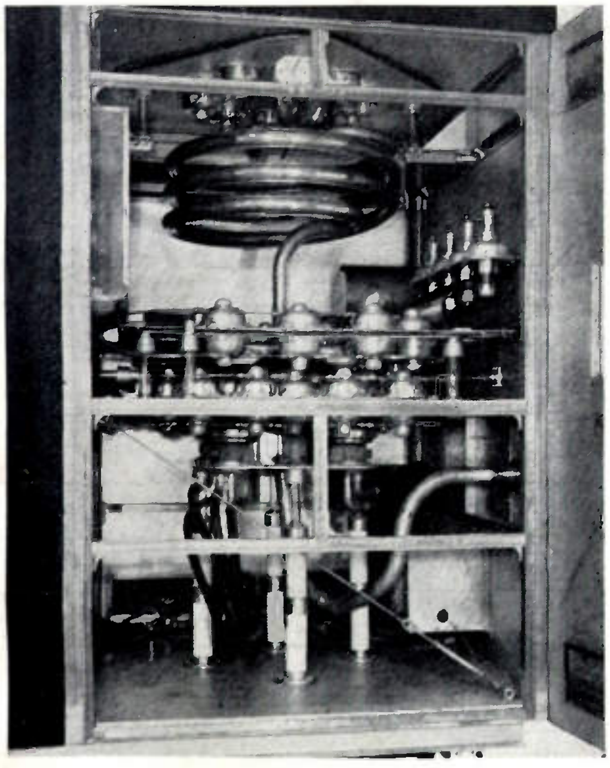
Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering
(Including the WAVES AND ELECTRONS Section)

April, 1949

Volume 37 Number 4



Collins Radio Company

THE UNIVERSITY OF ROCHESTER
CYCLOTRON OSCILLATOR

The principal frequency-determining element is the dee itself, which is inserted in the feedback line between anode and cathode circuits. This cyclotron, which produced mesons artificially on January 9, 1949, has pole pieces 130 inches in diameter and produces 270 Mev protons.

PROCEEDINGS OF THE I.R.E.

Waves and Electrons Section

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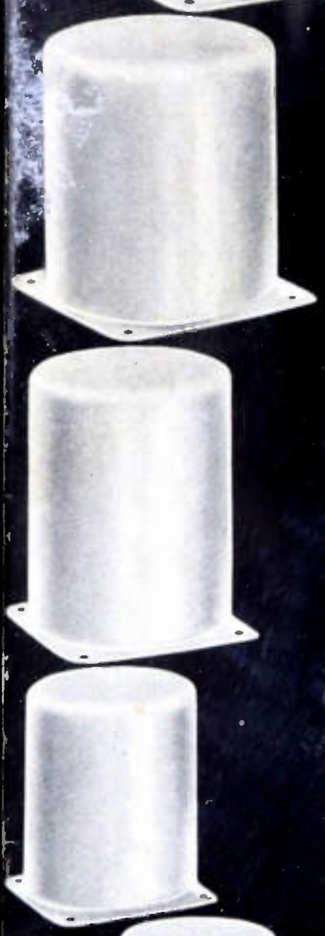
The Institute of Radio Engineers

COMMERCIAL GRADE COMPONENTS

A wide range of units for every application

U.T.C. Commercial Grade components employ rugged, drawn steel cases for units from 1" diameter to 300 VA rating . . . vertical mounting, permanent mold, aluminum castings for power components up to 15 KVA. Units are conservatively designed . . . vacuum impregnated . . . sealed with special sealing compound to insure dependability under continuous commercial service.

A few of the large number of standard C.G. units are described below. In addition to catalogued units, special C.G. units are supplied to customer's specifications.



CG VARIMATCH OUTPUTS FOR P. A.

Universal units designed to match any tubes within the rated output power, to line or voice coil. Output impedance 500, 200, 50, 16, 8, 5, 3, 1.5 ohms. Primary impedance 3000, 5000, 6000, 7000, 8000, 10,000, 14,000 ohms.

Type No.	Audio Watts	Typical Tubes	List Price
CVP-1	12	42, 43, 45, 47, 2A3, 6A6, 6F6, 25L6	$ 9.00
CVP-2	30	42, 45, 2A3, 6L6, 6V6, 6B5	14.00
CVP-3	60	46's, 50's, 300A's, 6L6's, 801, 807	20.00
CVP-4	125	800's, 801's, 807's, 4-6L6's, 845's	29.00
CVP-5	300	211, 242A's, 203A's, 838's, 4-845's, ZB-120's	50.00

CG VARIMATCH LINE TO VOICE COIL TRANSFORMERS

The UTC VARIMATCH line to voice coil transformers will match any voice coil or group of voice coils to a 500 ohm line. More than 50 voice coil combinations can be obtained, as follows:

.2, .4, .5, .62, 1, 1.25, 1.5, 2, 2.5, 3, 3.3, 3.8, 4, 4.5, 5, 5.5, 6, 6.25, 6.6, 7, 7.5, 8, 9, 10, 11, 12, 14, 15, 16, 18, 20, 25, 28, 30, 31, 40, 47, 50, 63, 69, 75.

Type No.	Audio Watts	Primary Impedance	Secondary Impedance	List Price
CVL-1	15	500 ohms	.2 to 75 ohms	$ 8.00
CVL-2	40	500 ohms	.2 to 75 ohms	11.50
CVL-3	75	500 ohms	.2 to 75 ohms	17.50

CG VARIMATCH MODULATION UNITS

Will match any modulator tubes to any RF load.

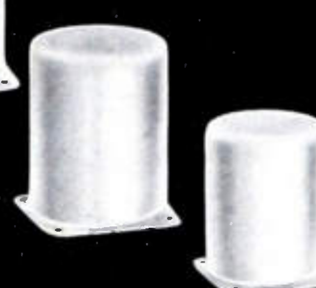
Primary impedances from 500 to 20,000 ohms
Secondary impedances from 30,000 to 300 ohms

Type No.	Max. Audio Watts	Max. Class C Input	Typical Modulator Tubes	List Price
CVM-0	12	25	30, 49, 79, 6A6, 53, 2A3, 6B5	$ 8.50
CVM-1	30	60	6V6, 6B5, 2A3, 42, 46, 6L6, 210	14.00
CVM-2	60	125	801, 6L6, 809, 4-46, T-20, 1608	20.50
CVM-3	125	250	800, 807, 845, TZ-20, RK-30, 35-T	30.00
CVM-4	300	600	50-T, 203A, 805, 838, T-55, ZB-120	50.00
CVM-5	600	1200	805, HF-300, 204A, HK-354, 250TH	115.00

INPUT, INTERSTAGE, MIXING AND LOW LEVEL OUTPUT TRANSFORMERS

(200 ohm windings are balanced and can be used for 250 ohms)

CG Type No.	Application	Primary Impedance Ohms	Secondary Impedance Ohms	List Price
131	1 plate to 1 grid	15,000	135,000 3:1 ratio	$ 9.50
132	1 plate to 2 grids	15,000	135,000 centertapped 3:1 ratio overall	10.00
133	2 plates to 2 grids	30,000 P to P	80,000 overall 1.6:1 ratio overall	12.50
134	Line to 1 grid hum-bucking	50, 200, 500	80,000	12.50
135	Line to 2 grids hum-bucking	50, 200, 500	120,000 overall	13.50
235	Line to 1 or 2 grids, hum-bucking; multiple alloy shielded for low hum pickup	50, 200, 500 ohms	80,000 overall	17.50
136	Single plate and low impedance mike or line to 1 or 2 grids Hum-bucking	15,000, 50, 200	80,000 overall	13.50
233	PP 6C5, 56, similar triodes to AB 45's, 2A3's, 6L6's, etc.	30,000 P to P	25,000 overall .9:1 ratio overall	11.00
333	PP 6C5, 56, similar triodes to fixed bias 6L6's	30,000 P to P	7,500 overall .5:1 ratio overall	11.00
433	PP 45, 2A3, similar tubes to fixed bias 2 or 4 6L6's	5,000 P to P	1,250 overall .5:1 ratio overall	12.00
137	Mixing	50, 200, 500	50, 200, 500	10.00
140	Triode plate to line	15,000	50, 200, 500	12.00
141	PP triode plates to line	15,000	50, 200, 500	13.50



United Transformer Co.

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Important Regional Meetings!

Cincinnati Spring Meeting

Third Annual Television Conference

April
23
1949

The Conference will be held at the Engineering Societies Building located at McMillan and Woodburn Sts., Cincinnati, Ohio.

Cincinnati
Section

The Third Annual Television Conference will be held at the Engineering Societies Building located at McMillan at Woodburn St., Cincinnati, Ohio, Saturday, April 23, 1949. The agenda is arranged to cover television horizons which have not widely been discussed, with emphasis being placed upon the UHF techniques as applied to monochrome reception in the 475-890 megacycle band.

Mr. Donald G. Fink member of the IRE Board of Directors and Editor of "Electronics" will present a review of factors which are especially important in the improvement of television receivers, most of which are available for immediate use.

Dr. Andrew V. Haeff, head of the Vacuum Tube Research Section at the Naval Research Laboratory will discuss fundamental ideas underlying the principle of operation of the three new types of UHF broad band amplifiers.

Papers to be Presented:

Propagation Characteristics of UHF R Radiation, Edward W. Allen, F.C.C. The Use of Stratovision in the UHF Band, C. E. Nobles, Westinghouse. UHF Television and Matching Techniques, O. M. Woodward, Jr., RCA Labs.

UHF Tuners for Receivers and Converter Use, Robert F. Romero, RCA Industry Service Laboratories. UHF Converter for Use with Present Receivers, Robert F. Wakeman, Allen B. DuMont Laboratories Inc. UHF Broad Band Amplifiers, Dr. A. V. Haeff, Naval Research Laboratory. The Influence of UHF Allocations on Receiver Design and Performance, John D. Reid, Crosley Div., Avco Mfg. Corp.

Psychophysiological Effects of Viewing Television, E. W. Commery, Nela Park, G. E. Co.

Trends in Television Receiver Design, Donald G. Fink, McGraw-Hill.

For registration or any questions concerning the conference, please contact Mr. Calvin Bopp, Spring Technical Conference Registration Chairman, Engineering Societies Building, McMillan at Woodburn St., Cincinnati, Ohio.

NEW ENGLAND

The 3rd Annual New England Radio Engineering Meeting sponsored by the New England division of the IRE will be held as before at the Hotel Continental, Cambridge, Mass. on May 21, 1949. Mr. P. K. McElroy of the General Radio Co. is chairman of the 1949 committee. Papers will be given in both morning and afternoon sessions. A luncheon and banquet will be held. Trips have been planned throughout the Boston area for those interested. Approximately 32 New England electronic firms will exhibit.

Calendar of COMING EVENTS

Annual Symposium, Engineers Council of Houston, Houston, Tex., April 2, 1949

Semiannual Convention, Society of Motion Picture Engineers, New York City, April 4-8

AIEE Conference on Electron Tubes, Buffalo, N.Y., April 11-12

AIEE Southwest District Meeting, Dallas, Tex., April 19-21

Third Annual Spring Conference, Cincinnati Section, IRE, Cincinnati, Ohio, April 23

IRE-RMA Spring Meeting, Philadelphia, Pa., April 25-27

IRE-URSI Spring Meeting, Washington, D.C., May 2-4

NAB Engineering Conference, Chicago, Ill., April 6-9

AIEE Summer General Meeting, Swampscott, Mass., June 20-24

AIEE Pacific General Meeting, San Francisco, Calif., August 23-26

1949 IRE West Coast Convention, San Francisco, Calif., August 29-September 1

AIEE Midwest General Meeting, Cincinnati, Ohio, October 17-21

1949 National Electronics Conference, Chicago, Ill., September 26-28

THE INSTITUTE OF RADIO ENGINEERS

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

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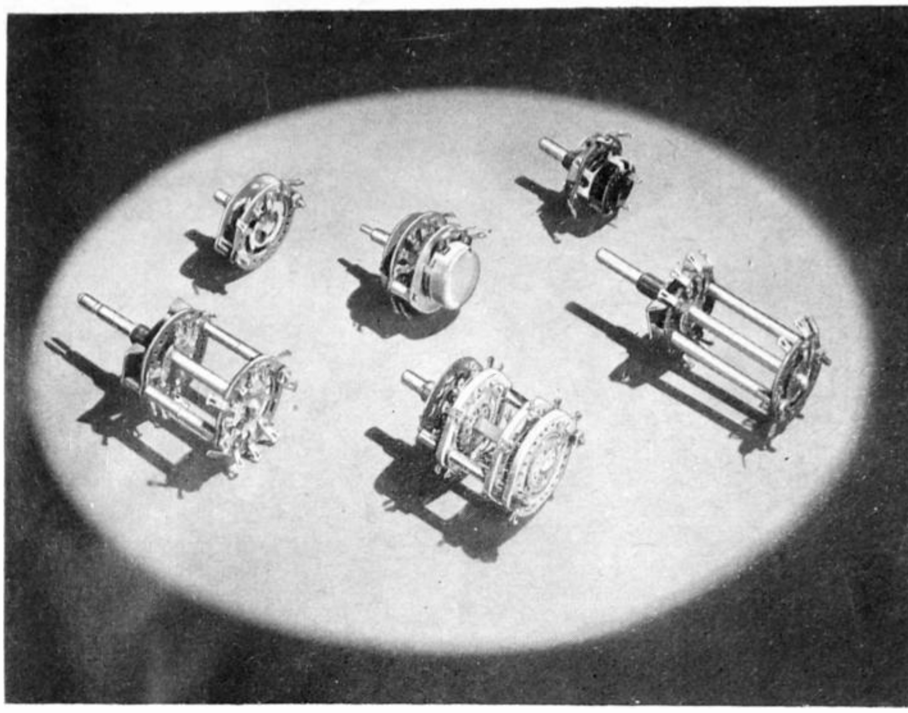

Use a Switch Worthy of Your Design

There is no substitute for MALLORY Quality Switches!

Mallory RS switches are designed to give you everything you want—maximum efficient service, substantial construction, precision manufacture. Mallory switches are constructed with cam and ball type index assembly, or with positive indexing hill-and-valley double roller type index assembly. Note these many features of the Mallory RS series which make their dependability and quality known wherever switches are used. These advantages are of extreme importance in television and high-frequency applications where stability is essential.

- Insulation of high-grade, low-loss laminated phenolic.
- Terminals and contacts of special Mallory spring alloy, heavily silver-plated to insure long life at low contact resistance.
- Terminals held securely by exclusive Mallory two-point fastening—heavy staples prevent loosening or twisting.
- Double wiping action on contacts with an inherent flexing feature—insures good electrical contact with the rotor shoes throughout rotation.
- Six rotor supports on the stator—insure accurate alignment.
- Brass rotor shoes, heavily silver-plated—insure low contact resistance.
- All shoes held flat and securely to phenolic rotor by rivets—prevents stubbing—insures smooth rotation—*minimum of noise in critical circuits.*

The Mallory RS series consists of RS-30, RS-40, RS-50, RSA-50, and RSA-60.

ENGINEERING DATA SHEETS

Send for the Mallory Engineering Data Sheets on the RS series. They contain complete specifications for available circuit combinations with respective terminal locations, dimensional drawings — everything the engineer needs.

SPECIFICATION SHEETS

Specification sheets for all RS switches have also been prepared. These sheets are printed on thin paper to permit blueprinting. The sectional drawings indicate standard and optional dimensions—make it easy for you to order production samples built to your requirements.

Precision Electronic Parts—Switches, Controls, Resistors

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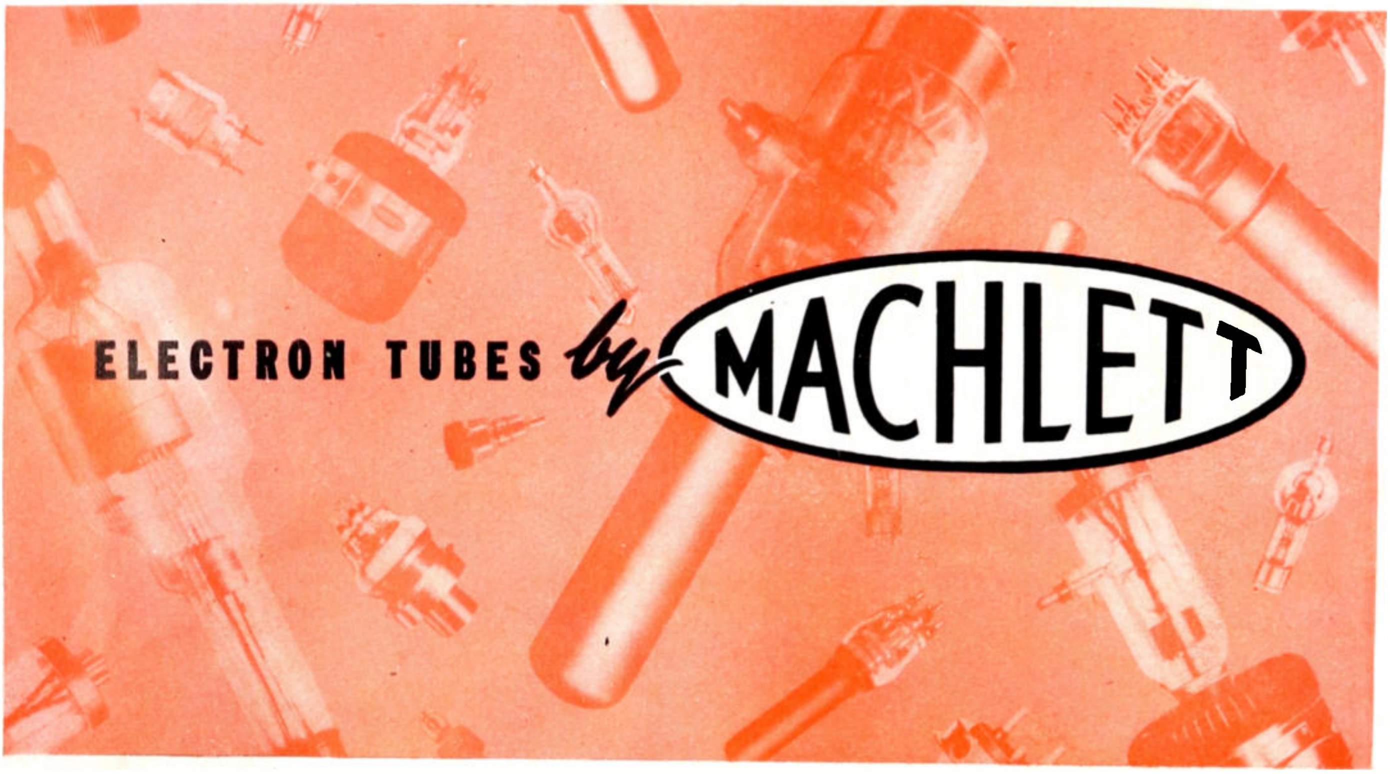

Combining Research—Development—Manufacturing Skill

Here's great news for broadcasters and industrial tube users. Machlett Laboratories and Graybar Electric Company have joined forces in a new distribution line-up to bring you more efficient and complete service on electron tubes.

For over a half century, Machlett has pioneered and made notable contributions to the development of the electron tube art. Today, through its modern plant, development laboratories and skilled personnel, Machlett tubes will set the highest standard of performance in broadcast and industrial service.

This combination of Machlett and Graybar is your best assurance of getting superior tubes. For better value—better service—try Machlett tubes now distributed via Graybar.

MACHLETT — OVER 50 YEARS OF ELECTRON TUBE EXPERIENCE

Famous as the outstanding manufacturer of X-ray tubes, the name, Machlett, on electron tubes has been the mark of quality, top performance and long life for more than 50 years. Experience, skill and a "singleness of purpose" to produce the best in electron tubes have made Machlett first choice around the world.

MACHLETT LABORATORIES, INC., Springdale, Connecticut

SERVE BROADCASTERS AND INDUSTRY

with National Distribution Service

In keeping with its policy of "Bringing You Broadcasting's Best Equipment," Graybar is proud to assign its Tag—the Symbol of Distribution—to the Machlett line of electron tubes for both broadcasters and industry.

This new connection will bring you dual benefits: (1) products from an outstanding manufacturer of electron tubes, (2) distribution service from an organization offering specialized assistance in choosing the best type of product for your requirements.

Machlett tubes can now be quickly and conveniently ordered through near-by Graybar "Supply Stations" located in over 100 principal cities from coast-to-coast. When you order Machlett tubes "via Graybar," you'll have the right combination for extra service and performance.

Call your local Graybar Representative. Graybar Electric Company, Inc., Executive Offices: Graybar Building, New York 17, N. Y.

EVERYTHING ELECTRICAL TO KEEP YOU ON THE AIR — *via* Graybar

These are the Graybar Broadcast Equipment and Electron Tube Specialists in key cities:

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CLEVELAND
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DALLAS
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P. L. Gundy, Temple 1-5500

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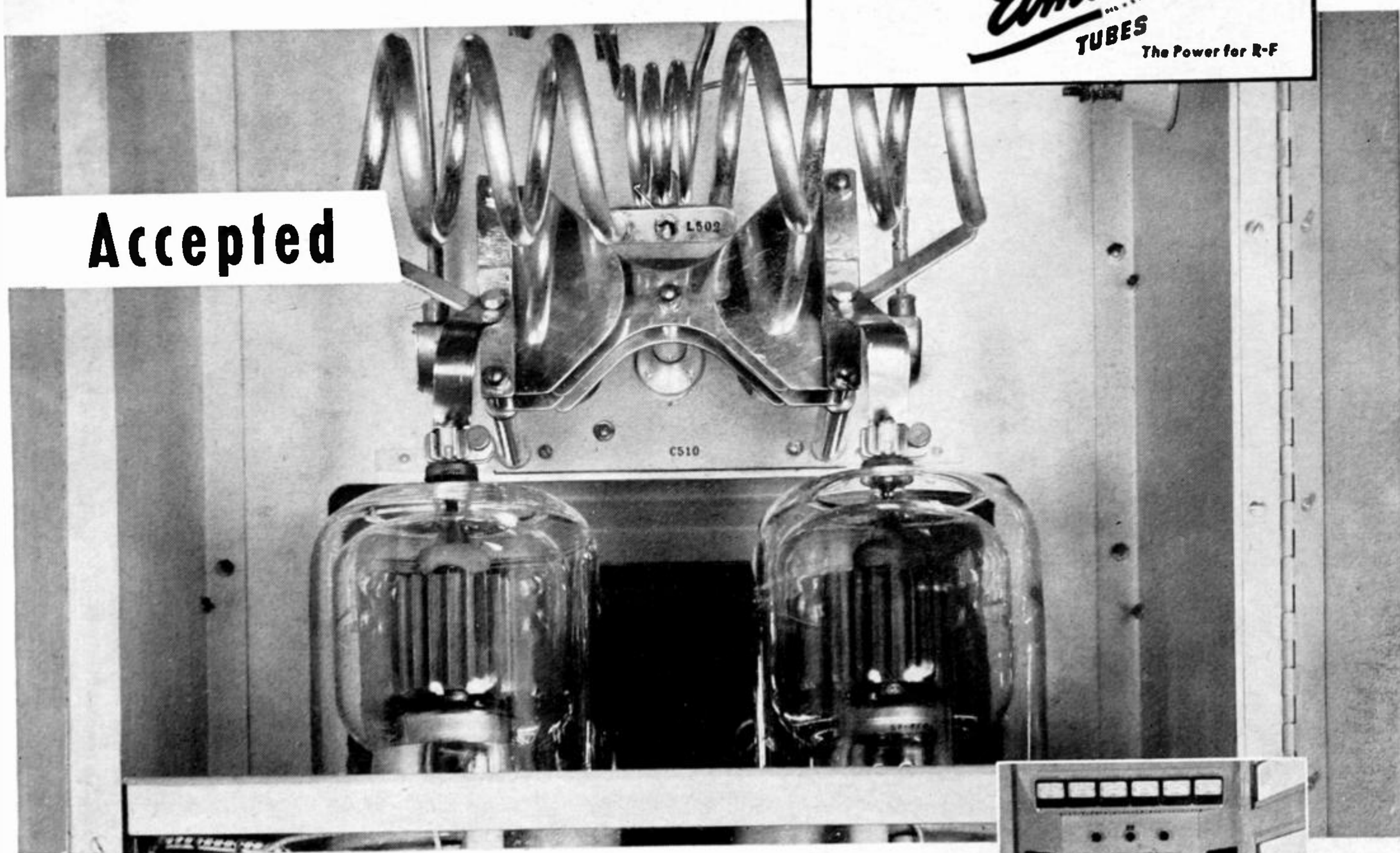
Eimac TUBES

The Power for R-F



Accepted

LINK 3000 UFS

For Emergency Services . . . .
The Link 3000UFS Transmitter
and Eimac 4-1000A Tetrodes

Here's a team that fills the bill by providing the dependability of performance required by police and other emergency communication services.

Link Radio, well known manufacturers of radio communication equipment, in designing their 3 kw 30-50 Mc. FM transmitter choose Eimac 4-1000A tetrodes to power the final amplifier. The high power-gain of these tubes enabled Link to use their standard 50 watt transmitter as a driver. The resulting compact simplified transmitter is ideally suited for control through telephone circuits from remote locations. A single pair of telephone lines carries transmitter modulation, power control, overload relay reset, and frequency selection plus receiver output and selection.

Because of their power-gain abilities, stability and other exceptional characteristics, the 4-1000A tetrode offers the design engineer interesting potentialities . . . write direct for further information, data is available.

EIMAC 4-1000A TETRODE

EITEL-McCULLOUGH, INC.

212 San Mateo Ave., San Bruno, California

Export Agents: Frazar & Hansen, 301 Clay St., San Francisco, California

Heat dissipation can be

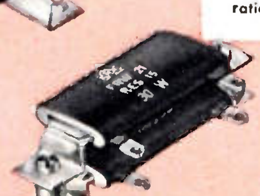
for resistors

Heat dissipation can be mighty tough . . . but not for IRC resistors. They are universally engineered for the lowest possible operating temperatures and maximum power dissipation within the smallest size units consistent with good engineering practice.

Long experience with the widest line of resistor types in the industry has provided IRC with a wealth of "know-how" on resistor heat dissipation. In Power Wire Wound Resistors for example, the complete range of tubular and flat types manufactured by IRC utilizes a special cement coating to attain rapid heat dissipation. This dark rough surface does double duty by effectively guarding the windings against harmful atmospheric moisture and corrosion. Use the handy coupon to get complete data on proven advantages of IRC Power Wire Wounds.

There are 52 specific types of IRC Power Wire Wound Resistors.

IRC Flat Wire Wound Resistors offer a higher space-power ratio.



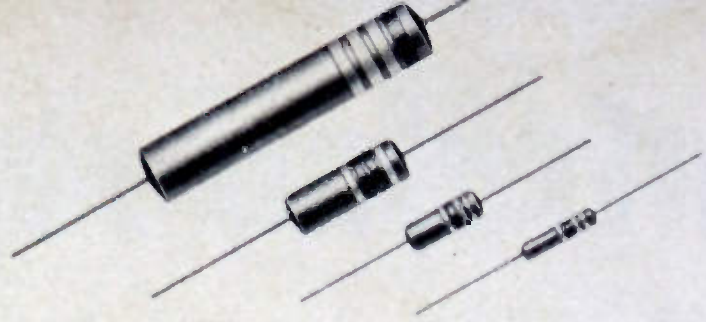
Adjustable types feature the exclusive IRC non-corrosive contact band.

Power ratings from 2 to 225 watts. Variety of 8 terminal types.

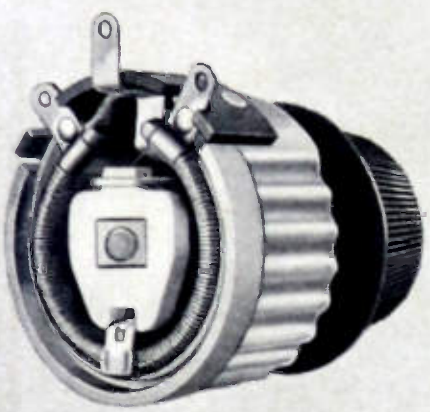
Windings are of highest grade alloy wire on tough ceramic forms.

tough

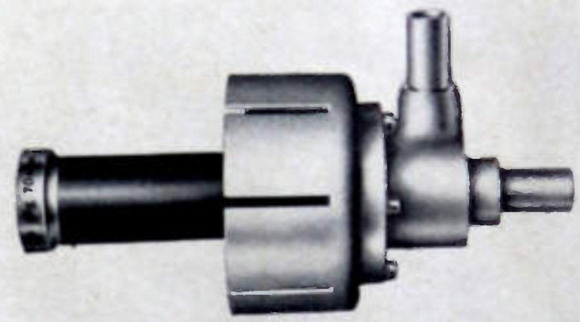
too!



New, ADVANCED BT Resistors obsolete present performance standards for fixed composition resistors. Extremely low operating temperature and excellent power dissipation in compact, light weight, fully insulated units at ⅓, ½, 1 and 2 watts. These ADVANCED resistors meet JAN-R-11 specifications. All the facts are included in 12-page technical data Bulletin B-1.



Heat dissipation properties of aluminum are used to full advantage in housing and winding core of IRC Power Rheostats, 25 and 50 watts. Type PR Rheostats operate at full rating at about half temperature rise of equivalent units. Can be operated at full power in as low as 25% of rotation without appreciable difference in temperature rise. Direct contact between rheostat and mounting panel allows rapid conduction to panel of a portion of heat dissipated. Send for Bulletin E-2.



Water-cooled LP Resistors utilize high velocity water stream flowing in spiral path against thin resistance film. High power dissipation is made possible by centrifugal force holding water in thermal contact with resistance surface. Resistance film less than 0.001'' thick with active length much less than ¼ wave length at FM and television frequencies, gives excellent frequency characteristics. Resistance values 35 to 1500 ohms; 15% tolerance standard; power dissipation up to 5 K.W. ac. Bulletin F-2 gives all the facts.

If you have the heat put to you for speedy service on small order resistor requirements for experimental work, pilot runs, etc., you'll appreciate the advantages of IRC's Industrial Service Plan. This enables you to get 'round-the-corner service from the local stocks of your IRC Distributor. He's a good man to know . . . we'll gladly send you his name and address.

IRC

Wherever the Circuit Says —ᴧᴧᴧ—

Power Resistors • Voltage Dividers • Insulated Composition Resistors • Low Wattage Wire Wounds • Controls • Deposited Carbon Precistors • Precisions • HF and High Voltage Resistors • Voltmeter Multipliers • Rheostats

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INTERNATIONAL RESISTANCE COMPANY
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Send me additional data on items checked below:

- ☐ Power Wire Wounds (tubular)
- ☐ Flat Power Wire Wounds
- ☐ Advanced BT Resistors
- ☐ Power Rheostats
- ☐ Water-Cooled Resistors
- ☐ Name and address of our local IRC Distributor

NAME...

TITLE...

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....

Centralab reports to

1 Simplified wiring and assembly . . . fewer individual components . . . fewer leads to be soldered — these are some of important production-boosting advantages you get with CRL *Printed Electronic Circuits*. In addition, P. E. C. — by combining up to 45 individual parts into one light, tiny unit — makes it possible to reduce the weight and size of the electronic products you manufacture. For complete P. E. C. information, see your Centralab Representative, or write direct.

Reaching Out in Ohio

WCLT at Newark

● For a full 78 hours every week, WCLT energizes central Ohio air with 8500 watts of effective radiated power—transmitted through a 4-bay General Electric FM antenna mounted atop a 290-foot Truscon Self-Supporting Steel Radio Tower. Total antenna height reaches 332 feet above ground level.

Truscon Radio Towers are on horizons everywhere, serving the needs of AM, FM and TV broadcasters. Strong and stable, these slender steel structures are

engineered for top operating efficiency under specific local conditions.

Truscon engineers are ready now to put their vast experience to work for you. Truscon can furnish exactly the tower you need—guyed or self-supporting, uniform or tapered in cross-section, of any height. A call or letter to our home office in Youngstown, Ohio, —or to any convenient Truscon District office—rates immediate attention, and action—with no obligation.

TRUSCON STEEL COMPANY
YOUNGSTOWN 1, OHIO
Subsidiary of Republic Steel Corporation

TRUSCON
SELF-SUPPORTING
AND UNIFORM
CROSS SECTION GUYED
TOWERS

Another
TRUSCON
TOWER OF STRENGTH
290 FT.
HIGH

AVIATION ASKED FOR THEM ...RADIO AND TV BENEFIT!

GENERAL ELECTRIC CUSTOM MINIATURES
Made and tested for supreme reliability!

"MORE dependable than any miniatures yet built." That was aviation's directive . . . and challenge! Thousands of premium-performance GL-5654's and GL-5670's now in use, prove how well the challenge has been met. In altimeters, radio compasses, radio control equipment, and high-frequency aircraft radio receivers, these fine General Electric tubes are doing the extra-reliable job for which they were painstakingly made.

You, as designer or user of radio-TV transmitter equipment, can have the protection of G-E custom-miniature dependability *now*—starting with Type GL-5654 (electrically the same as the 6AK5), and Type GL-5670 (similar to the 2C51 except for improved heater design and a somewhat higher heater current). Other types are being added.

These tubes are carefully manufactured one by one, from individually gaged and inspected heaters, cathodes, grids, and plates. Each gets not less than 50 hours' operation—ample assurance that when plugged in, tube performance will be in line with ratings consistently. Ask your G-E electronics office for further facts. Or write *Electronics Department, General Electric Company, Schenectady 5, New York.*

GL-5654
7-pin miniature h-f pentode

FIRST OF AN *Outstanding* NEW SERIES

GL-5670
9-pin miniature h-f twin triode

Characteristics

TYPE GL-5654

Heater voltage, a-c or d-c	6.3 v
Heater current	0.175 amp
Max ratings, design center values:	
plate voltage	180 v
Grid No. 2 voltage	140 v
plate dissipation	1.7 w
Grid No. 2 dissipation	0.5 w
Typical operation:	
plate voltage	180 v
Grid No. 2 voltage	120 v
cathode-bias resistor*	200 ohms
plate resistance (approx)	0.69 megohms
transconductance	5,100 micromhos
plate current	7.7 ma
Grid No. 2 current	2.4 ma

(*Fixed-bias operation not recommended)

TYPE GL-5670

Heater voltage, a-c or d-c	6.3 v
Heater current	0.350 amp
Max ratings, design center values, each triode section:	
plate voltage	300 v
plate dissipation	1.5 w
Typical operation, Class A1:	
plate voltage	150 v
cathode resistor, per section	240 ohms
plate current, per section	8.2 ma
transconductance, per section	5,550 micromhos
amplification factor	35
cut-off grid voltage, Ib equals 75 mu a (approx)	—10 v
Typical operation, Class AB1:	
plate voltage	300 v
cathode resistor	800 ohms
A-F grid-to-grid voltage, RMS	14 v
zero-signal plate current, per section	4.9 ma
max-signal plate current, per section	6.3 ma
load impedance, plate-to-plate	27,000 ohms
total harmonic distortion	10 per cent
max-signal power output	1.0 w

GENERAL ELECTRIC

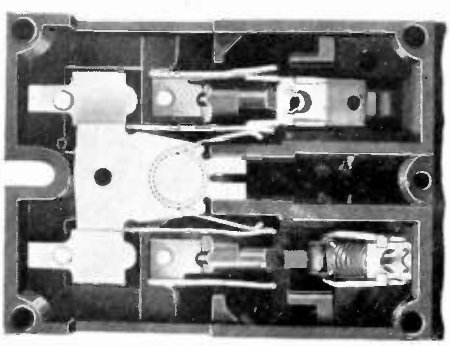
180-H3

FIRST AND GREATEST NAME IN ELECTRONICS

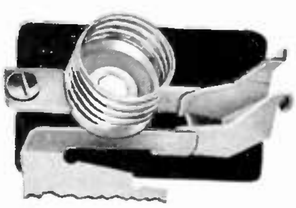

ALSIMAG
CUSTOM MADE
TECHNICAL CERAMICS
AMERICAN LAVA CORPORATION
Main Plant and Office
Production experience and
continuous research for
almost half a century have made
this building the headquarters
for problems on
Technical Ceramics

REVERE "KNOW HOW" made premium metal unnecessary!

Federal Noark NTPS Panelboard, made by the Federal Electric Products Company, Newark 5, N. J.



Showing the sure-contact spring in switch assembly.



These are the three current-carrying parts.

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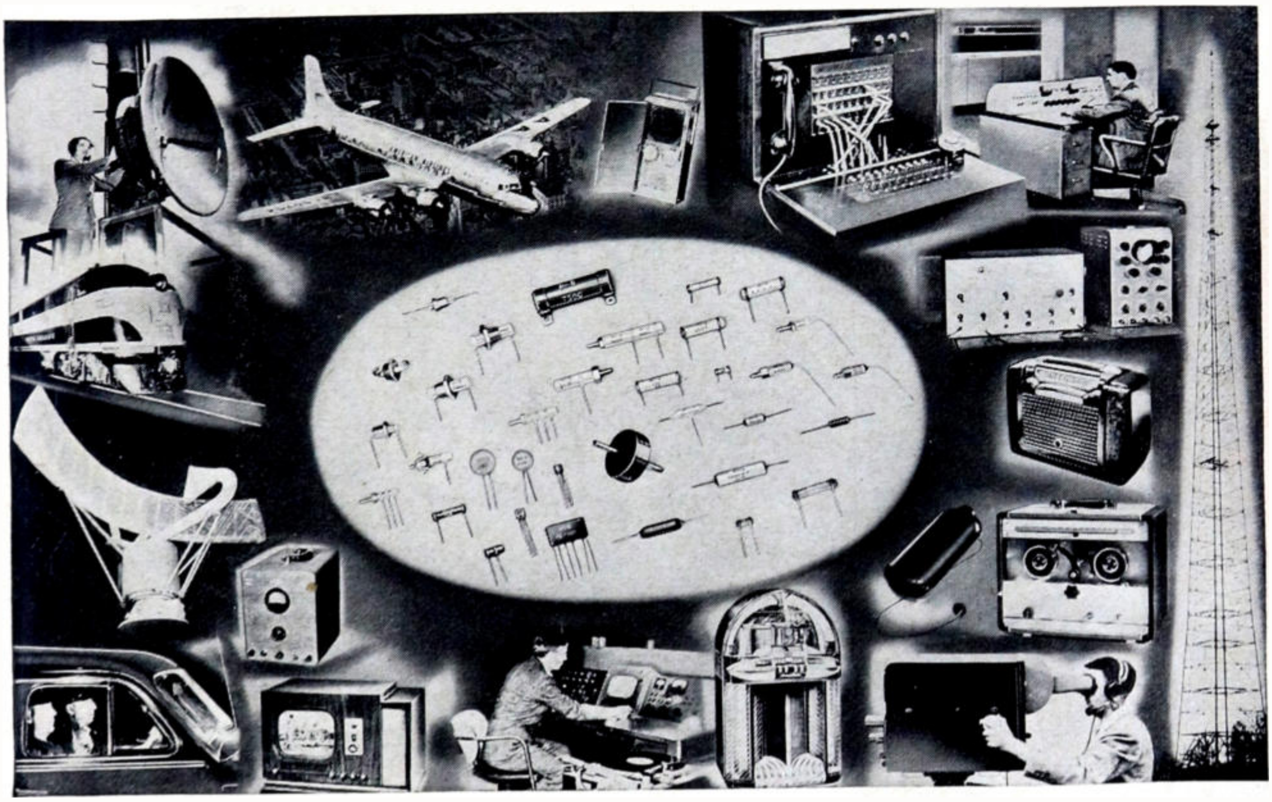

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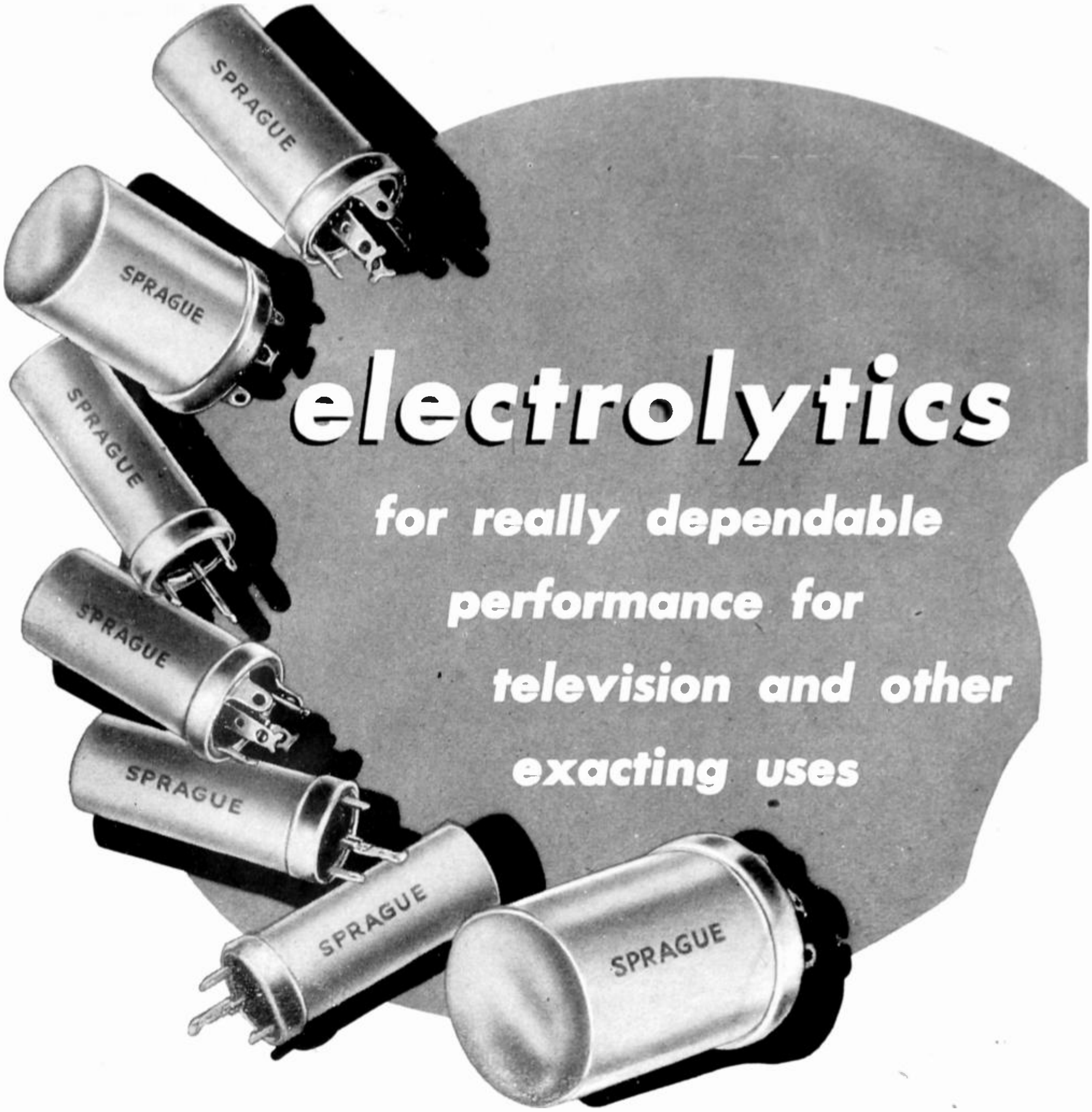

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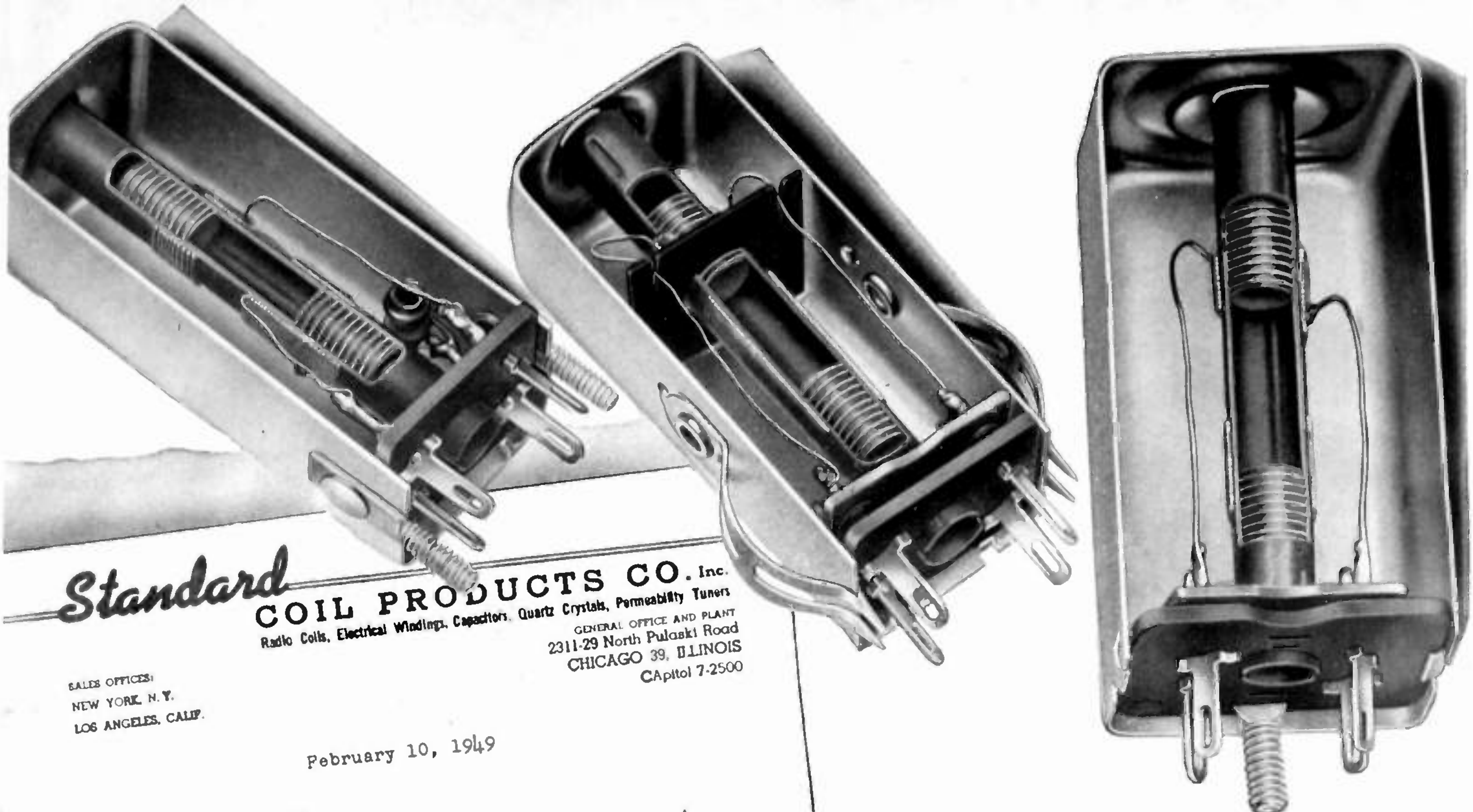

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February 10, 1949

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Dear Tom:

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Now . . low cost "E" powder for high frequencies!

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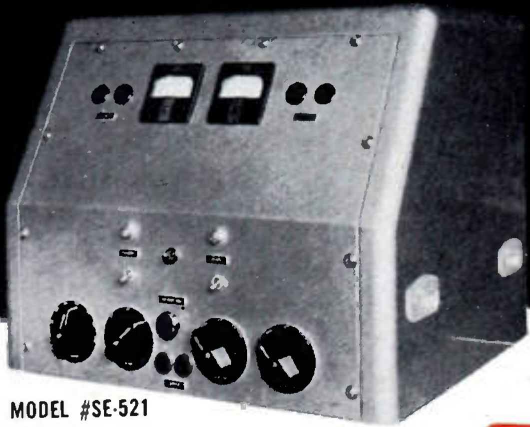

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This unit is designed for D.C. Amplifier study, Nuclear Physics Instrumentation, and all laboratory projects requiring power supplies of excellent characteristics.

A very high gain feed-back loop in the regulator accounts for its exceptional qualities. Its versatility makes it the standard laboratory power supply. Available in a sloping panel cubicle for bench use, or on a "tea-wagon", or attached to a standard rack-panel for incorporation in permanent equipment.

CIRCUIT

110 volts input is passed through a regulating transformer and is transformed, rectified and filtered to appear as approximately 800 volts D.C. This D.C. powers a regulator circuit designed to offer 300 to 600 volts D.C. regulated output. Passing tubes are 6-1614's. Three stage direct coupled amplifier produces the exceptional regulation features down to very low voltages.

SPECIFICATIONS

Range: 0-200 ma. 0-1200 volts, continuous.

Regulation: For line voltage variations from 95 to 130 volts, output will remain constant within .2% or .1 volts on lower ranges. For load current variations from 0 to 200 ma. output will remain constant within .5% or .25 volts on the lower ranges.

Ripple: .01% or 5 mv depending upon range.

Tube Complement: 12-1614; 2-6AC7; 2-6SL7; 2-VR150; 2-5R4GY.

Power Requirements: 115 Volts, 60 cps, single phase 500 watts.

CUSTOM ADDITIONS

Basic design can be modified to meet any special requirements. If large voltage range is unnecessary, supplies with lesser ranges can be furnished. (0-600 volts, 300 to 600 volts, etc.) Basic sections (regulator unit, rectifier unit, control unit) can be purchased separately.

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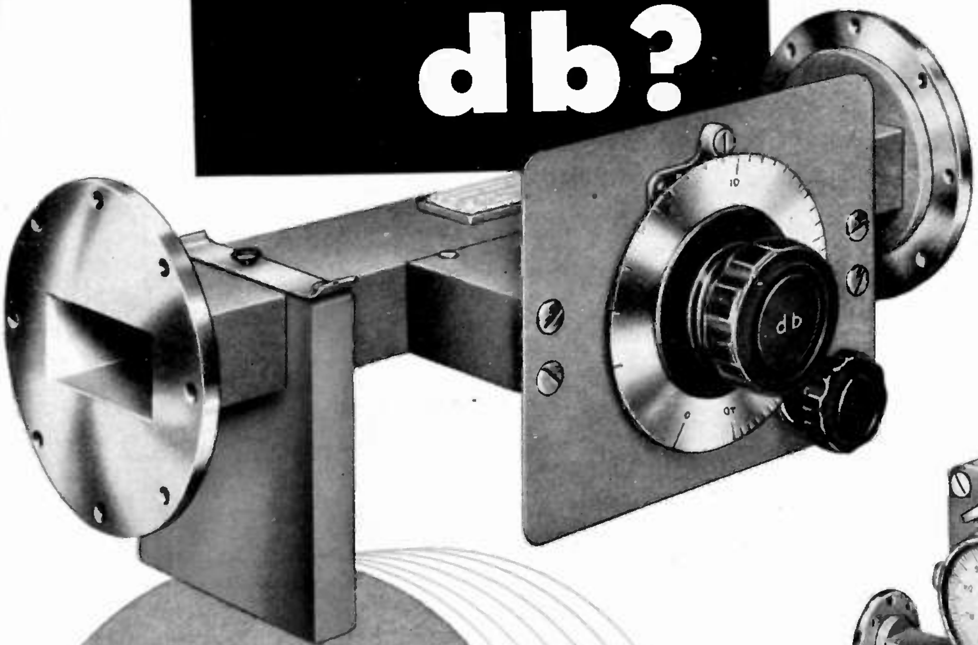
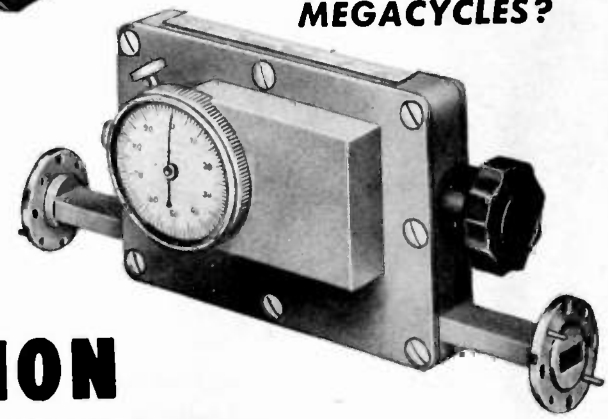


..AT 5,000 MEGACYCLES?

..AT 25,000 MEGACYCLES?



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Features:

- METALLIZED-GLASS ATTENUATING ELEMENTS
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- BACKLASH-FREE VERNIER DRIVE
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These attenuators, utilizing metallic-film-on-glass resistive elements, are of the most advanced design now available. Broadband matching, minimum frequency sensitivity, and constancy of attenuation with time, are but a few of the advantages which this technique affords. Casing designs available cover requirements from that of a simple power control attenuator to that of a precisely calibrated secondary standard.

Metallized glass attenuators are an important type of microwave measurement component in the complete PRD line. Available also are precision slotted sections and probes, impedance matching devices, frequency meters and standard cavities, and all of the other items which make up a complete measurements bench. An illustrated catalog and price list may be obtained by writing to Department R-5, on company letterhead.

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NEWS—NEW PRODUCTS

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

SU-4 Radioactivity Demonstrator

A new radioactivity demonstrator, type SU-4, has been developed in response to the need for an inexpensive instrument for teaching basic principles of radioactivity, by **Tracerlab Inc.,** 55 Oliver St., Boston 10, Mass. Radiation is indicated by three different means: by loudspeaker, by a flashing neon light, and quantitatively with a counting-rate meter. This instrument has sufficient accuracy to permit carrying out simple class-room and laboratory demonstrations of basic principles of radioactivity.

Included with the radioactivity demonstrator is a glass Geiger-Mueller tube capable of detecting gamma and high-energy beta rays. It is mounted inside the chassis to eliminate the possible danger of exposed high-voltage leads. A removable shield is mounted in front of the tube to completely eliminate beta rays when it is desired to measure only gamma radiation.

The meter reads in counts per minute and will indicate up to 2,500 counts per minute. A switch permits the meter to also indicate the voltage across the tube, which can be varied from about 500 to 1,100 volts. The instrument operates on 110 volts ac.

For information regarding the latest commercially available radioisotopes, request "Tracerlog" from this company.

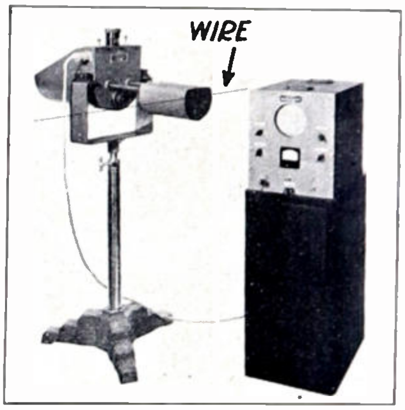
New Location

The Vickers Electric Division, Vickers, Inc., has recently moved into new quarters at 1815 Locust St., St. Louis 3, Mo., to facilitate the expansion of their line of products to the industry.

The firm will now be able to engage more efficiently in research, development and production of magnetic audio amplifiers, static voltage regulators, static motor-speed controls, power saturable reactors, rectifiers, photoelectric cells, servomechanisms, magnetic fluid clutches, special motors and generators, transformers, and controlled power rectifiers for electrochemical processes.

Visi-Limit Micrometer

The new type W1, an electronic control unit for production measurement of a dimension, designed for factory use where high speed and laboratory accuracy are required, is available from **Raymond M. Wilmotte, Inc.,** 1469 Church St., N.W., Washington 5, D. C.



The type W1 measures the outside diameter of wire, thread, tubing, or rod, and edge-to-edge dimensions of extruded parts, strip stock, etc.

This unit may be modified to measure thickness of sheet, the dimensions of certain machined or punched parts, or routine production test of several dimensions simultaneously.

Principal characteristics of the micrometer are: no contact with the object; no limit of speed of wire or strip; object may vibrate up to 0.4 inch; accuracy up to ±0.0001 inch; accuracy not dependent on electronic circuits, line voltages, etc.; two indicating instruments for instantaneous and average values; recorder provided when desired; warning lights, recorder, remote indicator; and power for automatic control up to 5 kw.

The Visi-Limit micrometer contains two component parts, a measuring head, and an indicating and power-supply unit. The measuring head houses the measuring aperture in front of which the material to be measured is fed. This opening, called the test aperture, is illuminated by a projection lamp. The object being measured casts a shadow, thereby reducing the illumination of the cathode of a photocell located behind the aperture.

The other two openings, described as limit apertures, are located close to the measuring aperture; these are separately adjustable, and are set to represent the positive and negative tolerance limits to which the material is to conform. All of these windows are relatively large.

A scanning disk rotates between the apertures and the photocell, exposing each of the three apertures in sequence. As a result of this rotation three voltage pedestals are fed to the amplifier circuit, each following the other in a time sequence synchronized with the disk rotation. Each

(Continued on page 55A)

WILCOX... first choice of MID-CONTINENT AIRLINES

MCA equips entire fleet with WILCOX 70 channel VHF transmitters and receivers

Mr. Daryl Devault, superintendent of communications for MCA, says:

"Wilcox VHF has brought closer the airlines' ultimate goal of all-weather flying — and in doing so it has proved an essential aid to Mid-Continent Airlines in maintaining a perfect safety record dating back to 1934 — and operating efficiency, which in 1948 reached a mark of 98.73 per cent.

"Mid-Continent pilots hail the equipment for the static-free, 'telephone clear' reception it assures them in plane-to-ground communications in all kinds of weather."

For many years Mid-Continent has used Wilcox ground station transmitters and receivers exclusively. Their proven performance, dependability, and easy maintenance earned the confidence of operation personnel throughout the system. That's why Wilcox was first choice again for MCA's airborne communications equipment.

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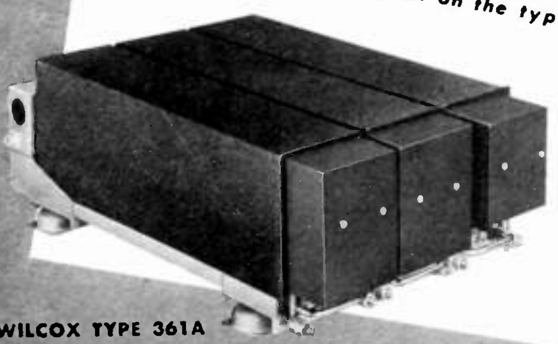

WILCOX TYPE 361A
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PROCEEDINGS OF THE I.R.E.

(Including the WAVES AND ELECTRONS Section)

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 37 — *April, 1949* — NUMBER 4

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

WAVES AND ELECTRONS SECTION

Donald B. Sinclair

TREASURER, 1949

Donald B. Sinclair was born at Winnipeg, Manitoba, Canada, on May 23, 1910. From 1926 to 1929 he was a radio amateur in Winnipeg under the call VE4FV, and during the last year he also worked as a part-time radio operator for the Western Canada Airways. At the same time he attended the University of Manitoba from 1926 to 1929, transferring in the latter year to a co-operative course in electrical engineering at the Massachusetts Institute of Technology. He received the S.B. degree from that institution in 1931, the S.M. in 1932, and the Sc.D. in 1935.

In 1930 Dr. Sinclair became a Junior Member of the IRE, and during that year and the next was associated with the Bell Telephone Laboratories and the New York Telephone Co. in New York, N. Y., and with the Western Electric Co. in Hawthorne, N. J. He was elected an IRE Associate in 1933. As a research assistant at MIT from 1932 to 1935 and research associate from 1935 to 1936, he worked on the electrocardiograph, electromechanical computing machines, and high-frequency measurements. Under a co-operative arrangement with the General Radio Co., his doctorate research work on high-frequency impedance measurements was sponsored by that organization and worked out in its laboratories during 1934 and 1935.

In 1936 Dr. Sinclair joined the staff of the General Radio Co. as an engineer, and was appointed chief engineer in 1944. There he worked chiefly on the general development and design of measuring instruments, with emphasis on high-frequency measuring equipment. Becoming a Member of the Institute in 1938 and a Senior Member in 1943, he was awarded the Fellow Grade in 1944.

Dr. Sinclair has been a member of the Board of Directors of the Institute since 1945, and is active on a number of committees.

The membership of The Institute of Radio Engineers in general, and of the IRE Audio Engineering Group in particular, will be interested in the following guest editorial dealing with the work of the IRE Electroacoustics Committee in standardization, carried out jointly with the American Standards Association and the Acoustical Society of America.

It is indicative of the considerable interest of the IRE in acoustical matters affecting the overall characteristics or performance of communications or electronic systems.

The attention of the IRE membership is particularly directed to the final paragraph of this editorial.—*The Editor*

New Standards on Electroacoustics

Definitions of Electroacoustic Terms Issued for Trial and Study

EGINHARD DIETZE

The Electroacoustics Committee, since the war, has been active in revising the standards on electroacoustics issued by The Institute of Radio Engineers in 1938. The initial effort was concerned with bringing the definitions up to date. A new *Standard on Electroacoustics* containing *Definitions of Electroacoustic Terms* has now been issued for trial and study, and is hereby brought to the attention of all IRE members in order that they may consider carefully the proposed definitions contained therein.

Because of the wide interest in the subject of Electroacoustics, it has been considered unwise by the IRE Technical Committee to carry on this work independently. The members of the Electroacoustics Committee therefore worked jointly with the ASA Subcommittee on Acoustical Terminology Z24A, which is sponsored by The Acoustical Society of America, and the proposed terminology is the combined product of these two committees who effectively merged for the purpose of carrying on this work. In addition, other committees of the ASA as well as technical committees of other interested societies were consulted, so that a wide cross section of all technical knowledge and experience in the field of acoustics is represented in this publication.

The need for a new terminology can be illustrated by considering that about five times as many terms are defined in this version as were included in the 1938 standards. Many new sections have been added reflecting recent scientific developments and activity during the late war, such as: Ultrasonics; Recording and Reproducing; Underwater Sound; Shock and Vibration; and MKS Units.

This terminology should be of value to workers in both theoretical and applied acoustics, manufacturers of radio, television, motion-picture and sound-recording and reproducing equipments; to the armed services; to research groups; and to medical groups investigating speech and hearing.

The proposed terminology is issued for trial and study for a six months' period, during which time comments are solicited. Any comments from IRE members should be sent directly to E. Dietze, Chairman, Electroacoustics Committee, c/o Bell Telephone Laboratories, Murray Hill, N. J.

Doppler Radar*

EDWARD J. BARLOW†, SENIOR MEMBER, IRE

Summary—**This paper contains a discussion of the principle of operation of cw doppler search radar systems and an analysis of their performance capabilities, with particular emphasis on the elimination of fixed targets. A comparison of these systems and MTI pulse radar systems is made.**

Introduction

Many kinds of radar sets have been developed for widely different applications during the past few years. Each application usually stresses some particular type of target. For example, a marine navigational radar system is built to show land masses, islands, buoys, and inlets as clearly as possible. On the other hand, an aircraft-warning radar should show aircraft or formations of aircraft clearly, but not land masses. Again, a small aircraft-interception radar system in a fighter plane should show adjacent aircraft but not clouds. From these few examples, it is clear that there are many applications for which response to some targets is desired, but response to others is objectionable.

One way in which targets differ is in that some are stationary and some moving. If a target is moving toward the radar system, the frequency of its radar echo is slightly higher than the frequency of the radar system itself; and if it is moving away, the frequency is slightly lower. This change in frequency with target motion is the doppler effect. Radar systems have been built which use this effect to distinguish between fixed and moving targets. They may be called doppler radar systems.

An ordinary pulse radar system may make use of this doppler principle by employing suitable modifications and additional circuitry. The system is then said to have MTI (moving-target indication) features. Systems of this type are discussed in the literature.[1]

An alternative approach is to abandon pulse-ranging techniques and make a cw (continuous-wave) radar system (or at least one with a duty cycle of the order of 1/2 to 1/10). Range information must then be determined by some other method or dispensed with altogether. This approach has been employed in some radar system development in the hope that greater discrimination between fixed and moving targets could be obtained together with greater flexibility in choosing relative emphasis of various target velocities. A limited amount of information is available on systems of this type.[2] These systems may be further subdivided into (1) systems which use frequency modulation of the transmitted signal to measure range,[3] (2) systems which use amplitude modulation, and (3) unmodulated cw systems[4] which do not measure range at all. It is the purpose of this paper to discuss cw systems of the latter two types in a sequence more or less paralleling the historical development of such systems, to discuss one particular amplitude-modulated range-measuring system with a duty cycle of 1/2, and then to compare the performance of this system with that of a typical pulse MTI system.

Before proceeding with this development, one or two general remarks might be made about the comparison of pulse and cw systems, not with respect to their relative ability to distinguish between fixed and moving targets, but rather with respect to their range capabilities and information rates. First, pulse and cw systems are, in principle, capable of approximately the same range performance for the same average power.[5] Practical considerations, however, sometimes prevent realization of this equivalence. Secondly, the number of independent elements of range information obtained by a radar system is directly proportional to the transmitted spectrum width, so that short-pulse systems and wide-band FM systems have a correspondingly large number of range elements[6] while an unmodulated cw system yields no range information.

The systems to be discussed in this paper all are either cw or have square-wave amplitude modulation so that the transmitted spectrum is narrow and the systems have, at most, one element of range information per beam position. However, their range performance may be expected to be roughly equal to that of an equivalent pulse system with the same average power. The equivalence may not be exact because of difference in the indicating device (for example, earphones instead of PPI), and because in some cases, the doppler-system bandwidths are not chosen to optimize system range performance but for some other reason, such as accommodating a certain band of doppler frequencies.

These general remarks will be replaced later by specific statements of the range performance and information rates of various doppler systems.

Doppler Principles

Many of the fundamental principles of doppler radar can be illustrated by the aid of a very simple system like that shown in Fig. 1. The transmitter output is an unmodulated carrier which is fed into the directive

* Decimal classification: R537. Original manuscript received by the Institute, August 25, 1947; revised manuscript received, September 10, 1948. Presented, 1948 IRE National Convention, New York, N. Y., March 22, 1948.

† Sperry Gyroscope Co., Great Neck, L. I., N. Y.

[1] L. N. Ridenour, "Radar Systems Engineering," Radiation Laboratory Series, vol. 1, McGraw-Hill Book Co., Inc., New York, N. Y., chap. 16.

[2] See chap. 5 of footnote reference 1.

[3] I. Wolff and D. G. C. Luck, "Principles of frequency-modulated radar," *RCA Rev.*, vol. 9, pp. 50–76; March, 1948.

[4] Two or more unmodulated cw systems could be used to measure range by a triangulation method.

[5] See pp. 123 and 124 of footnote reference 1.

[6] See p. 66 of footnote reference 3.

transmitting antenna. The receiver consists of a directive antenna followed by a detector (for example, a crystal rectifier) and an audio amplifier. The two antennas are mounted side by side and are arranged to rotate together. With such a system, some of the transmitter energy spills over from the transmitting antenna into the receiver, so that a spill-over or leakage signal is always present in the receiver. If there are objects present in the transmitter beam, energy reflected from them will also reach the receiver. This condition is illustrated in Fig. 2. Here are shown three types of signal entering the receiver, the leakage signal, a fixed-target signal (as that from a hill or building), and a moving-target signal (airplane, automobile, or bullet). The fixed-target signal will be of the same frequency as the leakage signal but the moving-target signal will, in general, be of a slightly different frequency because of the doppler effect.

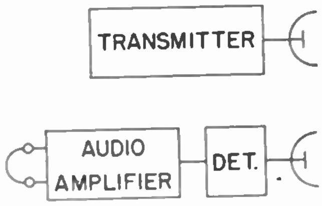

Fig. 1—Simple radar system.

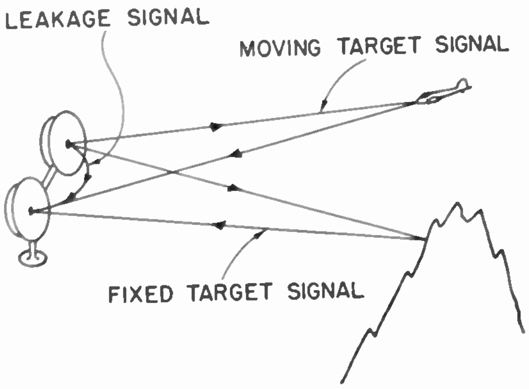

Fig. 2—Signals received by a radar system.

In the radar system of Figs. 1 and 2, the doppler effect can be shown very simply, as follows:

The phase difference between the leakage signal and the moving-target signal will depend on the distance to the target and on the system wavelength, being given by

$$\Delta\phi = \frac{4\pi r}{\lambda}$$

where

$\Delta\phi$ = phase difference
r = target range
λ = system wavelength.

As the target range changes, $\Delta\phi$ will change. A change in $\Delta\phi$ is equivalent to a frequency difference of the amount

$$\Delta f = \frac{1}{2\pi}\frac{d\Delta\phi}{dt},$$

which is equal to

$$\Delta f = \frac{4\pi}{\lambda}\cdot\frac{1}{2\pi}\frac{dr}{dt}$$

or

$$\Delta f = \frac{2v}{\lambda} = f_d,^{7} \text{ the doppler frequency,} \qquad (1)$$

where

v = the radial velocity of the target.

In the simple system of Fig. 1, the leakage signal and moving-target signal beat together in the detector and produce a signal at the difference of these frequencies which is the doppler frequency f_d. This doppler signal is amplified in the audio amplifier and is then fed to earphones. The fixed-target signal, being of the same frequency as the leakage signal, produces only dc in the audio amplifier and is, therefore, unnoticed by the operator. If a band-rejection filter were added to the audio amplifier, it would be possible to reject targets of some particular radial velocity; or, if desired, only a given radial velocity could be selected and all others rejected. It is this use of the doppler effect to select targets of certain velocities and discriminate against others which characterizes not only this simple system but all the radar systems discussed in this paper.

In Fig. 1 the output of the system is applied to earphones, but a frequency meter or velocity spectrograph or similar device could be used in their place. In any event, the information furnished by this simple system consists of an azimuth and elevation bearing, and radial velocity. Range information is not supplied. A system of this type has been described in the literature.[8]

One use for such a system might be detecting the presence of a moving object among a large number of fixed objects. For example, the system could be used as a watchman or sentry to detect a man walking or crawling in the vicinity of bushes, trees, or buildings. For this application, an ordinary pulse system would be practically useless because it would show all the fixed targets, and the moving target would be unnoticed among them. Furthermore, the minimum range of the system might be too great to detect near-by targets. Another application for this simple doppler system would be the accurate measurement of the velocity of bullets and projectiles of various sorts, and the investigation of their trajectories.

There are, however, several drawbacks to this simple system. First, there is no direct measurement of target range. Separate antennas are used for the transmitter and receiver, which makes the system physically larger

[7] This expression is only correct to the first power of v. The exact expression is

$$f_d = \frac{2v}{\lambda}\left(1 + \frac{v}{c} + \frac{v^2}{c^2} + \frac{v^3}{c^3} + \cdots\right).$$

[8] See chap. 5, sections 5 and 6, of footnote reference 1.

than a similar pulse system using only one antenna. Further, the presence of the leakage signal represents a limitation for several reasons. This signal makes it difficult to distinguish between approaching and receding targets. In the simple system described, a target approaching with a velocity v and one receding with this velocity both produce the same doppler frequency $f_d = 2v/\lambda$, and are thus indistinguishable. If the system were arranged so that, instead of producing a dc output for fixed targets, it would produce a signal at some frequency f, then approaching targets would have a frequency $f+2v/c$ and receding ones a frequency $f-2v/c$, so that the two could be distinguished. This can be accomplished by introducing into the receiver detector a signal at a frequency equal to the transmitter frequency minus f. The leakage signal in this case will produce a signal of frequency f which must be filtered out if moving targets are to be detected. It is quite possible for the leakage signal to be so much stronger than the desired target signals that it will be difficult to build a filter of sufficient attenuation at frequency f to suppress this leakage signal without seriously attenuating the moving-target signals. If there were no leakage signal present, a filter at frequency f might still be required to filter out fixed-target signals. However, these might be much weaker than the leakage signal. A discussion of the relative intensity of various signals will be considered later in this paper.

The leakage signal also causes difficulty if the system is operated from a moving base. In this case, the signals from fixed targets are doppler-shifted due to the velocity of the system itself. When these signals beat with the leakage signal, a doppler-frequency signal appears in the audio amplifier. This must be filtered out if it is desired to detect only objects moving with respect to the ground. The difficulty is that the frequency of the fixed-target signals is not constant, but depends both on the velocity with which the system itself is moving and on the angle between the system and the direction of movement. For example, suppose a system is mounted in an airplane flying at a speed v_0 while the system is scanning through 360° in azimuth. The frequency of the fixed-target signals will vary between a value of $2v_0/\lambda$ and zero as the system scans. Any filter to take out the ground-target signal would have to have its rejection frequency variable in accordance with the airplane velocity and with system azimuth. If, however, there were no leakage signal, the ground targets would show up as dc, and moving targets among the ground targets could be detected without the necessity for the complicated filter described above.

Another limitation imposed by the leakage signal is that its magnitude is normally large compared to that of the desired moving-target signal. Suppose, for illustration, that it is 1,000 times greater in voltage. Then modulation of this leakage signal due to hum, microphonics, fluctuation noise, etc., must be kept below 0.1 per cent or this modulation will mask the desired signal. In general, the greater the range of the desired target, the greater will be the ratio of the leakage signal to the desired signal, and the more stringent the noise-modulation requirements will be. Actually, doppler systems have been built which required a leakage modulation of less than 0.01 per cent. This stringent modulation requirement is one of the basic characteristics of the doppler type of system. Elimination of the leakage signal does not completely solve the difficulty, since the fixed-target signals would be modulated in accordance with the transmitter modulation. If fixed-target signals were large compared to the desired moving-target signals, there would still be stringent modulation requirements, but not as severe as before.

The importance of the noise modulation on the leakage signal will depend on the absolute amount of leakage power, which, in turn, will generally depend on the transmitter power. For a small system of very limited range, the problem may not be at all serious, and a system like that of Fig. 1 may be perfectly satisfactory. This would be the case for the watchman or sentry application mentioned previously, where the range on a person might be about one-half mile. For a system designed to detect an airplane 200 or more miles away, the transmitter power and hence the leakage signal will be several orders of magnitude greater, and the leakage signal may be prohibitively large. In this case, it can be eliminated by separating the transmitter and receiver. One method of accomplishing this is to locate the receiver at a point some distance from the transmitter. This is inconvenient for some applications, while satisfactory for others. For example, a system used for ballistics studies at a firing range could be permanently installed. In this case, space separation is quite satisfactory if one is willing to make certain geometric corrections to the doppler data. On the other hand, a system for continuous search or automatic tracking of moving targets, which must be transportable and might be sited in rough terrain, would be very inconvenient if physical separation of the transmitter and receiver were required. In this case, the time-separation principle can be used.

Time Separation

The transmitter and receiver are separated in time simply by switching them on and off alternately. This is similar to pulse-radar-system practice. It permits both transmitter and receiver to use the same antenna if a suitable switching arrangement is employed in the antenna feed to switch the antenna back and forth between transmitter and receiver at the correct time. Time separation also removes the leakage signal from the receiver input, so that there is complete control over the amplitude and frequency of whatever reference signal is injected to provide a beat with the doppler-shifted target signal. Such a reference signal is necessary, in general, since otherwise a single moving target will produce in the audio output only dc and harmonics of the switching frequency. This is exactly the effect produced by a fixed target, so that the two cannot be distinguished, and the whole point of doppler radar will be lost. In some applications, it might be that a fixed-target signal itself will provide a satisfactory reference signal. In general this is not true, because, if the system scans, there will probably be some ranges and azimuths where there is little or no fixed-target signal. A moving-target signal in these regions would be lost. Another disadvantage of relying on the fixed-target signal as a reference signal is that this signal, even in an unkeyed system, is not a pure frequency but, instead, a spectrum of frequencies due to system instability, ground motion, and system scanning. This means that the moving-target signal, even if pure itself at the receiver input, will appear in the audio amplifier as a spectrum of frequencies after beating with the ground-target signal in the detector. This is not a serious disadvantage if the signal merely goes to a pair of earphones, but in more complex systems, to be described later, in which range information and possibly tracking information are extracted from the signal, such a spectrum of frequencies may seriously reduce the sensitivity and accuracy of the system.

The time-separation principle is illustrated in Fig. 3. The switching or repetition frequency F is given by $F=1/T$. The fraction of the time that the transmitter is

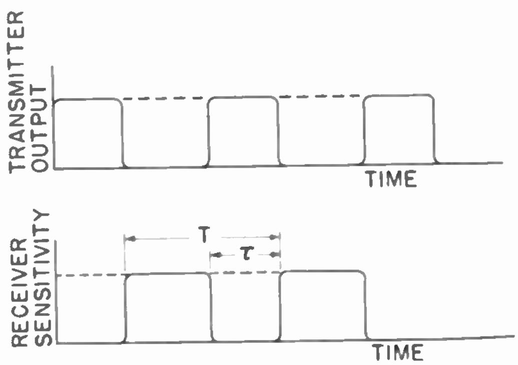

Fig. 3—Time-separation principle.

on, or its duty cycle, is given by τ/T. In a normal pulse system, τ/T will be of the order of 1/1,000. In general, this small value for τ/T is not characteristic of doppler systems because of the way in which fixed-target signals are eliminated, as will be explained.

Fixed Targets

In a time-separated system, the signal returned from an isolated fixed target when viewed in the audio amplifier (assumed wide-band) will consist of dc, the switching frequency F, and its harmonics. The number of harmonics that it is necessary to pass in order to preserve most of the energy in each pulse is approximately equal to T/τ, or the reciprocal of the duty cycle. For a normal pulse system, this means there will be something like 1,000 separate frequency components. The resulting spectrum for a fixed target and a moving target is shown in Fig. 4. Here, f_0 is the transmitter frequency, f_d is the doppler shift, and F is the repetition frequency. If a reference signal of frequency f_0 is introduced at the detector, the spectrum in the audio amplifier will be as shown in Fig. 5.

If it is desired to eliminate the fixed-target signals while still preserving the moving-target signals unaltered, a filter must be built having a rejection band at each harmonic of the repetition frequency F but passing all components of the moving-target signal. Such a filter would have a pass characteristic of the type indicated in Fig. 6. Each rejection band of this filter should have sufficient attenuation to reduce the strength of the fixed-target signal down to system noise level. It should also have sufficient bandwidth to reject the modulation of the fixed-target signal due to transmitter noise, small motions of the fixed target, motion of the radar beam across the fixed target, etc. Such a filter, applied to a pulse radar system to which a reference signal had been added, would result in a system responding only to moving targets; furnishing range, azimuth, and elevation information. The range information would be obtained by observing the time of arrival of the moving-target pulse with respect to the emission time of the transmitter pulse.

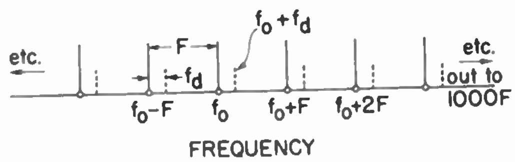

Fig. 4—Frequency spectrum of signals at the input to a radar receiver.

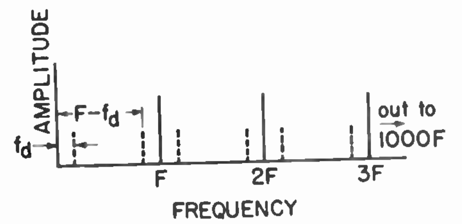

Fig. 5—Frequency spectrum of signals at the output of a radar receiver detector.

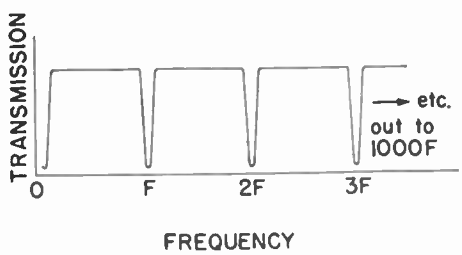

Fig. 6—Frequency characteristic of the desired fixed-target rejection filter.

Unfortunately, the filter described above is extremely complex and to date has only been approximated in practice. Approaches to this ideal filter have been along two paths. The first has been to make an equivalent filter having as many rejection sections as desired, but by its very nature possessing a depth and width of each section which cannot be satisfactorily adjusted. An example of such a filter is shown in Fig. 7. This shows a delay line which delays the signals passing through it by a time exactly equal to T, the time between successive transmitter pulses. This means that, in the subtractor, the pulse due to each target is subtracted from its own previous pulse. Fixed-target pulses should then cancel out, since they do not change from one period to the next, while moving-target pulses should not cancel out, since their phase with respect to the reference signal, and hence their amplitude in the audio circuit, is continually changing. This scheme has the equivalent filter characteristic shown in Fig. 8. This is of the form of a rectified sine wave.[9]

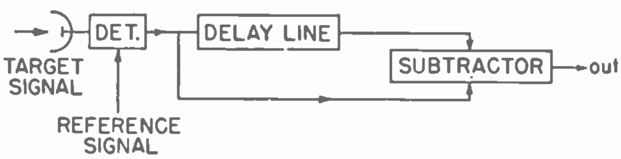

Fig. 7—Fixed-target rejection filter using a delay line.

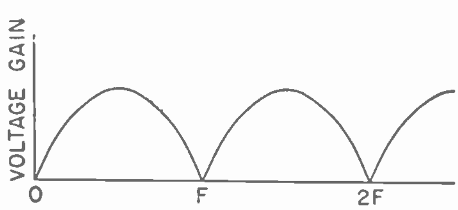

Fig. 8—Frequency characteristic of the delay-line filter.

Each rejection section of such a filter is too narrow at the bottom and too wide at the top for completely satisfactory rejection of fixed-target signals. Furthermore, in passing through the delay device, pulses are slightly distorted, so that fixed-target pulses do not cancel exactly in the subtractor, and, to date, attenuation of fixed-target signals has not generally been greater than about 30 db.

The other approach has been to build each rejection section of R, L, and C elements and electron tubes, as necessary, so that the desired attenuation characteristics could be achieved. Filters have been built for which each rejection band has satisfactory characteristics, but, so far, they have not contained more than a few sections because of their complex nature.

If a one- or two-section filter is used, it must be accompanied by a low-pass filter to remove all the high harmonics of the fixed-target signals. Unfortunately, this removes also the high harmonics of the moving-target signals and so destroys their character as sharp pulses. It is then impossible to determine range by observing the time at which a moving-target pulse is received, and some other method of measuring range must be used. Since high harmonics of the repetition rate F are thrown away, there is no longer any point in transmitting short pulses. Consequently, a duty cycle is chosen having the most possible energy in its low harmonics. This is a duty cycle of 1/2. It is this second approach—namely, limiting the number of signal harmonics to one or two and making the ground-target rejection filters deep enough and wide enough for satisfactory rejection—which will be discussed in the remainder of this paper.

Although this approach permits ground-target rejection filters of great attenuation to be built, it should be pointed out that the sacrifice in range resolution necessitated by limiting the number of signal harmonics increases the magnitude of the fixed-target signals relative to the moving-target signals. This is so because any given moving target must compete with the ground targets within the same "pulse packet" and the doppler system with a duty cycle of 1/2 has a much larger "pulse packet" than a high-resolution pulse system. Thus, going from a high-resolution design to a duty cycle of 1/2 permits the construction of fixed-target rejection filters of greater attenuation, but, by increasing the relative size of the fixed-target signals, greater attenuation is required for the same detectability of the moving targets among fixed targets. Which of these factors outstrips the other and provides greater attenuation of fixed-target signals relative to moving-target signals will depend on the state of development of system components and techniques.

It should be pointed out that the amount by which the fixed-target signals are reduced relative to the moving-target signals as the pulse packet is made smaller will depend on the nature of the fixed targets. If a large number of small fixed targets are more or less uniformly distributed in range, the fixed-target signals will vary in power roughly as the pulse length. On the other hand, if the fixed targets exist in a few sharply defined places, a reduction in the pulse length will tend to reduce the fraction of the system response area covered by fixed-target signals without much reducing their peak magnitude. Thus the difference in magnitude of fixed-target signals in the pulse and cw cases may be much less than the difference in system duty cycles would at first lead one to suppose.

For a cw radar system, a theoretical calculation can be made of the magnitude of the ground-target signal and hence of the attenuation required in the ground-target rejection filter, assuming an idealized "rough earth." This calculation can be supported by experimental evidence gathered from the operation of radar systems during World War II. There is also experimental evidence about the fluctuations of the fixed target signals due to motion of trees, bushes, etc., from which the signal is being reflected.

From all this information, fixed-target rejection filters of a few sections have been designed which will satisfactorily reject the fixed-target signals in a keyed cw doppler system. By "fixed target" is meant hills, mountains, cities, water tanks, large buildings, etc. Of these, the targets which are the most difficult to reject are mountains covered with trees, especially on a windy day. The swaying of the trees causes rapid fluctuations

[9] See p. 650 of footnote reference 1.

in the intensity and phase of the returned signal. This necessitates a large bandwidth in the rejection filters. Some other targets require even larger bandwidths for satisfactory rejection. They should not really be spoken of as fixed targets but rather as undesirable slowly moving targets. Examples of these are rain clouds and rain storms, snow storms, waves and spray on the surface of the ocean (called "sea echo"), and the strips of tin foil called "window" dropped by airplanes to confuse radar systems. All these targets can be eliminated by proper design of the fixed-target rejection filters. Of course, some desired targets may also be eliminated if the rejection bandwidths become too great, so that some compromise of bandwidth is necessary.

The simple system of Fig. 1 now can be improved by the incorporation of the time-separation principle, suitable fixed-target rejection filters (FTR), and a low-pass filter. It then becomes the system shown in Fig. 9. This system is an improvement on the simple system in several ways. First, it uses only one antenna. There is no leakage signal, so that greater transmitter powers can be used, and noise troubles in the transmitter are less acute. It is possible to operate this system from a moving base and it is possible to distinguish between approaching and receding targets.

This system furnishes azimuth, elevation, and radial-velocity information just as did the simple system of Fig. 1. It will supply all the information that the simple system did, and in addition can be used for such applications as detecting moving cars, trucks, trains, etc., from an airplane, studying projectile trajectories out to much greater distances, and detecting the approach of men or vehicles not only in the presence of a large number of fixed targets but also in the presence of other moving targets.

With all these advantages, this system still has a serious drawback, for it lacks range information. It has been pointed out above that, because of the way high harmonics of the repetition frequency are filtered out, pulse ranging technique cannot be used and some other method of ranging must be found.

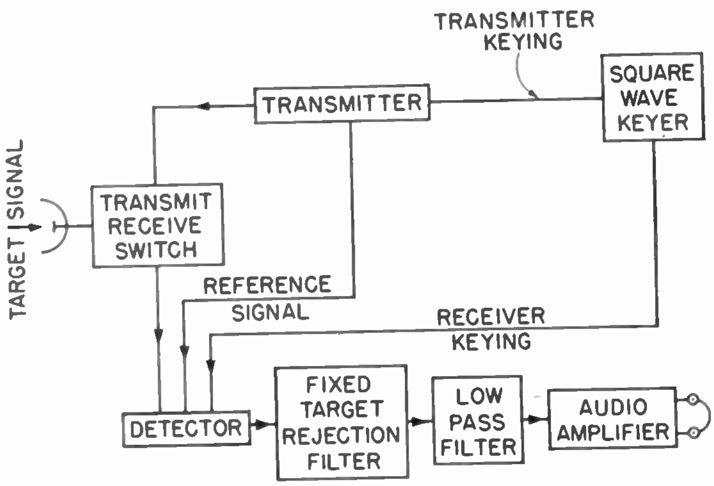

Fig. 9—Doppler radar system using the time-separation principle.

Ranging Methods

One ranging method which is very simple in principle is the two-frequency method. This can best be illustrated by visualizing two entirely separate doppler systems, side by side, having carrier frequencies which differ by a frequency f_r. This difference in carrier frequencies implies a difference in doppler frequencies of

$$\Delta f_d = 2f_r \frac{v}{c}.$$

Now, imagine a target starting from the system and moving away at the velocity v. Initially, the two doppler-frequency signals from the two separate systems will be n phase, but as the target moves away they will differ in phase by an amount $\Delta\phi$ where

$$\Delta\phi = 2\pi\Delta f_d t = \frac{4\pi f_r}{c} vt = \frac{4f_r}{c} \cdot r$$

where r is the target range. The frequency difference f_r will be called the ranging frequency. The formula shows that the target range is directly proportional to the phase angle $\Delta\phi$ between the two doppler signals. A complete system employing this ranging principle is shown in Fig. 10. In this system, the time-separation principle and a two-channel superheterodyne receiver

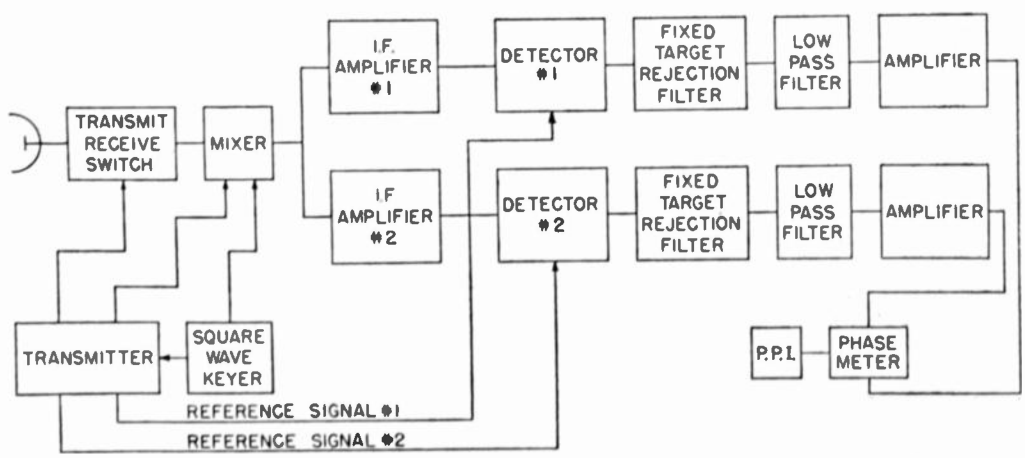

Fig. 10—Doppler radar system using the two-frequency ranging principle.

are combined in such a way as to make two completely separate radar systems unnecessary. One of the receiver channels works with the transmitter frequency, while the other channel works with one of the keying side-bands. This means that the ranging frequency f_r is equal to the keying frequency F or some harmonic of F.

The output of the phase meter gives the range of a target, while the position of the antenna gives the target azimuth. These two quantities are sufficient to permit a PPI (plan-position indicator) to be used. This plots the target on co-ordinates which are a scaled-down replica of the area swept by the system. This PPI will be similar to that of an MTI pulse radar system, in that only moving targets will be shown. The smaller system bandwidth means, however, that the noise background will be much "coarser." In addition to the PPI, targets are also indicated by a pair of earphones connected to one of the audio channels. These earphones could be replaced by a frequency meter which would give radial-velocity information.

This system makes obvious one drawback of the doppler ranging system. Since the phase meter can only indicate one phase at a time, the system as a whole can indicate only one target at a time, or, if the system scans, only one target per beam width. This is the limit to the information rate of the system. If two targets are in the beam at the same time, the phase meter will indicate a range somewhere between them depending on the relative strength of their two signals. This makes such a system useless in any area which is congested with moving targets. Hence, it cannot be used for airport traffic control if the traffic is at all heavy. Such a system as this finds its primary application in detecting or tracking an isolated target through or near many fixed targets which might be large enough to confuse pulse MTI systems.

Another system which furnishes range information is that shown in Fig. 11. This has the advantages over the previous system that only one channel is required in the receiver and that the phase meter operates at one frequency, the keying frequency, rather than being required to operate at any doppler frequency as in the previous system. Also, there are three types of presentation of target information, PPI, earphones, and A-scope. The operation of this system is as follows: The transmitter and receiver are keyed by the square-wave keyer so as to be on alternately. This means that, at the detector output, a fixed-target signal will have frequency components at dc, the keying frequency F, and harmonics of F. The low-pass filter cuts off at F, so the harmonics are all eliminated. The FTR filter has two rejection bands, one at dc and one at frequency F, so that all components of fixed targets are eliminated. At the detector output, a moving target will have components at the frequencies f_d, $F-f_d$, $F+f_d$, $2F-f_d$, $2F+f_d$, etc. The low-pass filter removes all of these except the two at f_d and $F-f_d$. Provided that $f_d<F$,[10] these are unaffected by the FTR filter, and so they pass to the rectifier. The rectifier, being a nonlinear device, creates new components at frequencies equal to the sum and difference of the original frequencies plus various multiples and combinations of these. One component will be at a frequency equal to the sum of the frequencies of the entering components which is just F, the keying frequency. The narrow-band-pass filter is centered at this frequency, so that it selects only this component, eliminating the others. The phase of this component relative to the square-wave keying voltage is a measure of the target range, and so this information is put on the PPI.

As indicated in Fig. 11, there is provision for earphones (or a loudspeaker). These can be replaced or supplemented by a frequency meter, if desired, so that radial-velocity information will be furnished. The new type of presentation is the A-scope. This is simply an oscilloscope showing the detector output, plotting amplitude on the vertical axis against time on the horizontal. In a region of ground targets, the A-scope shows so

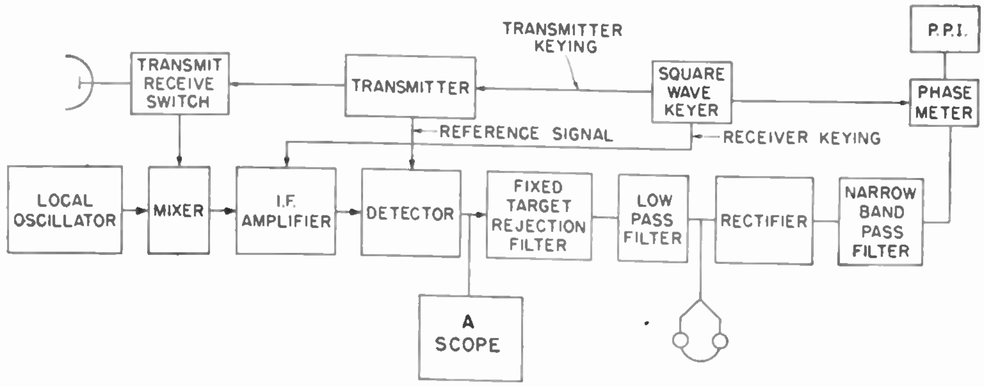

Fig. 11—Doppler radar system using the keying frequency modulation of the target signal to measure range.

[10] If f_d is greater than F, the low-pass filter will pass a different pair of components of the moving-target signals, but these will also contain the desired target information. If the doppler frequency is very high, the signal components passed by the low-pass filter may contain very little energy, so that the system sensitivity is reduced. This difficulty may be circumvented by providing one or more additional receiver channels having band-pass rather than low-pass filters, and choosing the pass bands in accordance with the desired doppler frequency ranges.

much clutter that moving targets are unnoticeable but, of course, the earphones and PPI are unaffected by the ground targets if the ground-target rejection filters are performing satisfactorily. Just as for the previous system, this one will only respond to one target at a time, and so its applications and limitations are the same.

To make the operation of this system clearer, some sketches of the various signals are shown in Fig. 12. Figs. 12(a) and (b) show the transmitter and receiver square-wave keying. Fig. 12(c) shows the signal due to a single moving target. It consists of a sine wave at the doppler frequency multiplied by a rectangular wave due to the combined effects of transmitter and receiver keying. Fig. 12(d) shows the same signal after full-wave

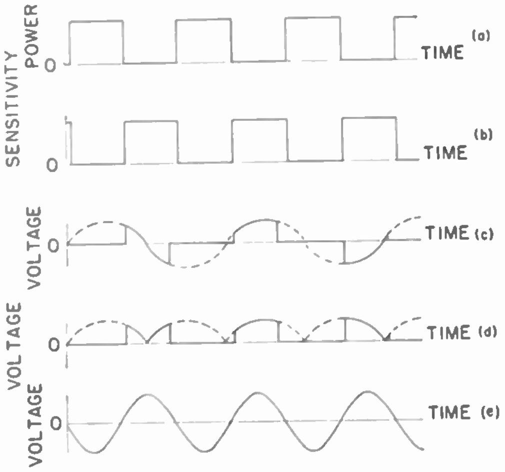

Fig. 12—Wave shapes of various signals present in the doppler radar system of Fig. 11.
(a) Transmitter output.
(b) Receiver sensitivity.
(c) Signal at the input to the fixed-target rejection filters due to a single moving target.
(d) The same signal after passage through a full-wave rectifier.
(e) The same signal at the output of a narrow-band-pass filter.

rectification. The signal has in it a component at the keying frequency which can be extracted by passage through a narrow filter turned to this frequency, as shown in Fig. 12(e). This last sketch shows a sine wave at the keying frequency whose phase relative to the phase of the square-wave keying voltage is a measure of target range.[11]

Range Accuracy and Resolution

For the doppler ranging methods so far described, the range information is in the form of the phase angle of a sine wave which is in the presence of inevitable noise voltages of the receiver. This noise makes measurement of phase somewhat uncertain, and thus introduces a range error. The range error gets less as the S/N (signal-to-noise) ratio increases, but there will always be some error due to this effect. For very weak signals the range error could easily be as large as 25 to 50 per cent of maximum range. The case is very different for a pulse radar set because any target which can be detetected at all has an uncertainty in range of less than the pulse length, which means a maximum range uncertainty of less than the duty cycle or 0.1 per cent in a typical case.

By "range resolution" is meant the ability of the system to indicate two targets differing only slightly in range, both on the same azimuth. Since the doppler systems previously described cannot handle more than one target on any azimuth, they cannot be said to have any range resolution.

The doppler radar system of Fig. 11 can be compared to a pulse system in order to assess the advantages and disadvantages of the doppler approach. Since doppler information is inherent in all radar systems, it seems most reasonable to compare the performance of the system of Fig. 11 to that of a pulse system with MTI features. Pulse systems of this type are described in the literature.[12]

Advantages and Disadvantages of Doppler Radar

Some of the differences between the two systems are:

(1) The doppler system can handle only one target at a time, or roughly one target per beamwidth for a scanning system. By contrast, a high-resolution pulse system has something like 1,000 separate range elements, and hence can handle many targets per beam width.

(2) The range accuracy of the doppler system depends on the target signal-to-noise ratio and, in general, is worse than that of a pulse system. For certain critical doppler frequencies (related to the ranging frequency), the range accuracy is particularly poor.

(3) The doppler system can be arranged to give an accurate indication of the radial velocity of targets. This information is also available from the pulse system but is difficult to extract without loss of accuracy or sensitivity.

(4) The doppler system permits of great flexibility in choosing the relative system response to targets of different radial velocity. Thus, provision can be made for adequate rejection of mountain echoes even if the mountains are heavily wooded. Storm-cloud echoes, sea echoes, etc., can also be eliminated, and various relative responses to different airplane velocities, etc., can be arranged. Among the limitations to the flexibility of design is the fact that only radial velocity counts, so that a fast-moving target traveling almost tangentially must be accorded the same treatment as a slow, radially moving target. Also, if an undesirable storm echo and a desirable car, truck, etc., echo occur at the same doppler frequency, they must be accorded the same treatment.

Another restriction on the velocity-response characteristics of the doppler system is that targets of doppler frequency equal to the ranging frequency must be rejected. This is also true of the simple pulse MTI system. For both types of system it is possible to avoid

[11] See chap. 5, section 11, of footnote reference 1 for a description of this system.

[12] See chap. 16 of footnote reference 1.

this difficulty by several methods which, however, increase the system complexity.

The pulse MTI system using one delay line and cancellation circuit has essentially a fixed velocity-response characteristic.[13] This characteristic can, in principle, be modified, but only at the price of greatly increased system complexity. For example, the delay line could be replaced by a series of rejection filters, one at each repetition-rate harmonic, having the desired frequency-response characteristics. Each such filter would be as complicated as the filter needed by the doppler system for the same velocity-response characteristic, and the pulse system might have as many as 1,000 such filters.

(5) The doppler and pulse systems differ in their ability to reject large fixed-target signals. Field experience with pulse MTI systems indicates relative fixed-target attenuation of roughly 30 db, while the doppler system of Fig. 11 has achieved over 90 db attenuation of such signals.[14] Bearing in mind that the doppler system needs more fixed-target attenuation because of the larger "pulse packet" involved, field experience has shown that in really hilly terrain the doppler system has been able to reduce the fixed signals below the noise level of the set, while pulse MTI systems at the same location have been seriously troubled with fixed-target echoes. Part of the difference between the ground-target attenuation realized by the two systems is because of the different design problems and practical limitations of the fixed-target rejection filters and delay lines, and part is also due to the difference in the amplitude and frequency stability of the transmitter, local oscillator, and reference signal in the two cases. Pulse MTI systems have generally employed magnetron transmitters and reflex-klystron local oscillators, while the doppler system described in Fig. 11 was completely crystal-controlled. A master crystal oscillator followed by frequency multipliers provided the local oscillator, reference signal, and transmitter drive. The transmitter output tube was a triode.

The relative importance of some of these advantages and disadvantages of doppler systems will depend greatly on the application considered. For example, suppose an airport surveillance radar is requested in a location not extremely hilly. For this application, the velocity-response characteristic of the pulse MTI system is fairly satisfactory. The 30-db fixed-target attenuation is adequate, and the range accuracy and range resolution of the pulse system are needed. The doppler system will fall down on the counts of target-handling capacity and, possibly, range accuracy, and will have more clutter rejection than necessary.

On the other hand, if there is need for a radar in very rough country, and clutter rejection is important and further moving targets are infrequent but imperative to detect when present, the doppler system of Fig. 11 would perform very well, while the pulse system might be inadequate.

[13] See p. 650 of footnote reference 1.
[14] See p. 155 of footnote reference 1.

Calculation of Performance

It is possible to replace the qualitative statements which have been made about range accuracy, ground-target size, etc., by specific calculations.

I. Sensitivity

The calculation will first be made for the simple doppler system of Fig. 1 and then modified for the system of Fig. 11. Let

P_t = transmitter power
λ = transmitter wavelength
d = antenna diameter (the antenna is assumed to be a paraboloidal reflector fed by a dipole)
r = target range
θ = antenna beamwidth at the half-power points
G_0 = antenna gain relative to an isotropic radiator (assumed the same for transmitter and receiver).

The power per unit area at the target due to the transmitter will be

$$P = P_t \times \frac{1}{4\pi r^2} \times G_0.$$

Let the target be represented by an equivalent scattering cross section Σ visualized as scattering back uniformly over a sphere all the energy incident upon it. The proper value of Σ for an actual target can be found by experiment or can be calculated for some simple cases. On this basis, the power received by the target is

$$P = P_t \frac{1}{4\pi r^2} G_0 \Sigma,$$

and the power at the receiver is

$$P_r = P_t \frac{\lambda^2 \Sigma G_0^2}{64\pi^3 r^4}. \tag{2}$$

The quantity which is significant is not the absolute magnitude of the received power but rather the signal-to-noise ratio at the output of the af amplifier. This can be calculated in terms of two additional quantities, the receiver noise figure (NF) and the effective bandwidth ΔF. The equivalent noise power at the receiver input is

$$P_n = (NF) kT\Delta F.$$

Thus the signal-to-noise ratio in power, S/N, is given by

$$S/N = \frac{P_r}{P_n} = \frac{P_t \lambda^2 \Sigma G_0^2}{64\pi^3 r^4 (NF) kT\Delta F}. \tag{3}$$

It has been shown that, for a paraboloid reflector, G_0 is related to the paraboloid diameter d in the following way:

$$G_0 = \frac{\pi^2 d^2 \eta}{\lambda^2} \tag{4}$$

where η is an efficiency factor whose value depends on the uniformity of illumination of the paraboloid. A representative value of η for a dipole feeding a paraboloid, for example, is about 0.6.

Using this expression for G_0 in (3), we get

$$S/N = \frac{\pi}{64kT} \frac{P_t \Sigma d^4 \eta^2}{\lambda^2 r^4 (NF)\Delta F}. \tag{5}$$

The noise figure of a microwave receiver has been defined and methods of measurement discussed elsewhere.[15,16] The effective bandwidth ΔF depends, in general, on the bandwidths of the if and af amplifiers and the type of detector used, but in the case of the simple doppler set, as long as the leakage carrier is large compared to the noise voltage at the detectors and as long as the if bandwidth is greater than twice the af bandwidth, the effective bandwidth ΔF is just twice the af bandwidth.

On applying this formula to the system of Fig. 11, it can be seen that there are really two af bandwidths, the bandwidth of the low-pass filter and the bandwidth of the band-pass filter. The effective bandwidth depends on both of these and on the S/N ratio at the rectifier. Without going into details, it will be stated that the S/N ratio at the input to the phase meter is given very nearly by

$$S/N^{17} = \frac{F_1}{2F_2} \frac{\left(\frac{\pi}{64kT} \cdot \frac{P_t \Sigma d^4 \eta^2}{2F_1 NF \lambda^2 r^4}\right)\left(\frac{r}{r_0}\right)^2}{\left\{1 + \frac{1}{4}\left(\frac{r_0}{r}\right)^2 \left(\frac{\pi}{64kT} \cdot \frac{P_t \Sigma d^4 \eta^2}{2F_1 NF \lambda^2 r^4}\right)^{-1}\right\}} \tag{6}$$

where

F_1 = bandwidth of low-pass filter

F_2 = bandwidth of band-pass filter

P_t = average transmitter power

$2r_0$ = maximum unambiguous range

and $F_1 \gg F_2$. This formula assumes that, if there is any modulation of the doppler signal, the modulating frequencies are smaller than $F_2/2$.

Equation (6) can be written in a somewhat more convenient form by realizing that, for the simple system of Fig. 1, the range for which $S/N = 1$ is given by

$$r = \left[\frac{\pi}{64kT} \frac{P_t \Sigma d^4 \eta^2}{2F_1 NF \lambda^2}\right]^{1/4} \tag{7}$$

where now F_1 is the af bandwidth of the simple system. If this range be called r_1, (6) can be rewritten as

$$S/N = \frac{\left(\frac{r}{r_0}\right)^2 \left(\frac{r_1}{r}\right)^4 F_1}{F_2 \left\{2 + \frac{1}{2}\left(\frac{r}{r_1}\right)^4 \left(\frac{r_0}{r}\right)^2\right\}}. \tag{8}$$

This formula is valid only for $r < r_0$, and assumes a square-law rectifier. If, as a further simplification, the repetition frequency is chosen so that $r_0 = r_1$, the relation becomes:

$$S/N = \frac{\left(\frac{r_0}{r}\right)^2 F_1}{F_2 \left\{2 + \frac{1}{2}\left(\frac{r}{r_0}\right)^2\right\}}. \tag{9}$$

This is not too unreasonable an assumption, since, for the radar system to be used as an example later, $r_0 = 47$ miles and $r_1 = 60$ miles.

The preceding analysis has yielded an expression for the output S/N ratio of the doppler system of Fig. 11. This S/N ratio is, however, only one of the factors entering into the analysis of the detectability of target signals. Other important factors include type of presentation (i.e., PPI, A-scope, earphones, etc.) and, if cathode-ray-tube presentation is used, such things as spot size, sweep speed, minimum change in brilliance which can be detected by the eye, etc. The duration of the signal, the amount of signal integration which is employed, and the behavior of the integrating device also are important. In order to determine the true range performance of a radar system, an analysis of the effects of these factors is necessary. Some work has been done along this line in connection with the detection of short-pulse radar signals on PPI tubes and A-scopes.[18,19]

Relatively little work seems to have been published on the detectability of signals from cw-type radar systems. For this reason, it is difficult to compare the range performance of cw and pulse systems on a target-detectability basis. A definite comparison can be made, however, on an output S/N ratio basis, and in the absence of further information this must suffice.

It was stated earlier that, in principle, pulse and cw systems were capable of approximately the same range performance for the same average power. This statement may now be amplified by the addition of the above restriction to an equal S/N ratio comparison basis. Furthermore, in order to achieve its maximum range performance, the cw-type system should have an effective system bandwidth chosen solely to maximize the S/N ratio, with no regard for required doppler-frequency pass bands. In the system of Fig. 11, the bandwidth of the narrow-band filter is chosen in accordance with this criterion, but the bandwidth of the low-pass filter has been made much greater in order to pass signals of various doppler frequencies. This results in a reduction in the system's range performance. The S/N ratio of this system has been plotted as a function of range in Fig. 13. For comparison, the S/N ratio of a simple cw sys-

[15] D. O. North, "The absolute sensitivity of radio receivers," *RCA Rev.*, vol. 6, p. 332; January, 1942.

[16] H. T. Friis, "Noise figures of radio receivers," Proc. I.R.E., vol. 32, pp. 419–422; July, 1944.

[17] For derivation, see Appendix I.

[18] A. V. Haeff, "Minimum detectable radar signal and its dependence upon parameters of radar systems," Proc. I.R.E., vol. 34, pp. 857–862; November, 1946.

[19] B. M. Ashby, V. Josephson, and S. Sydoriak, "Signal threshold studies," NRL Report R-3007, December 1, 1946.

tem with optimum bandwidth to maximize the S/N ratio has also been plotted. The system of Fig. 11 has a S/N ratio less than the optimum-bandwidth system because of the large effective bandwidth, because of the fact that the keying of the receiver prevents it from receiving all the signal energy at the antenna, and because not all the frequency components of the signal spectrum are used to produce the output wave. If the system of Fig. 11 were converted to a pulse system of the same average power and scanning speed, it would still not have the performance of an optimum-bandwidth system because it would have many hits per target which would have to be integrated. These hits cannot be integrated with complete effectiveness. The estimated performance of such a pulse system is also shown in Fig. 13.[20]

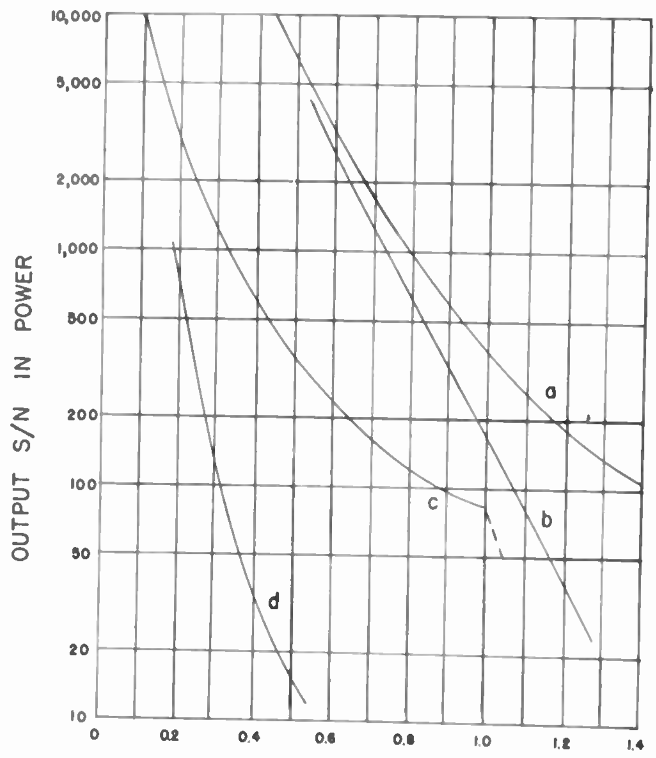

Fig. 13—Comparative radar system performance.
(a) Optimum-bandwidth system.
(b) Equivalent pulse system.
(c) Doppler system of Fig. 11.
(d) Simple doppler system.

The simple doppler system of Fig. 1 and equation (5) has a performance much worse than the optimum-bandwidth system because of its relatively large bandwidth. Its performance is also shown in Fig. 13. For all these curves, the values of F_1 and F_2 chosen were

$$F_1 = 1{,}000 \text{ cps}$$

$$F_2 = 5 \text{ cps}.$$

[20] The integration devices assumed in the computation of this curve were a series of range gates followed by narrow-band video filters.

This corresponds to a radar system requiring a doppler frequency range of ±1,000 cps and a scanning speed such that the time spent in sweeping through one beam width is approximately $\frac{1}{5}$ second.

Although, in Fig. 13, the performance of the simple doppler system looks very poor, it should be remembered that, if the output of the system is fed into earphones, the effective system bandwidth will depend on the ability of the human listener to pick an audio tone out of a noise background. Some experimental work on this point indicates an effective noise bandwidth of the human auditory mechanism of roughly 100 cps for sustained tones. For tones of short duration, this effective bandwidth increases. This means that the relative performance of the simple doppler system using earphones is probably not so bad as is indicated by Fig. 13.

II. Range Error

If the phase angle of a sine wave is measured in the presence of noise, there will be a fluctuation in the answer obtained, there being a certain probability of observing any given phase. The probability distribution of the phase angle will depend on S/N, the signal-to-noise ratio in power. This distribution is given by[21]

$$P(\theta - \theta_0, S/N) = \frac{1}{2\pi} e^{-S/N} + \frac{1}{2}\sqrt{\frac{S/N}{\pi}} \cos(\theta - \theta_0)$$

$$\cdot e^{-S/N \sin^2(\theta-\theta_0)} \left\{ 1 + erf\left(\sqrt{\frac{S}{N}} \cos(\theta - \theta_0) \right) \right\} \quad (10)$$

where θ_0 is the phase of the sine wave, and θ is the phase actually measured.

From this distribution, the rms phase error $\sqrt{\overline{(\theta-\theta_0)^2}}$ or $\Delta\theta_{rms}$ can be determined. For large S/N,

$$\Delta\theta_{rms} \cong \frac{1}{\sqrt{2S/N}} \cdot \quad (11)$$

Equation (10) has been combined with (9) and the rms error has been plotted as a function of $(r)/r_0$ in Fig. 14. The relation between θ_0 and r has been assumed to be $\theta_0 = \pi r/2r_0$. This is consistent with the derivation of (8), in which the maximum unambiguous range was $2r_0$.

To give a better idea of what this performance actually corresponds to, the following values have been chosen to compute r_0:

$P_t = 100$ watts
$\Sigma = 50$ square feet (a small airplane)
$d = 12$ feet
$\eta = 0.6$
$\lambda = 1$ foot ($f_0 = 1{,}000$ Mc)
$NF = 20$
$F_1 = 1{,}000$ cps.

[21] See Appendix II.

These values give an r_0 of approximately 60 miles. Thus, for a small airplane 60 miles away, the rms range error can be expected to be about 3 miles.

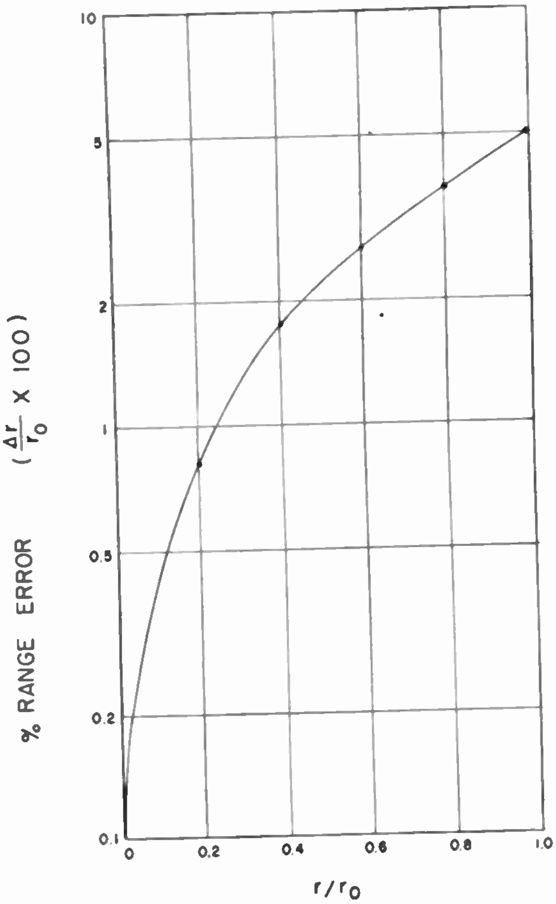

Fig. 14—Range error of doppler radar system of Fig. 11 as a function of target range.

III. *Fixed Target Rejection*

Experimental measurements have been made of the properties of the following:

1. Ground targets (hills, mountains)
2. Sea echo
3. Rain clouds
4. "Window."

A frequency analysis has been made in terms of what will be called the "power spectrum." This means that, if a voltage $V(t)$ is analyzed and its Fourier transform $G(f)$ is found, the power spectrum $W(f)$ is defined as

$$W(f) = \frac{G(f)^2}{G(0)^2}$$

where $G(0) = G(f)$ at zero frequency. The power spectrum of the fluctuations in the return of the four types of "fixed" target mentioned above has been found to be approximately of the form:

$$W(f) = e^{-af^2/f_0}. \tag{12}$$

The value of a will depend on the nature of the ground targets or clouds or waves, as the case may be, and for all types of fixed targets there will be a dependence on the weather conditions. Some representative values for a are given in Table I.

TABLE I

Target	*a*	*Curve in Fig. 15*
Heavily wooded hills, 20-mph wind blowing	2.3×10^{17}	1
Sparsely wooded hills, calm day	3.9×10^{19}	2
Sea echo, windy day	1.41×10^{16}	3
Rain clouds	2.8×10^{15}	4
Window "jamming"	1×10^{16}	5

For the transmitter frequency of 1,000 Mc, used in the calculation of r_0, the power spectra of these types of target are plotted in Fig. 15. The width of the spectrum band needed to reject these undesired signals can be determined directly from these data.

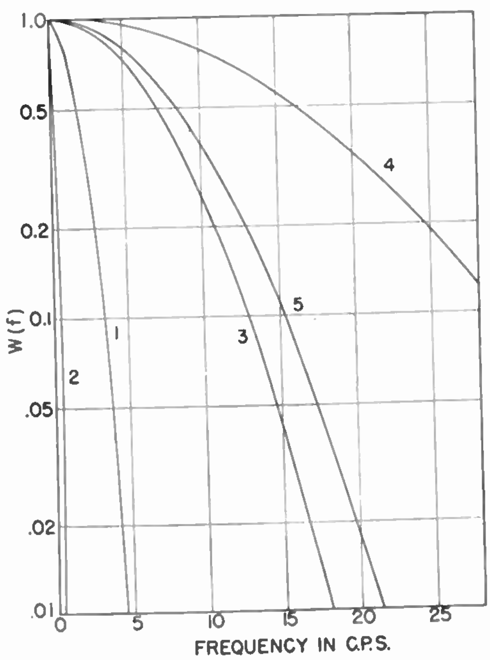

Fig. 15—Frequency spectra of various types of fixed targets. See text to identify curves.

A rough calculation can be made of the intensity of the return to be expected from ground targets. Assume that at a distance r there are mountains large enough to fill the transmitter beam and rough enough to scatter uniformly over a hemisphere all the energy incident upon them. In this case, the average power of the signal received at the antenna terminals is

$$P_r = P_t \times \frac{\lambda^2}{4\pi} G_0 \frac{1}{2\pi r^2}$$

$$= \frac{P_t \lambda^2 G_0}{8\pi^2 r^2}. \tag{13}$$

To apply this result to the system of Fig. 11, the receiver gating and the effect of the detector action and filter bandwidths should be taken into account. An approximate result may be obtained by multiplying P_r by $(r/r_0)^2$ to account for the receiver gating, and by taking the effective system bandwidth to be the geometric mean of the two bandwidths F_1 and F_2.[22] In this case, it follows that the attenuation required to get the ground-target signal equal to effective receiver noise is given by

$$\beta = \frac{P_r\left(\frac{r}{r_0}\right)^2}{NFkT\sqrt{F_1F_2}} = \frac{P_t\lambda^2G_0}{8\pi^2r_0^2\,NFkT\sqrt{F_1F_2}}$$

$$= \frac{P_td^2\eta^2}{8r_0^2NFkT\sqrt{F_1F_2}}. \tag{14}$$

To get an idea of the order of magnitude of β, the following values will be taken for the quantities in the equation above:

$P_t = 100$ watts
$d = 12$ feet
$\eta = 0.6$
$r_0 = 3\times10^5$ feet (approximately 60 miles)
$NF = 20$
$F_1 = 1{,}000$ cps
$kT = 4.1\times10^{-21}$
$F_2 = 5$ cps.

This choice gives a β of 91 db.

Several experimental measurements of the magnitude of ground targets give somewhat smaller results than the preceding calculations; in fact, 10 to 20 db smaller (in hilly country, not jagged peaks). Presumably, this is because they are not hemispherical and their radar albedo is less than unity. The experimental data give the doppler system which was used in calculating β an actual ground-target attenuation of about 80 db required. This is a representative value for doppler systems built to date, and gives an idea of the magnitude of the filter problem.

For sea echo, clouds, and "window" the intensity will be lower than for ground targets, so that a filter having enough attenuation for ground targets will be satisfactory for the other types of clutter, provided it also meets the frequency requirements.

So far, only the frequency spectrum of a fixed target resulting from motion of the target itself has been presented. Another effect which can modify the spectrum as seen at the filter is the motion of the radar beam from one section of the fixed target to another. As the beam swings past ground targets, for example, the amplitude of the ground-target signal in the receiver will fluctuate up and down. This will result in a frequency spectrum for the ground-target signals. The width of this spectrum will increase as the scanning speed increases and as the transmitter beam is sharpened. It can be shown that the highest frequency in this so-called "scanning spectrum" is the doppler frequency associated with the fastest-moving part of the antenna structure as it rotates. If the antenna paraboloid diameter is d and if the speed of rotation is N rpm, this highest frequency will be

$$f_{max} = \frac{\pi Nd}{30\lambda} \tag{15}$$

where, again, λ is the system wavelength. The FTR filters must have rejection bands wide enough to take care of this spectrum.

An alternative way of looking at this relationship is to say that the ratio of the system repetition frequency to the scanning spectrum width is approximately equal to the number of repetition periods the system spends scanning past any one target. It can be seen that, once the width of the rejection bands of the fixed-target rejection filters of a doppler radar system are chosen, an upper limit to the system scanning speed is set.

IV. System Stability Requirements

It was pointed out in the discussion of the leakage signal that one of the characteristics of a doppler radar system is the stringent stability requirement imposed on the transmitter output and reference signal. The stability required will depend on the application and on the particular system being used. To illustrate some of the fundamental requirements, consider first the simple system of Fig. 1. Any amplitude modulation of the leakage signal will be detected and will appear in the af amplifier of this system. Very low and very high frequencies can be filtered out without detriment to the operation of the system. Any modulation at a frequency corresponding to the doppler frequency of the moving targets for which the system is designed cannot be filtered out without removing the desired target signals. Such modulation components must be kept down to the receiver noise level, or the system sensitivity will be impaired. Amplitude modulation of the leakage signal might be caused by power-supply hum, microphonics, fluctuation noise, intermittent contacts, etc. Suppose, first, that only hum is present. In this case, if the modulation coefficient of the leakage carrier is m_a, the power in this modulation referred to the receiver input (which may be called P_n) is

$P_n = \gamma P_t m_a^2$
P_t = the transmitter power
γP_t = the leakage power into the receiver.

The requirement that this be smaller than receiver noise can be expressed as

$$\gamma P_t m_a^2 < 2NFkTF_1,$$

which is equivalent to the requirement on m_a of

$$m_a < \sqrt{\frac{2NFkTF_1}{\gamma P_t}}. \tag{16}$$

It may be seen that, the greater the attenuation γ of the leakage signal, the less severe is this requirement. A typical value of γ obtained in practice, with two paraboloidal antennas side by side, is about 10^{-7}. In this case,

[22] See Appendix III.

using the values

$$NF = 20$$
$$F_1 = 1{,}000 \text{ cps}$$
$$\gamma = 10^{-7}$$
$$P_t = 10^2 \text{ watts,}$$

yields a value of m_a required of

$$m_a < 4 \times 10^{-6}. \tag{17}$$

This is an extremely difficult requirement to meet and necessitates extreme care in eliminating hum and microphonics. Voltage-regulated power supplies, shock-mounting, and acoustic shielding are needed. Care must be taken with the cooling of the transmitter output tube to prevent an impinging air or water blast from introducing microphonics in the output.

If, instead of a single hum frequency, the leakage signal is modulated by a noise spectrum, the modulation can no longer be described by a single modulation coefficient unless a bandwidth is specified in which the noise-modulation sidebands are to be counted. If all the modulation sidebands in a bandwidth equal to the receiver bandwidth (F_1 cps on each side of the carrier) are counted in defining m_a for noise modulation, then (17) applies directly to this case. This will be assumed to be true in what follows.

Phase modulation of the leakage carrier will not produce any signal in the audio amplifier directly, since the detector is an amplitude-modulation detector. It can, however, cause trouble in several ways. In the first place, if there are ground targets present, the signal returned from them will beat with the leakage signal in the detector. Even if the transmitter output is phase-modulated, this beat would still produce only dc except for the fact that the fixed-target signal is slightly delayed because of the distance it has traveled, so that the phase modulations of the leakage signal and fixed-target signal are slightly out of step and their instantaneous frequencies differ slightly. This results in a signal in the audio amplifier whose strength increases linearly with the delay time of the fixed-target signal for close-by targets. There is a delay time for which the effect is a maximum and, at this point, it is just twice as bad as the amplitude-modulation case. That is, for the case of phase modulation by a sine wave, where the peak phase deviation is $\Delta\phi$, the requirement must be met that

$$\Delta\phi < \frac{1}{2}\sqrt{\frac{2NFkTF_1}{\beta P_t}} \tag{18}$$

where βP_t now represents the power in the ground-target signal. The theoretical value of βP_t is given by (13), so that (13) and (18) can be combined to give an expression for $\Delta\phi$. This is modified by the fact that actual fixed-target signals have been found to be approximately 10 db smaller than indicated by (13). Using the values of the previous examples and, in addition, the value $r = 7$ miles gives a value for $\Delta\phi$ of

$$\Delta\phi < 0.4 \times 10^{-5} \text{ radians.} \tag{19}$$

If there are no fixed targets, there will be no requirement on $\Delta\phi$ unless it is desired to make an accurate velocity measurement. Phase modulation can cause an error in this case. To see this, suppose the transmitter output to be phase-modulated as indicated by the expression

$$e_t = E_t \sin(2\pi f_0 t + \Delta\phi \sin 2\pi F t).$$

In this case, the doppler frequency due to the moving target will have some frequency modulation superimposed. Suppose that the doppler frequency is measured by counting the cycles during the time of measurement. The result of the measurement will probably not be the true doppler frequency but something slightly different. The modulation will produce a maximum error in the velocity given by

$$\frac{\Delta v}{v} = \frac{2\Delta\phi}{\pi f_d \tau} \tag{20}$$

where f_d is the doppler frequency involved. This can be rearranged in terms of $\Delta\phi$ as

$$\Delta\phi = \left(\frac{\Delta v}{v}\right)\frac{\pi f_d \tau}{2}. \tag{21}$$

Suppose, for illustration, that values are taken of

$$\frac{\Delta v}{v} = 10^{-3} \text{ or } 0.1 \text{ per cent}$$
$$f_d = 10^3 \text{ cps}$$
$$\tau = 10^{-2} \text{ seconds.}$$

This gives a value of $\Delta\phi$ permitted of

$$\Delta\phi < 1.5 \times 10^{-2} \text{ radians.} \tag{22}$$

By comparing this to (19), it is seen to be a much less stringent requirement.

All of the foregoing results can be applied to the system of Fig. 11 by making two changes. First, the leakage signal of (16) may be replaced by the injected reference signal, normally about as great as the largest fixed-target signal expected. Second, the bandwidth $2F_1$ should be replaced by the effective system bandwidth, which can be taken as $\sqrt{F_1F_2}$. In (18), β should be multiplied by $(r/r_0)^2$, as was indicated in the discussion leading to (14).

Appendix I

If there is one moving-target signal present in the receiver, the output of the detector (into which a coherent reference signal is injected) can be written as

$$e = \sum_{i=1}^{\infty} a_i \sin\left(\frac{2\pi i t}{\tau} + \phi_i\right)$$

$$+ E_r \sin 2\pi f_d t \left\{ n + \sum_{l=1}^{\infty} \frac{2 \sin ln\pi}{l\pi} \cos 2\pi F_1 l t \right\}$$

$n = r/2r_0$ for $r < r_0$

$2r_0$ = maximum unambiguous range.

The first term represents the receiver noise voltage in a Fourier series with base period τ, and the second term represents the signal which is a doppler-frequency component multiplied by a rectangular wave due to the combined transmitter and receiver square-wave keying. The keying is not explicitly stated for the noise voltage because the correlation which it introduces for the a_j and ϕ_j does not affect the final result for the system S/N ratio. The only effect of the receiver keying on the noise is to reduce the noise power per cycle of bandwidth by a factor of 2. Assume that the detector is followed by a low-pass filter passing frequencies in the range $0-F_1$. Then the voltage becomes

$$e = \sum_{j=1}^{F_1\tau} a_j \sin\left(\frac{2\pi jt}{\tau} + \phi_j\right) + E_r n \sin 2\pi f_d t$$

$$+ E_r \frac{\sin n\pi}{\pi} \sin 2\pi(F_1 - f_d)t.$$

The action of the full-wave rectifier can be approximated by assuming a square-law characteristic. That is,

$e_{out} \sim e_{in}^2$

$$e_{out} \sim E_r^2 n \frac{\sin n\pi}{\pi} \cos 2\pi F_1 t$$

$$+ \sum\sum_{l=j=1}^{F_1\tau} a_l a_j \sin\left(\frac{2\pi lt}{\tau} + \phi_l\right) \sin\left(\frac{2\pi jt}{\tau} + \phi_j\right)$$

$$+ 2 \sum_{j=1}^{F_1\tau} a_j \sin\left(\frac{2\pi jt}{\tau} + \phi_j\right) \left\{ E_r n \sin 2\pi f_d t + E_r \frac{\sin n\pi}{\pi} \sin 2\pi(F_1 - f_d)t \right\}$$

+ other signal terms at frequencies different from F_1.

This voltage is now passed through a narrow-band filter of bandwidth F_2 centered at frequency F_1. The signal voltage comes out unchanged as

$$e_{signal} \sim E_r^2 \frac{n \sin n\pi}{\pi} \cos 2\pi F_1 t.$$

To compute the noise voltage, the indicated multiplication must be performed and the terms of frequencies falling within the bandwidth of the narrow-band filter must be kept. If this is done and the mean-square noise voltage is determined, the result is

$$\overline{e_n^2} = F_2\left(\frac{(\overline{a_j^2})^2 F_1 \tau^2}{4} + n^2 \frac{E_r^2 \overline{a_j^2}\tau}{2} + \frac{\sin^2 n\pi}{2\pi^2} E_r^2 \overline{a_j^2}\tau\right),$$

so that the output S/N ratio in power becomes

$$S/N = \frac{E_r^2 n^2 \sin^2 n\pi}{\overline{a_j^2}\tau 2\pi^2 F_2 \left\{ \frac{\overline{a_j^2} F_1 \tau}{4E_r^2} + \frac{1}{2}\left(n^2 + \frac{\sin^2 n\pi}{\pi^2}\right)\right\}}.$$

The quantity $\overline{a_j^2}$ can be related to receiver noise power as follows:

$$\frac{E_r^2}{\overline{a_j^2}\tau} = \frac{P_r}{P_n/\text{cycle}} = \frac{P_r}{NFkT}.$$

This yields

$$S/N = \frac{P_r n^2 \sin^2 n\pi}{2\pi^2 F_2 NFkT \left\{ \frac{1}{2}\left(n^2 + \frac{\sin^2 n\pi}{\pi^2}\right) + \frac{F_1}{4} \frac{NFkT}{P_r} \right\}}.$$

Now, let $n = r/2r_0$, and approximate $(\sin n\pi)/\pi$ by $r/2r_0$. Then,

$$S/N = \frac{F_1/F_2 P_r (r/2r_0)^2}{(2F_1 NFkT)\left\{1 + \frac{1}{8}\left(\frac{2F_1 NFkT}{P_r}\right)\left(\frac{2r_0}{r}\right)^2\right\}}.$$

The peak received power Pr is given by

$$Pr = \frac{P_t \Sigma \pi d^4 \eta^2}{32\lambda^2 r^4};$$

hence,

$$\frac{Pr}{2F_1 NFkT} = \left(\frac{\pi}{64kT} \frac{P_t \Sigma d^4 \eta^2}{F_1 NF\lambda^2 r^4}\right).$$

Thus,

$$S/N = \frac{F_1\left(\frac{\pi}{64kT} \frac{P_t \Sigma d^4\eta^2}{2F_1 NF\lambda^2 r^4}\right)\left(\frac{r}{r_0}\right)^2}{2F_2\left(1 + \frac{1}{4}\left(\frac{r_0}{r}\right)^2\left(\frac{\pi}{64kT} \frac{P_t \Sigma d^4\eta^2}{2F_1 NF\lambda^2 r^4}\right)^{-1}\right)}.$$

Appendix II

Let the output of the narrow-band filter be written as a signal and noise in the form:

$$e = S \sin 2\pi F t + X(t) \sin 2\pi F t + Y(t) \cos 2\pi F t$$

where X and Y have the joint probability distribution

$$P(X, Y) dX dY = \frac{e^{-(X^2+Y^2)/2\sigma^2}}{2\pi\sigma^2} dX dY$$

and σ^2 is the mean-square noise voltage. X and Y obey the relations:

$$\overline{X^2} = \overline{Y^2} = \sigma^2 \overline{XY} = 0.$$

The combined probability distribution is, thus,

$$P\,(\text{signal, noise}) = \frac{e^{-\{(X-S)^2+Y^2)/2\sigma^2\}}}{2\pi\sigma^2}.$$

Let

$$X = r\cos\theta$$
$$Y = r\sin\theta.$$

This makes r the magnitude and θ the phase of the resultant vector. Then,

$$P(r,\theta)drd\theta = \frac{rdrd\theta}{2\pi\sigma^2} e^{-(r^2+s^2-2rs\cos\theta)/2\sigma^2}.$$

From which it follows that

$$P(\theta)d\theta = d\theta \int_0^\infty P(r,\theta)dr$$

Hence,

$$P(\theta) = \frac{e^{-S^2/2\sigma^2}}{2\pi\sigma^2}\int_0^\infty rdre^{-(r^2-2rS\cos\theta)/2\sigma^2}$$

$$= \frac{1}{2\pi}e^{-S^2/2\sigma^2} + \frac{1}{2}\sqrt{\frac{S^2}{2\pi\sigma^2}}\cos\theta \;.$$

$$\cdot e^{-S^2\sin^2\theta/2\sigma^2}\left\{1 + \operatorname{erf}\sqrt{\frac{S^2}{2\sigma^2}}\cos\theta\right\}.$$

It can be recognized that $S^2/2\sigma^2$ is the S/N ratio in power. This derivation took the phase angle of the signal to be zero. If, instead, it was taken as θ_0, the expression in the text would result.

Appendix III

To see that the effective bandwidth is approximately the geometric mean bandwidth, some of the analysis of Appendix I may be used. Starting with the expression,

$$S/N = \frac{F_1}{F_2}\frac{(Pr/2F_1NFkT)(r/2r_0)^2}{\left\{1+\frac{1}{8}\left(\frac{2r_0}{r}\right)^2\left(\frac{2F_1NFkT}{Pr}\right)\right\}},$$

let

$$\frac{Pr}{2F_1NFkT}\left(\frac{r}{2r_0}\right)^2 = x \quad \text{and set} \quad S/N = 1.$$

Then,

$$1 = \frac{F_1X}{F_2\left(1+\frac{1}{8x}\right)}.$$

Solving for x, assuming $F_1/F_2 \gg 1$, yields

$$x = \sqrt{\frac{F_2}{8F_1}} = \frac{Pr}{2F_1NFkT}\left(\frac{r}{2r_0}\right)^2,$$

so that

$$Pr = NFkT\sqrt{\frac{F_1F_2}{2}}\left(\frac{2r_0}{r}\right)^2.$$

Recognizing that Pr is a peak value of the received power, it is to be seen that the average received power is

$$P = NFkT\sqrt{2F_1F_2}\left(\frac{r_0}{r}\right)^2.$$

Since the term (r_0/r) was introduced separately in the text, the "effective" system bandwidth is seen to be $\sqrt{2F_1F_2}$, or, very closely, the geometric mean bandwidth.

Signal-to-Noise-Ratio Improvement in a PCM System*

A. G. CLAVIER†, FELLOW, IRE, P. F. PANTER†, SENIOR MEMBER, IRE, AND W. DITE†

***Summary*—It is shown for a PCM system that the output signal-to-noise power ratio expressed in decibels is approximately equal to twice the input signal-to-noise power ratio. It is independent of the number of code pulses when a sufficient number are used. The distortion due to quantization varies greatly with the number of levels. Adopting the criterion that the output noise power should equal the distortion power, a relation giving corresponding values of the number of digits and input signal-to-noise ratio is found.**

I. Introduction

One of the newer methods of transmitting intelligence by means of pulses is known as pulse-count or pulse-code modulation (PCM). Several[1-3] papers have discussed the advantages of PCM from the point of view of distortion and crosstalk. The purpose of this paper is to discuss another significant aspect of any communication system, which is the system signal-to-noise improvement.

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† Federal Telecommunication Laboratories, Inc., Nutley, N. J.

[1] A. G. Clavier, P. F. Panter, and D. D. Grieg, "PCM distortion analysis," *Trans. AIEE* (*Elec. Eng.*, November, 1947), vol. 66, pp. 1110–1122; November, 1947.

[2] H. S. Black and J. O. Edson, "PCM equipment," *Trans. AIEE*, (*Elec. Eng.*, November, 1947) vol. 66, pp. 1123–1124; November, 1947.

[3] D. D. Grieg, "Pulse count modulation system," *Teletech*, vol. 6, pp. 48–52; September, 1947.

II. General Consideration of Noise

Noise may be divided into two general types; namely, fluctuation noise and impulse noise. In what follows, however, only fluctuation noise will be considered. Fluctuation noise is assumed to be composed of an infinite number of equal infinitesimal components covering the whole frequency spectrum, and assumed to have a random and continually varying phase.

If fluctuation noise is observed for a sufficiently long time, it will be found that any given voltage will be exceeded for a certain fraction of the time of observation. The differential probability that the instantaneous noise voltage lies between V and $V+dV$ is

$$dp = \frac{1}{\sqrt{2\pi e}}\epsilon^{-V^2/2e^2}dV = \frac{1}{\sqrt{2\pi e}}\epsilon^{-r^2/2}dV \qquad (1)$$

where e is the rms value of noise voltage and $r = V/e$.

This is often expressed by saying that the amplitude of the noise voltage is distributed normally about the mean value.

If the circuit includes an ideal low-pass filter, with a cutoff frequency f_0, it has been shown by Rice[4] that the probable number of positive noise bursts per second of amplitude greater than V is

$$n(V) = \frac{f_0}{\sqrt{3}} \epsilon^{-r^2/2}. \tag{2}$$

There is an equal probable number of negative bursts exceeding V in amplitude.

An approximate derivation of (2) is given in the Appendix.

The noise bursts will be considered in what follows as being independent of each other. Further, the number of noise bursts in a given interval of time will be considered as independent from what occurs in any other interval. Of course, over a sufficiently great period of time, the average number $n(V)$ of noise bursts per second above V is determined. Accordingly, we may assume that the distribution of noise bursts obeys Poisson's law, which states that the probability of obtaining exactly k disturbances in an interval of time T is

$$p(k) = \frac{(nT)^k}{k!} \epsilon^{-nT}. \tag{3}$$

As shown in Fig. 1, the curve of $p(k)$ versus k assumes a variety of forms according to the value of nT. If nT is small—say, 0.1—then $p(o)$ is nearly unity and $p(k)$ decreases rapidly as k increases. When nT is large—say, 100—the curve is sharply peaked at $k = 100$.

III. Application to PCM

In order to evaluate the effect of noise in the video section of the receiver, it will be recalled that in a binary PCM system intelligence is conveyed by a given number μ of code pulses, giving in all $2^\mu - 1$ discrete amplitude levels for the output PAM pulses. These PAM pulses are subsequently converted to audio by means of a low-pass filter. If the amplitude of the code pulses is taken as $2V$, the following simplifying assumptions will be made as to the effect of noise bursts:

1. A negative noise burst of any shape or duration, and of amplitude greater than V, when combined with a code pulse, will cause complete obliteration of the code pulse.

2. A positive noise burst of any shape or duration, and of amplitude greater than V, occurring in the absence of a code pulse, will cause a spurious code pulse to appear.

These assumptions are justified by (26) of the Appendix, which shows that the average duration of a noise burst above V is small compared with the width of the PAM pulse, provided the cutoff frequency f_0 of the filter is chosen as low as the duration τ of the code pulse will allow (that is to say, $1/2\tau$), and the input signal-to-noise ratio is sufficiently large. Hence, a noise burst will cause the insertion or cancellation of only one code pulse, except for very rare cases which may be neglected. Of course, when noise power is comparable to the signal power this is no longer true, but the system would then become inoperative.

For the time being, suppose that only one noise burst occurs during the time T allotted for the transmission of the μ code pulses. In what follows, the interval T will be referred to as the subframe. As a noise burst may occur in any of the μ positions, the mean square change in the level of the output PAM pulses is

$$M_1 = \frac{1}{\mu} \sum_1^\mu 2^{2(\mu-1)} = \frac{4^\mu - 1}{3\mu}. \tag{4}$$

This is expressed in terms of the square of the value of one level taken as an arbitrary unit, and consequently has the dimension of a voltage squared.

Taking account of the fact that on the average there is only half of the noise bursts (positive or negative) which effectively alter the code, the other half being composed of those positive bursts which fall simultaneously with a code pulse, and those negative noise bursts which occur when there is no pulse, the probability of obtaining one effective noise burst in a subframe is $p(1)$, by (3), and the output noise power (for a unit resistor) of the output PAM pulses is, therefore,

$$N_1 = \frac{4^\mu - 1}{3\mu} \frac{\tau}{T} p(1) \tag{5}$$

where τ is the duration of an output PAM pulse, and (1) indicates that there is only one noise burst per subframe.

Provided the input signal-to-noise ratio is sufficiently large, the quantity nT in Poisson's equation (3) will be small compared to unity, and, as previously mentioned, the probability $p(k)$ decreases rapidly as k increases. The case of two noise bursts in a subframe is thus al-

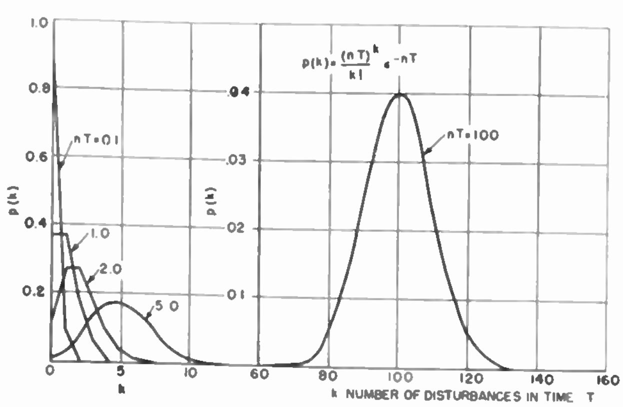

Fig. 1—Curves of Poisson's law.

[4] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. Jour.*, vol. 23 pp. 282–332; July, 1944.

ready very much less probable than the case of only one burst, for all practical cases of PCM transmission systems. An approximate calculation of the contribution of this case to the output noise is obtained as follows. Let two indices (i, j) designate the position of the modified code pulses. The amplitude of the output pulse is modified by the burst (i, j) in any of the following ways:

$$2^i + 2^j,\ 2^i - 2^j,\ -2^i + 2^j,\ -2^i - 2^j,$$

depending on the particular subframe being considered. Assuming that any of these combinations is equally probable, the mean square variation of the corresponding output PAM pulses is

$$m_{i,j} = \tfrac{1}{4}\left[(2^i + 2^j)^2 + (2^i - 2^j)^2 + (-2^i + 2^j)^2 + (-2^i - 2^j)^2\right] = 4^i + 4^j.$$

It is now necessary to sum $m_{i,j}$ over all possible values of (i, j), noting, however, that the burst (i, i) is not more damaging than the single burst i. The result is

$$M_2 = \sum_{i=1}^{i=\mu} \sum_{j=1}^{j=\mu} m_{i,j} = \frac{2\mu - 1}{\mu} M_1. \qquad (6)$$

If μ is large, this is very close to

$$M_2 = 2M_1. \qquad (7)$$

A similar reasoning may be applied to three or more bursts per frame; in fact, as long as the number of bursts per frame does not become equal to the number of code pulses. As the probability for this to occur is negligible if the signal-to-noise ratio is large enough for the system to be operative, the output noise power (for a unit resistor) is closely approximated by the following expression:

$$N_0 = \frac{4^\mu - 1}{3\mu} \frac{\tau}{T} \left[p(1) + 2p(2) + 3p(3) + \cdots\right] = \frac{4^\mu - 1}{3} \frac{n(V)\tau}{\mu}. \qquad (8)$$

The number of noise bursts $n(V)$ is given by (2), so that

$$N_0 = \frac{4^\mu - 1}{3\sqrt{3}} \frac{\tau}{T} \left(\frac{f_0 T}{\mu}\right) \epsilon^{-r^2/2}. \qquad (9)$$

The output signal power is determined by the amplitude and form factor of the quantized signal. For a sinusoid of peak value $m/2(2^\mu - 1)$ where m is the degree of modulation, this output signal power (in a unit resistor) is[1]

$$S_0 = \frac{1}{8} m^2 \frac{\tau}{T} (2^\mu - 1)^2. \qquad (10)$$

The output signal-to-noise ratio is

$$(S/N)_0 = \frac{3\sqrt{3}\, m^2}{8} \cdot \frac{2^\mu - 1}{2^\mu + 1} \left(\frac{\mu}{f_0 T}\right) \epsilon^{r^2/2}. \qquad (11)$$

Noting that T/μ is the duration of a code pulse, we see that $f_0 T/\mu$ is a significant parameter of a pulse system which determines the shape of the pulse. An acceptable minimum value for this quantity is 0.5. We may write further that

$$r^2 = \frac{V^2}{e^2} = (S/N)_I \qquad (12)$$

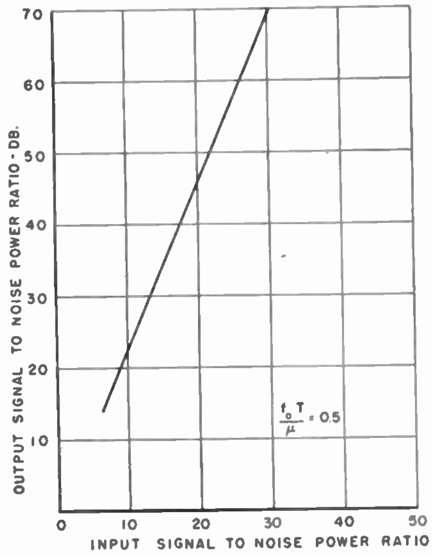

Fig. 2—Output signal-to-noise ratio versus input signal-to-noise ratio.

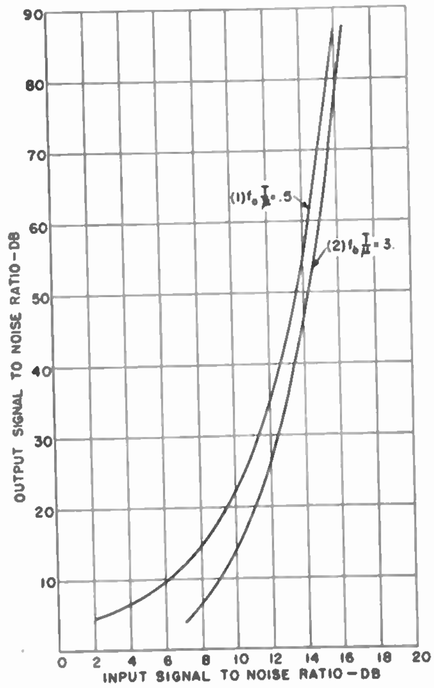

Fig. 3—Signal-to-noise improvement in a PCM system (S/N) output db versus (S/N) input db.

where $(S/N)_I$ is the input signal-to-noise ratio. Thus the final result is

$$(S/N)_0 \simeq 1.3 m^2 \epsilon^{1/2(S/N)_I} \qquad (13)$$

where both $(S/N)_I$ and $(S/N)_0$ are expressed as power ratios. Letting $m=1$ and expressing $(S/N)_0$ in decibels gives the relation

$$(S/N)_{0_{db}} \simeq 2.2(S/N)_{I_{power}}. \tag{14}$$

Thus the output signal-to-noise power ratio expressed in decibels is proportional to the input signal-to-noise ratio expressed as a power ratio. This result is independent of the number of code pulses, provided enough are used; that is, more than 3 or 4. The results of (14) are shown graphically in Figs. 2 and 3.

In any PCM transmission system, although the effect of noise is substantially independent of the number of code digits, the per cent distortion is, of course, directly influenced by it. The distortion power of the output pulses is given by[1]

$$D = \frac{A^2}{12N^2}\frac{\tau}{T} = \frac{1}{12}\frac{\tau}{T} \tag{15}$$

where

$2A$ = peak-to-peak value of signal $= 2^\mu - 1$

$2N$ = total number of levels $= 2^\mu - 1$.

Taking $f_0T/\mu = 0.5$, the output noise power is, from (9),

$$N_0 = \frac{4^\mu - 1}{6\sqrt{3}}\frac{\tau}{T}\epsilon^{-1/2(S/N)_I}. \tag{16}$$

Thus,

$$\frac{D}{N_0} = \frac{\sqrt{3}}{2}\frac{1}{4^\mu - 1}\epsilon^{1/2(S/N)_I} \simeq \frac{1}{4^\mu}\epsilon^{1/2(S/N)_I}. \tag{17}$$

The output noise power will thus be equal to the distortion power when

$$\epsilon^{1/2(S/N)_I} = 4^\mu.$$

A curve of $(S/N)_0$ in decibels versus μ is plotted in Fig. 4 under these conditions. It gives corresponding values of output signal-to-noise ratio and number of code pulses for equality of the distortion and output noise power.

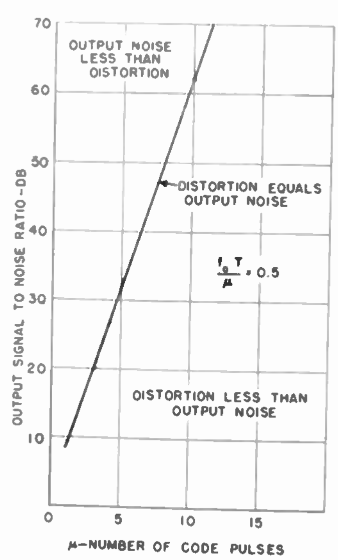

Fig. 4—Output signal-to-noise ratio versus number of code pulses when distortion equals noise.

This criterion is useful in the design of cable systems where the signal power is constant. It gives the highest fidelity consistent with other factors. For instance, about 11 digits are required for an output signal-to-noise and distortion ratio of 60 db. The improvement secured by a further increase in the number of digits would be small and not warranted by the added complexity.

IV. Effect of Noise on PCM Systems with Repeating Stations

Let the relay system in Fig. 5 consist of k identical linear amplifiers and k identical paths, the gain of an amplifier being just sufficient to overcome the path at-

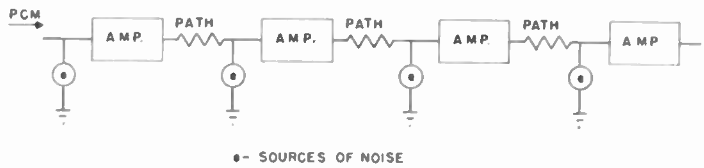

Fig. 5—Relay system.

tenuation. The various noise sources of rms value e may then be replaced by an equivalent noise source of rms value $\sqrt{k}e$ at the input. Equation (14) gives

$$(S/N)_{0_{db}} = \frac{2.2}{k}(S/N)_I \tag{18}$$

where $(S/N)_I$ is the input signal-to-noise power ratio as defined previously; the output signal-to-noise ratio again is expressed in db. In practice, the results obtained may be somewhat worse due to the progressive deterioration of the pulse shape.

In another type of relay system, the repeaters regenerate the code pulses so that the output consists of idealized PCM pulses. The gain, again, is just sufficient to overcome the path attenuation. As noise enters in the form of wrong code pulses, each noise source contributes n noise bursts, and consequently n wrong codes independently of the others. If we neglect the possibility of some noise bursts canceling the effect of others, (9) for the output noise power becomes

$$N_0 = k\frac{4^\mu - 1}{3\sqrt{3}}\frac{\tau}{T}\left(\frac{f_0T}{\mu}\right)\epsilon^{-r^2/2}.$$

Using this value, (14) becomes

$$(S/N)_{0_{db}} = 2.2(S/N)_I - 10 \log k. \tag{19}$$

From this we see that the curve of Fig. 2 is moved parallel to itself, whereas in the first case considered its slope is changed.

To show the effect more clearly, the input signal-to-noise ratio for an assumed output signal-to-noise ratio of 60 db versus the number of repeaters is plotted in Fig. 6. It is seen that, for the linear repeaters, the input level rises considerably with the number of repeaters, and requires a 17-db increase for 50 repeaters. On the other hand, for regenerative repeaters a small increase of 1 db takes care of the cumulative effect of noise. This

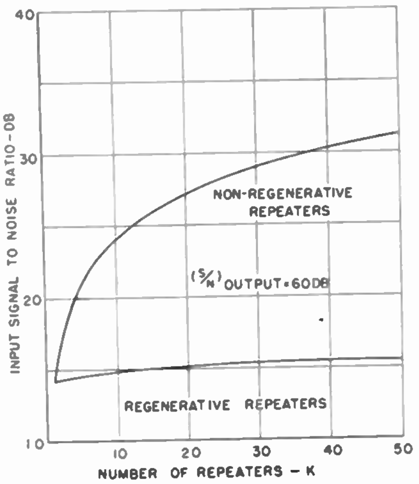

Fig. 6—Input signal-to-noise ratio versus number of repeaters for constant output signal-to-noise ratio.

represents a considerable economy in total power installed, since each repeater power must necessarily be increased. Systems combining the two types may be used, thereby obtaining intermediate results.

V. Conclusions

A study of the effect of fluctuating noise on PCM signals shows that the output signal-to-noise power ratio expressed in decibels is proportional to the input signal-to-noise power ratio. It is substantially independent of the number of code digits when a sufficient number is used, superior to, say, 3. The distortion due to quantization, however, varies greatly with the number of levels, and consequently the number of code digits. A relation is given showing the number of digits for which the output noise power is equal to the distortion power, when a definite input or output signal-to-noise ratio is assumed. It is, of course, unnecessary to increase the number of digits beyond that value for which these two powers are equal.

In case the communication link includes a number of relays, PCM shows a striking advantage over other modulation systems when use is made of regenerative repeaters. A very small increase in input signal-to-noise ratio accounts for the cumulative effect of noise along the whole chain of repeaters.

Acknowledgment

Acknowledgment is made to G. Deschamps for his mathematical assistance in the preparation of this paper.

VI. Appendix

An approximate derivation of (2) will be given here. If wide-band fluctuation noise is passed into an ideal low-pass filter, the large peaks on the input side may be considered as being essentially isolated impulses. Thus, the output noise peaks are determined by the impulse response of the filter, which is

$$v(t) = V_p \frac{\sin 2\pi f_0 t}{2\pi f_0 t} \tag{20}$$

where V_p is the peak voltage and f_0 the cutoff frequency. From the definition of probability $p(V)$, we have

$$p(V) = n(V)\cdot t(V) \tag{21}$$

where $n(V)$ is the number of peaks above V, and $t(V)$ is the average duration of such peaks.

On the other hand, the expression of $p(V)$ is

$$p(V) = \frac{1}{\sqrt{2\pi}\, e}\int_V^\infty \epsilon^{-V^2/2e^2} dV.$$

Utilizing an asymptotic series for the integral, it is thus found by limiting the expansion to two terms that $p(V)$ is expressed by

$$p(V) = \frac{1}{\sqrt{2\pi}}\, \frac{e}{V}\left(1 - \frac{e^2}{V^2}\right)\epsilon^{-V^2/2e^2} dV. \tag{22}$$

The average duration $t(V)$ may be found by replacing the noise peaks above V by an average noise peak of height $\overline{V}$ given by

$$\overline{V} = \frac{\dfrac{1}{\sqrt{2\pi}\, e}\displaystyle\int_V^\infty V\epsilon^{-V^2/2e^2} dV}{\dfrac{1}{\sqrt{2\pi}\, e}\displaystyle\int_V^\infty \epsilon^{-V^2/2e^2} dV} \tag{23}$$

To the approximation of (22), this gives

$$\overline{V} = \frac{V}{1 - \dfrac{e^2}{V^2}} \simeq V\left(1 + \frac{e^2}{V^2}\right). \tag{24}$$

The average duration $t(V)$ is then found by solving

$$V = V\left(1 + \frac{e^2}{V^2}\right)\frac{\sin \pi f_0 t(V)}{\pi f_0 t(V)}. \tag{25}$$

Assuming $\pi f_0 t(V)$ is a small angle, the result is

$$t(V) = \frac{\sqrt{6}}{\pi f_0}\, \frac{e}{V}\left(1 - \frac{e^2}{2V^2}\right). \tag{26}$$

The average number of noise peaks is then derived from (21), and is found to be

$$n(V) = \frac{\sqrt{\pi}}{2}\, \frac{f_0}{\sqrt{3}}\left(1 - \frac{e^2}{2V^2}\right)\epsilon^{-V^2/2e^2}. \tag{27}$$

This is very nearly the result given by Rice and utilized in the text.

Temperature Variations of Ground-Wave Signal Intensity at Standard Broadcast Frequencies*

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Summary—Variations of ground-wave signal intensity at standard broadcast frequencies appear to be more closely related to changes in temperature than to any other single commonly observed meteorological measurement. Results presented here were obtained from an analysis of signal intensities and weather conditions over six paths between 30° and 45° north latitude. It was found that (1) the intensities tend to decrease markedly from their peak values when the temperature becomes high, (2) the amount of such decrease is approximately proportional to the path length in wavelengths, and (3) the temperature at which the peak value occurs varies with frequency. Sample curves of experimental data are presented. General relationships deduced from all paths are combined in a nomographic chart showing intensities relative to the peak value for various frequencies, path lengths, and temperatures.

I. Introduction

Ground-wave signal intensities at standard broadcast frequencies exhibit considerable variation, even over short paths. The shortest of the six paths reported here showed, for 475 days of recording, a mean intensity of 289 microvolts per meter, with a maximum of 470 and a minimum of 175, representing a ratio of 2.7 to 1. The longest path showed, for 504 days of recording, a mean intensity of 5.3 microvolts per meter, with a maximum of 26 and a minimum of 1.3, representing a ratio of 20 to 1.

The practical result of such variations, regardless of the physical explanations ultimately involved, is equivalent to changes in the effective conductivity of the path. Therefore, in cases where the consideration of close margins in conductivity measurements is necessary, the variations produced by changing conditions over the path may become exceedingly important.

In a preliminary investigation on the analysis of signal intensities and weather conditions over the Philadelphia-to-Baltimore path, the effects of various weather conditions were studied at some length.

Careful comparisons of signal intensities with many sets of data on temperature, precipitation, humidity, atmospheric pressure, dew point, and vapor pressure showed clearly that there was closer and more continuous correlation with temperature than with any other single type of data.

Fig. 1, in giving all of the signal intensities for one of the paths, shows the type of distribution obtained for the data in general. There are indications of both the range of variation within a month and the change in general level from month to month. It will be seen later that much of both of these variations can be explained in terms of temperatures.

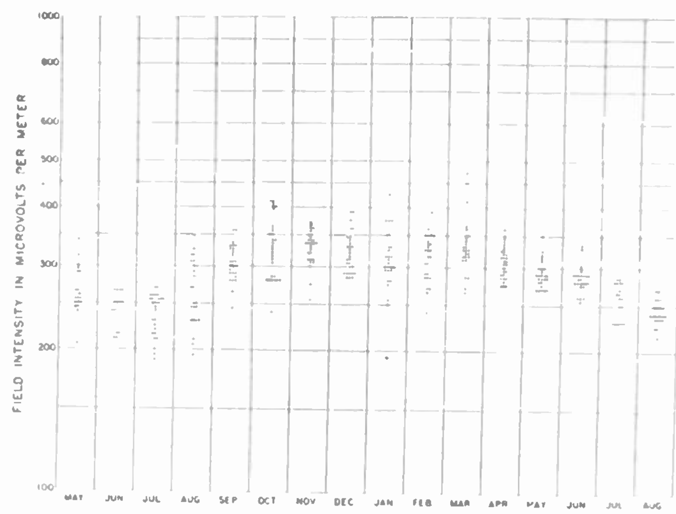

Fig. 1—Daily 1:30 P.M., EST, field intensities, Philadelphia, Pa., (WCAU) to Baltimore, Md. 1,170 kc, 76 miles. May, 1939–August, 1940.

Fig. 2 gives an example of one of the better point-by-point correlations of signal intensities with temperatures. It is a comparison of the three-day running averages of the signal intensities of radio station WCAU, Philadelphia, received at Baltimore, with the three-day running averages of the mean of the Philadelphia and Baltimore 24-hour mean temperatures.

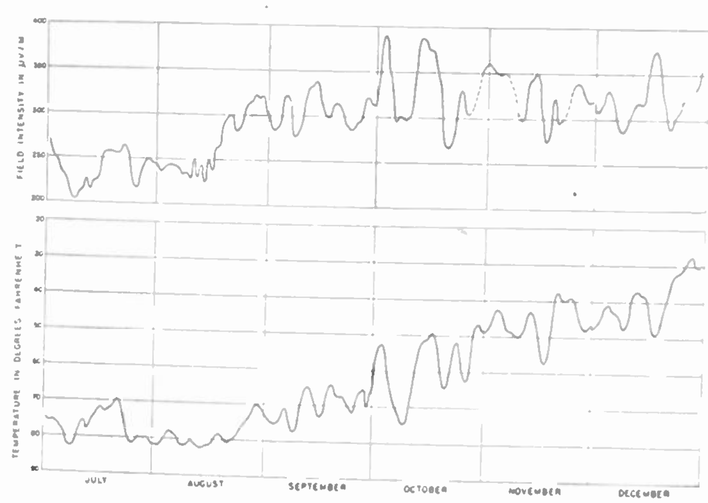

Fig. 2—Comparison of three-day running averages, Philadelphia-to-Baltimore signal intensities and temperatures.

Probably the most promising data, next to temperatures, are those on precipitation. On most of the paths there were frequent instances of marked increases in signal intensity during, and for a few days following, periods of heavy rainfall. However, there were other pe-

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riods, some of them with equally heavy rainfall, when no such increase occurred.

This partial correlation, and particularly the lag of the effect after the end of the rainfall, led to an investigation of a ground-moisture figure. Such a figure attempts to make allowances for the rates at which the ground gains moisture by precipitation and loses it by runoff and evaporation.

The application of this relative moisture figure to a portion of the Philadelphia-to-Baltimore data resulted in another partial correlation, which, while interesting, was not as continuous and reliable as that for temperature.

Comparisons of intensities with atmospheric pressure, dew point, and vapor pressure were, in general, inconclusive. A few highly localized coincidences with humidity were observed; enough to indicate a partial relationship, but not enough to form a basis for generalized results.

Limitations of the present data on signal intensities are coupled with additional limitations imposed when it is necessary to get simultaneous, detailed measurements of several types of meteorological data over a specific path. Therefore, it is especially fortunate for the purpose of analysis at the present time that signal intensities show a quite close relationship with single, directly obtained types of data such as temperature. The following development is concerned with extracting the maximum amount of information about temperature variations from the available data.

Before discussing the details of such a development, however, it seems desirable to make brief comments on the probable physical explanations of ground-wave variations. Three chief possibilities appear to be (1) the conductivity of the earth and (or) its vegetation, (2) the gradient of the index of refraction of the lower troposphere, and (3) the dielectric constant of the earth.

An extensive treatment of the first of these is contained in a recent paper[1] which discusses the effects of vegetation, particularly of trees, on ground-wave variations. Although it has not been possible to make a detailed study of the vegetation on the paths reported below, the results, so far as the intensity versus temperature relationship is concerned, are in substantial agreement with those given in the paper.

Exceptions appear, however, when an attempt is made to obtain correlations seasonally with vegetation. Signal intensities at the same or similar temperature in various seasons were compared, but no appreciable trend or tendency toward grouping was apparent. Furthermore, there were short periods of high atmospheric temperature on a few winter days when signal intensities corresponding to the same temperatures in summer were closely duplicated, although presumably the vegetation would have been in a condition much different from that of summer.

Two points should be further emphasized in this connection: (1) The survey reported below consists of a large number of observations distributed over a range of frequencies, path locations, and path lengths; and (2) the main bodies of data in the two reports are in substantial agreement concerning the intensity versus temperature relationship.

Another possibility for explaining ground-wave variations appears to be the gradient of the index of refraction of the lower troposphere. Measurements of reflections from regions just a few kilometers above the earth's surface have been reported in detail in papers by Watson Watt, Wilkins, and Bowen,[2] and by Friend and Colwell.[3] The possibility exists that such a reflected wave may combine with the surface wave to produce the variations in the total intensity.

Just as some of the winter-day intensities become difficult to explain in terms of vegetation, so the many regular intensity versus temperature observations become difficult to explain in terms of a reflected wave component which is varying irregularly in height of reflection and length of path.

It has also been suggested as a third possibility that changes in the dielectric constant of the earth caused by changes in its moisture content may be involved. As previously mentioned, however, the correlation with varying moisture is only partial.

Thus, none of the explanations available at present appears to cover all of the observed results, but each is capable of explaining part of them.

II. Data and Method of Analysis

On many field-intensity records at standard broadcast frequencies and at distances on the order of 100 to 400 miles, the rapidly fluctuating sky-wave trace of the dark hours gives way between sunrise and sunset to a comparatively steady trace representing a signal intensity produced wholly by the ground wave. Although this trace is remarkably constant compared to the sky-wave trace, there are gradual, but significant, changes occurring in its general level from hour to hour and from day to day.

First impressions were that the ground-wave signal intensities were nearly constant during any given day, but that the general daily levels varied considerably from day to day. A test of one month's portion of the Philadelphia-to-Baltimore data, however, indicated otherwise. The monthly mean of the daily mean deviations for the period from 9 A.M. to 3 P.M. is 8.5. The monthly mean of the day-to-day deviations of the daily

[1] W. Gerber and A. Werthmuller, "Ueber die vegetabile Absorption der Bodenwelle," *Techn. Mitt., Jg.*, vol. XXIII, No. 1, S. 12; 1945.

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

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

mean signal intensities is 16.6. Since the period during the day is 7 hours and the day-to-day period is 24 hours, the comparison will be 8.5:16.6 = 7:24. Therefore, since the deviations are roughly proportional to their corresponding lengths of period, the indication is strong that no essentially different kind or magnitude of change would be observable *between* days from that observed *during* days.

Considerable testing of the comparison of various lengths of period of signal-intensity measurements with various periods of temperature measurements indicated the following as preferred methods of handling the present data:

(1) Comparison of a spot measurement of signal intensity at a given time each day with the simultaneous temperature of the path as determined by one or more measurements at suitably located weather stations.

(2) Comparison of signal intensity as above with the 24-hour mean temperature obtained at one or more suitably located stations (somewhat less desirable, but quite usable in cases where simultaneous temperature measurements are not available).

The time 1:30 P.M., EST, was chosen as the best compromise between having the measurements near the middle of the day (for assurance of 100 per cent ground wave) and having the maximum amount of simultaneous weather data.

The main details of the data material used are shown in Table I.

TABLE I

Path	Frequency, kc	Path Length in Miles	Path Length in Wave-lengths
(1) Philadelphia (WCAU)—Baltimore	1,170	76	477
(2) New York (WABC)—Baltimore	860	163	752
(3) Ames, Iowa (WOI)—Grand Island, Neb.	640	259	889
(4) Des Moines (WHO)—Grand Island	1,040	268	1,495
(5) Minneapolis (WCCO)—Grand Island	830	390	1,737
(6) Dallas (WFAA)—Grand Island	820	558	2,454

Period of Recording	Interval	Temperature Measurements
(1) May, 1939–Aug., 1940	Daily	Mean of Philadelphia–Baltimore 1:30 P.M. EST
(2) May, 1939–Dec., 1940	Daily	Phila., 1:30 P.M., EST
(3) Jan., 1941–Dec., 1941	Every 4th day	Omaha, 24-hour mean
(4) Jan., 1939–Nov., 1942	Every 4th day	Omaha, 24-hour mean
(5) Jan., 1939–Dec., 1939	Every 4th day	Sioux City, Iowa, 1:30 P.M., EST
(6) May, 1938–Feb., 1942	Every 4th day	Concordia, Kan., 24-hour mean
Mar., 1942–Dec., 1942	Daily	

Plots of the 1:30 P.M., EST, field intensities in microvolts per meter versus temperatures in °F for four of the paths listed above are shown in Fig. 3.

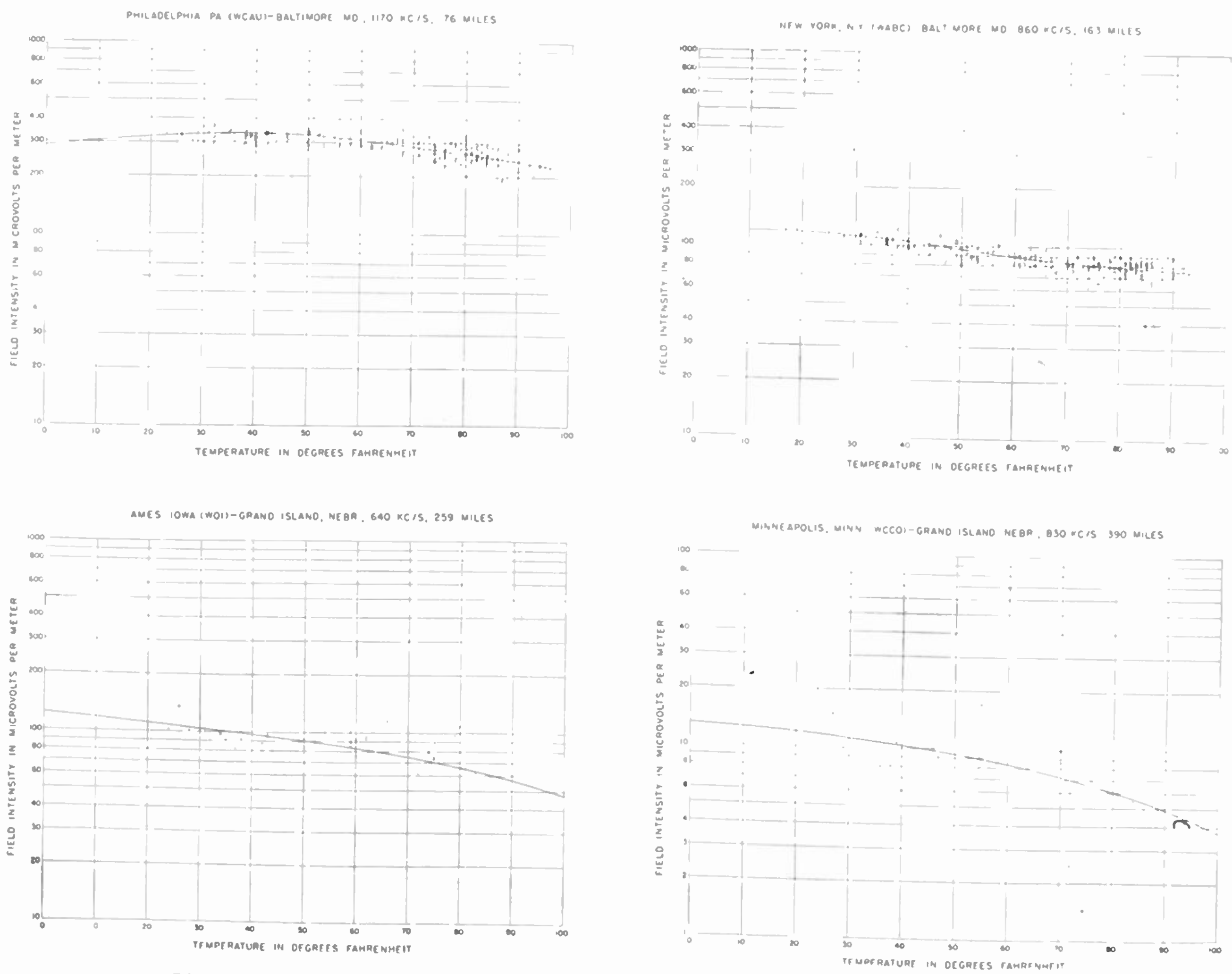

Fig. 3—Samples of signal-intensity versus temperature data for four of the paths measured.

The trend curve shown by the solid line in each was drawn by locating the median levels of segments 10°F wide and drawing a smoothed curve through these levels.

The scatter of individual values from the trend curves increases, in general, with distance. Part of the scatter may be only an apparent one, due to the difficulty of obtaining a close approximation to the mean temperature over the whole of a long path. Whatever the cause, it also happens that, when the scatter begins to get troublesome, the distance limit is also being closely approached, so that the signal intensity begins dropping below the noise level frequently at certain times of the year. The Dallas-to-Grand Island path of 558 miles represents about the maximum distance, and is, indeed, exceptionally long for this type of use.

III. Conclusions

All of the curves obtained from measurements on the six paths, together with values applying to a frequency of 1,500 kc (obtained from the second harmonic of a 750-kc station), were studied for indications of general trends.

The following general observations appear to be established with reasonable certainty:

(1) There is a marked decrease in the intensities at the higher temperatures.

(2) The amount of this decrease increases monotonically with distance.

With less certainty but strongly indicated, in the writer's opinion, by an inspection of all available data are:

(1) A maximum intensity at some intermediate temperature with a slight decrease at lower temperatures.

(2) A continuous shift of the maximum toward lower temperatures for lower frequencies. (The temperature range available does not permit the lower parts of the curves for the lower frequencies to be completely defined.)

(3) For any given frequency, a rate of decrease from the maximum intensity which is approximately proportional to the distance in wavelengths. (The possibility of expressing the distance in miles rather than in wavelengths is not ruled out, but such comparisons as have been possible indicate a slight advantage in favor of wavelengths.)

Numerous attempts were made to express these indicated general relationships in the form of empirical equations. It was planned to derive an expression which would permit the substitution of various combinations of frequency, path length, and temperature as parameters. A fourth-degree polynomial was found to fit considerably better than anything of lower degree. However, the complexity resulting from its use when the proper coefficients were supplied, together with the indication that it would not be enlightening for physical explanations anyway, made other methods advisable.

Preparation of a nomographic chart seemed the only practical method of representing all the general relationships compactly. Fig. 4, therefore, incorporates and summarizes all these results. Estimates of relative signal intensities for various temperatures, frequencies, and distances in wavelengths may be read from the chart by the procedure stated in the lower left of the figure.

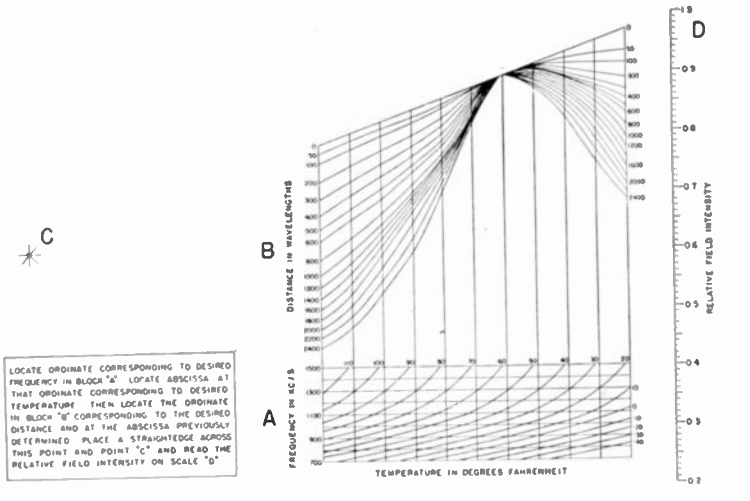

Fig. 4—Nomographic chart suggesting possible relationship of intensities with frequencies, temperatures, and path lengths.

It should be emphasized, perhaps, that Fig. 4 is not intended to represent conclusions established by the data with complete finality, but represents a concise expression of the writer's effort to interpret the data presently available as completely as possible.

IV. Acknowledgments

The collection of the radio data on which this report is based was made possible by the co-operative efforts of the author's colleagues within the engineering department of the Federal Communications Commission. Particular acknowledgment for helpful discussions is made to Edward W. Allen, Jr., and Edgar F. Vandivere, Jr., of the Commission's Technical Information Division.

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4. G. W. Pickard, "Some correlations of radio reception with atmospheric temperature and pressure," Proc. I.R.E., vol. 16, pp. 765–772; June, 1928.
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A Phenomenological Theory of Radar Echoes from Meteors*

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Summary—**The Dominion Observatory and the National Research Council, Ottawa, Canada, have undertaken a program of combined visual, photographic, and radar observations of meteors. The observational data is summarized, and general conclusions have been drawn.**

The radar echoes obtained from meteors have been classified into basic types according to their appearance on the range-time record of the radar display. These types include echoes indicating approach, or recession, or both. Other observed features are: durations of echoes up to several minutes, complexity of structure for echoes from brighter meteors, and appreciable delays in the appearance of the echoes.

A phenomenological theory is proposed, involving a number of postulates concerning the physical conditions in an *M* region in the upper atmosphere. A kinetic-energy mechanism, together with an ultraviolet radiation mechanism, are suggested to account for the ionization produced by the meteor. In the *M* region are visualized striae, or patches, which form a fine structure such that, within the striae, the physical properties of the atmosphere emphasize the creation and maintenance of meteoric ionization as compared to the ionization produced in the intervening spaces. A qualitative explanation for all the observed echo forms is advanced on the above hypothesis. The results of other investigators on different wavelengths are consistent with this analysis.

INTRODUCTION

EVER SINCE the first tentative suggestions were made by Pickard,[1] Skellett,[2,3] and others that short-duration *E*-region radar reflections might be due to meteoric ionization, considerable interest has been aroused in many quarters and a great deal of observational and theoretical work has been done. The result is that, at this date, there is no serious doubt that meteors actually produce transient echoes which are obtained from heights near the *E* region. The evidence is overwhelming: the diurnal and annual echo rates vary with the visual meteor rates; the rates increase during the well-known visual showers; many definite coincidences have been obtained between the stronger echoes and the brighter visual meteors.

Up to the present, however, the literature does not show that the correlation between the visual meteors and the radar echoes has been carried out in as thorough a manner as might be desired. With the object of giving particular attention to this phase of the problem, a combined project was initiated in August, 1947, at Ottawa under the auspices of the Dominion Observatory and the National Research Council. The combined observations have so far been made on the Perseid shower of August, 1947,[4] on the Geminid shower of December, 1947,[5] and on the Lyrid shower of April, 1948.[6] The radar equipment has been operated alone at other times. The meteor-shower periods are selected for the combined work chiefly for the reason that direction of motion of the shower meteors is known. Hence, from visual or photographic observations made at a single point on the earth's surface, not only can the position of a meteor trail[6] in the sky be found but, if that meteor is identified as a shower meteor, the angle between the meteor's path and the line of sight can be determined. Further, the geocentric velocity is also known for most of the annual showers, and the number of meteors in the sky during a shower is considerably greater than on a nonshower night.

In the program on which this paper is based, the visual observations were carried out by a team of six or more observers, and the reduction of their data yielded meteor positions with an average error in the neighborhood of 3°. Direct and spectrographic cameras were also operated, and when photographs of the brighter meteors were available, the error in the observed position was of the order of 0°.05. The timing error of the visual meteors was about ½ to 1 second. The times were recorded both by a recording observer associated with the observing group, and by an Esterline-Angus multipen recorder with individual pens remotely controlled by each observer. During the Lyrid shower, push buttons were connected to an illuminated indicator which marked the radar film directly.

The parameters of the radar system during the Perseid shower were as follows: frequency, 32.7 Mc; peak power, 150 kw; pulse length, 8 microseconds; receiver sensitivity, 5×10^{-14} watt; single half-wave horizontal dipoles on both transmitter and receiver, each mounted a quarter-wave above the ground plane, both dipoles

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[1] G. W. Pickard, "A note on the relation of meteor showers and radio reception," PROC. I.R.E., vol. 19, pp. 1166–1170; July, 1931.

[2] A. M. Skellett, "The effect of meteors on radio transmission through the Kennelly-Heaviside layer," *Phys. Rev.*, vol. 37, p. 1668; June 15, 1931.

[3] A. M. Skellett, "The ionizing effect of meteors in relation to radio propagation," PROC. I.R.E., vol. 20, pp. 1933–1940; December, 1932.

[4] Peter M. Millman, D. W. R. McKinley, and M. S. Burland, "Combined radar, photographic and visual observations of the 1947 Perseid meteor shower," *Nature*, vol. 161, pp. 278–280; February 21, 1948.

[5] Peter M. Millman and D. W. R. McKinley, "A note on four complex meteor radar echoes," *Jour. Roy. A. Soc. Can.*, vol. 42, pp. 121–130; May and June, 1948.

[6] In this paper, we define *meteor path* as the line of motion of a meteor described in a three-dimensional co-ordinate system referred to the earth. *Meteor trail* is the meteor path projected on the celestial sphere, as seen by the observer.

oriented north-south. The power was increased to about 400 kw peak, with a 12-microsecond pulse length, just prior to the Geminid shower, with the other parameters remaining approximately the same. An A-scope display was used for monitoring, and a range-time display on a long-delay-screen cathode-ray tube was employed for recording and timing the radar echoes. In August, the range-time display was continuously photographed with a single-shot movie camera open for 1-minute intervals. Since November, a continuously moving film camera has been employed with a stationary, intensity-modulated range trace. Timing seconds pulses were available from the Dominion Observatory's time service, radio station CHU. Starting in December, these seconds pulses have been applied directly to the range trace. The same pulses were made audible for use by the visual observers. The timing error of the radar echoes was of the order of one-quarter of a second or better, and the combined error of visual and radar timing was thus of the order of one second, on the average.

The experimental data obtained from these showers have been described briefly elsewhere, and more complete accounts are being prepared for publication. However, it was felt that several interesting conclusions could be drawn from this work without recounting the vast amount of small detail involved in making and reducing the observations. This paper summarizes the results and outlines a phenomenological theory of the mechanics of radar reflections from meteors.

Several theories of meteoric ionization and the radar reflections obtained from the meteor trails have been advanced.[7-9] One of the salient points in these discussions is the fact, clearly demonstrated experimentally, that the trail of ionization behaves as a long, thin reflecting cylinder immediately after the passage of the meteor. Thus the most efficient reflection will be obtained from a meteor when its path is normal to the line of sight. Our observations have added further confirmation to the large amount of data verifying this normal reflection law. However, we should like to point out that this law applies only *immediately after* the passage of the meteor, as we believe that the character of the radar echoes and the mechanics of reflection change radically with time and with the orientation of the meteor path with respect to the line of sight. That is, while the bulk of the small meteor echoes observed are probably seen as a consequence of the normal reflection law, the larger meteors definitely produce complicated echoes which are not explained by this law.

[7] E. W. Allen, Jr., "Reflections of very-high-frequency radio waves from meteoric ionization," Proc. I. R. E., vol. 36, pp. 346–352; March, 1948.

[8] J. S. Hey and G. S. Stewart, "Radar observations of meteors," *Proc. Phys. Soc.*, vol. 59, pp. 858–883; September, 1947.

[9] A. C. B. Lovell and J. A. Clegg, "Characteristics of radio echoes from meteor trails: I. The intensity of the radio reflections and electron density in the trails," *Proc. Phys. Soc.*, vol. 60, pp. 491–498; May, 1948.

Statistical Data from the Radar Observations—Determination of Shower Radiant

We shall define the angle α as the angle between the line drawn from the meteor to the observer (measured from the midpoint of the visual path unless otherwise specified) and the meteor path. For example, $\alpha = 90°$ defines a meteor traveling normal to the line of sight. Values of α less than 90° will represent meteors approaching the observer, while values greater than 90° denote meteors receding from the observer. For every recorded visual meteor, the position of the trail can be written down in terms of the azimuthal angle A and the angle of elevation h. If the visual meteor was identified as a shower meteor (by projecting its trail back to the shower radiant),[10] then it was possible to compute the angle α. Only shower meteors were used in these correlations.

It is of interest to study the distribution of the angle α over the visible sky, for a meteor radiant at a given angle of elevation h. For this purpose we shall assume, tentatively, that the region in which the meteors occur is a thin layer at a height of 100 km above a flat earth. The assumption of a curved earth complicates the calculations unnecessarily, as the corrections introduced are not significant at this stage. (When discussing elevation-range correlations it is desirable to take into account the curvature of the earth, and this has been done.) The geometry of the situation is outlined in Fig. 1 (a) (elevation). R_{max} is the maximum slant range available on the radar display.[11] The curves, α = constant, are de-

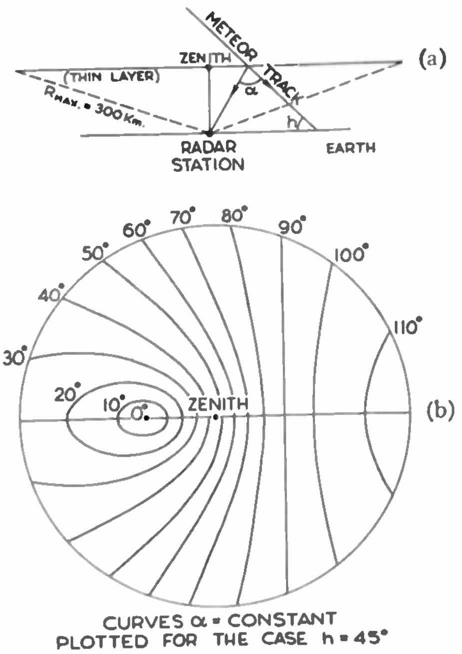

Fig. 1—(a) Elevation. (b) Plan.

[10] A meteor radiant is defined as the point on the celestial sphere which is the origin of the meteor trail when the meteor path is projected back to infinity.

[11] The terms "range" or "slant range" define the distance from the observer to the target.

fined by the intersection of the layer with the cone of half-angle α, apex at the radar station, and axis directed toward the radiant. These curves, which are conic sections, are plotted in Fig. 1 (b) (plan) for 10° increments in α and for an arbitrary radiant elevation $h=45°$. The particular case $\alpha=90°$ is of especial interest. From Fig. 1 it is apparent that the perpendicular distance from the radar station to the straight line $\alpha=90°$ is given by $R_{min}=100/\cos h$.

We shall examine briefly the statistical information available in the radar records alone. In this discussion, a typical shower period, December 12 and 13, 1947 (the maximum of the Geminid shower), and a typical nonshower or standard period, February 5 and 6, 1948, will be selected as samples. The expression "standard period" is open to question, since the minor shower radiants are so numerous it is difficult to select a time interval in which meteors could be described as wholly sporadic. However, the statistical effects of the minor radiants are small and tend to average out. Hourly counts were made of the number of echoes in arbitrarily selected slant-range classes of the range-time records. The numbers in each range class were further broken down into groups containing short echoes less than 1 second in duration, and groups containing echoes 1 second and longer. The hourly average ranges of the short echoes, of the long echoes, and of all echoes were then calculated.

Let us first analyze the data for the standard period. The hourly rates (all echoes) are plotted in Fig. 2. This rate curve is in general agreement with other meteor echo counts, and with the usual diurnal variation in number of visual meteors. Next, Fig. 2 also shows the hourly average range of all echoes for the standard period. If we separate the data into two periods, period A from 0000 to 1000 hours (7,500 echoes) and period B from 1300 to 2300 hours (4,000 echoes), and plot the frequency distribution in range, we get the curves shown in Fig. 3. The change in shape of the curves could be explained by a depression of the over-all radar parameters of several db during the period B, but this possibility was checked and eliminated. The probable explanation is that the morning echoes, as well as being more numerous, are actually stronger (therefore, visible at a greater range) than the evening echoes, since the average geocentric velocity of nonshower meteors is highest at about 0600 hours. A meteoroid of a given physical size presumably produces ionization proportional to some function of its geocentric velocity.

The curves in Fig. 3 can be represented by semi-empirical equations of an exponential form if certain assumptions are made concerning the reflecting properties of the meteoric ionization clouds and their distribution in space. However, as several widely divergent theoretical assumptions led to the same distribution curves, no significant conclusions were reached.

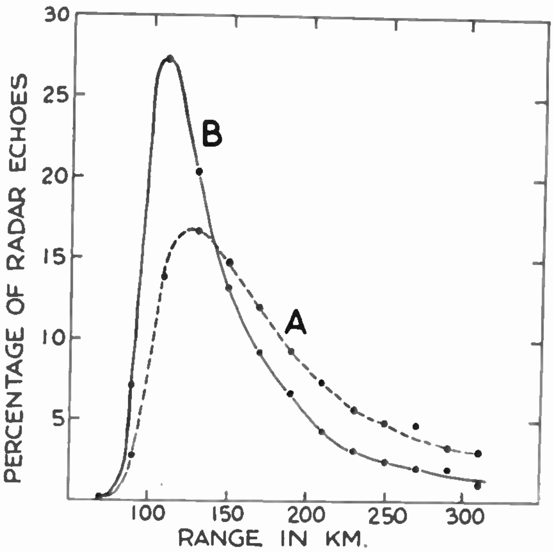

Fig. 3—Standard-period range-distribution curves.

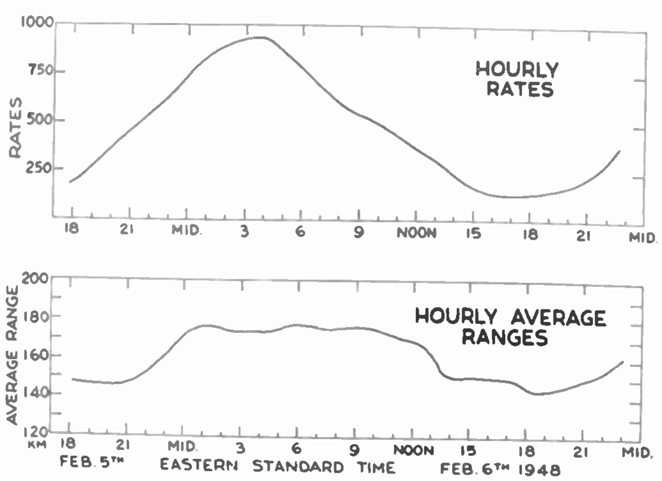

Fig. 2—Typical standard-period curves.

Let us now examine our typical shower period. The radar parameters were the same during the standard period and the shower period, so that many of the variables can be eliminated in comparing the observational data. The hourly rates of all echoes during the shower period are plotted in Fig. 4, together with the standard hourly rates from Fig. 2 for comparison purposes. A pronounced minimum in the shower period curve at 0200 hours is obvious. In Fig. 4 we have also plotted the hourly average ranges during the shower period. The solid curve is for all echoes, the dotted curve for the short echoes, and the dashed curve for the long echoes. The trend is the same for all three curves, each showing a minimum at 0210 hours. The long echoes show a consistently greater average range than the short ones, and this might be expected from noting that a long echo is usually a strong echo (the converse is not necessarily true).

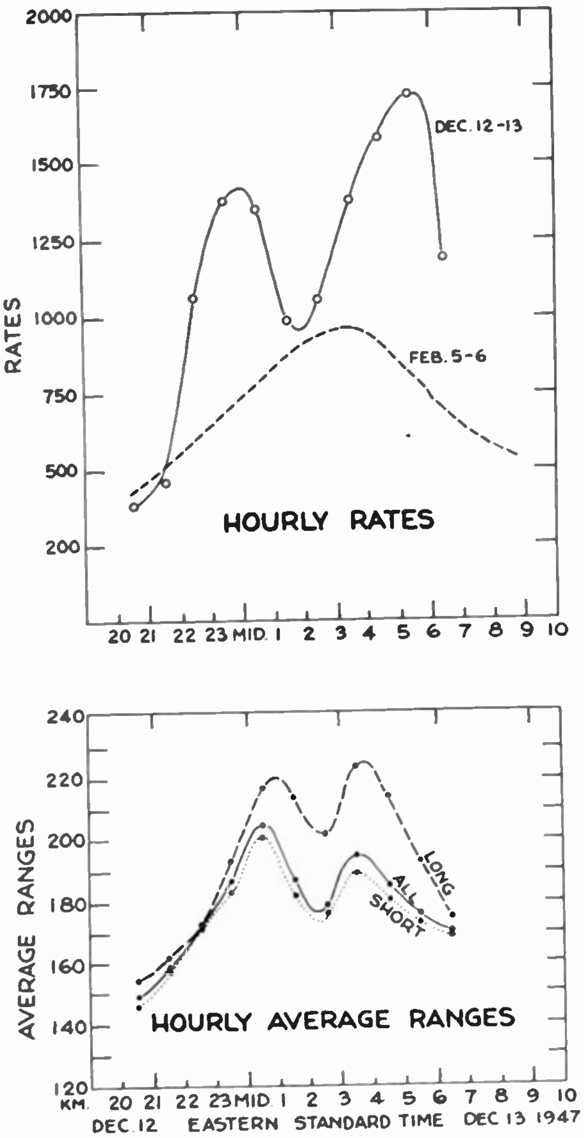

Fig. 4—Geminid shower period curves.

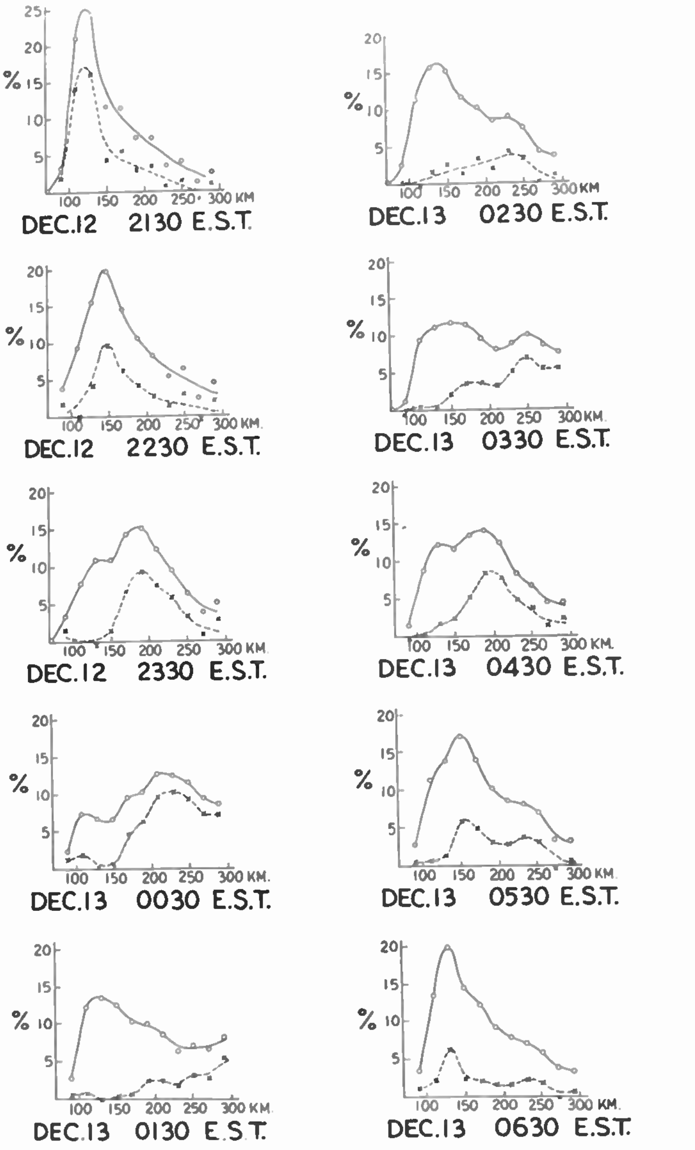

Fig. 5—Hourly range-distribution curves, Geminid shower period. Solid curves—observed distribution of all echoes; dashed curves—calculated distribution of Geminid echoes.

In Fig. 5 the solid curves show the hour-by-hour range distributions for the shower period, plotted for all echoes. We now make the assumption that the progressive deviations in shape of these shower period curves from the standard curves for the corresponding hours are due to the presence of the shower echoes, and then attempt to segregate the shower echoes from the standard background of sporadic echoes. From Fig. 4 a rough estimate of the hourly ratios of shower rates to standard rates is available and can be verified by fitting "rubber" standard-distribution curves inside each of the curves of Fig. 5. By subtracting the estimated percentages of sporadic echoes from the shower curves, the dotted curves of Fig. 5 have been obtained. These indicate the range distributions of the pure shower echoes.

It has been noted above that, in most cases, the meteoric ionization reflects the radar wave best when $\alpha=90°$. From Fig. 1 it is seen that the curve $\alpha=90°$ is a straight line across the thin ionizing layer, with the minimum slant range given by $R_{min}=100/\cos h$. In practice, there will be a spread in range and height of shower meteor echoes about this line. However, the greatest concentration of meteor echoes will be found in a small region around the line. Attenuation with range will reduce the contributions from meteors at the extremities, so that most of the echoes should be seen at about the range R_{min}. In other words, the ranges of the maxima of the pure shower curves should be a good first approximation to use in the above formula. The hourly values of h for the Geminid radiant are calculated in this way and plotted as points on Fig. 6.

The values of h in the neighborhood of 0100 to 0300 hours are in doubt, and it is obvious from Fig. 5 that the maximum range of 300 km available on the display was not great enough to permit a reliable determination of the maxima for those hours. This fact also accounts for the fortuitous sharpness of the dips in the hourly-rate and average-range curves, Fig. 4. By drawing a smooth curve (solid line, Fig. 6) through the remaining points, it is possible to determine the maximum elevation of the radiant. The time of the maximum elevation also can be estimated from this curve, but it can be obtained more accurately from either the average-range

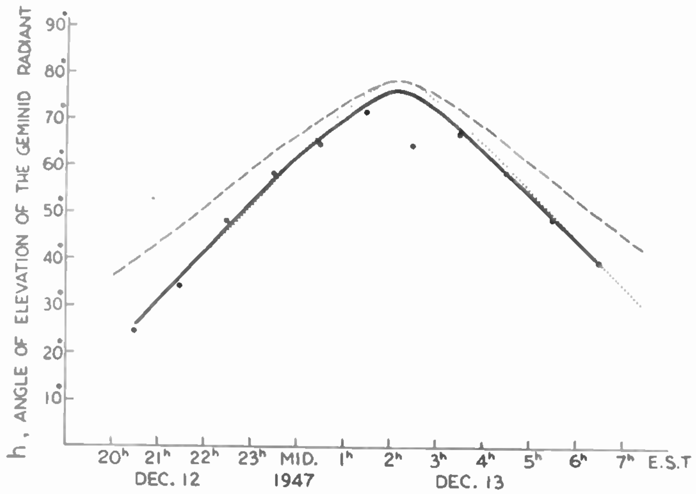

Fig. 6—Hourly angle of elevation of Geminid radiant; solid curve—radar data; dotted curve—computed for radiant at R. A. 112°, declination +33°; dashed curve—computed for radiant at R. A. 112°, declination +57°.

or the hourly-rate curves. The dotted curve in Fig. 6 shows the computed hourly elevation of the apparent Geminid radiant[12] with a maximum elevation of 78° at Ottawa, plotted by assuming the radiant at right ascension 112°, declination +33°, a position taken from the best available data.[13] The dashed curve shows the hourly elevation of an alternative radiant, right ascension 112°, declination +57°, which also has a maximum elevation of 78° at Ottawa. It is apparent that the radar observations agree with the curve for the first radiant rather than that for the second, and hence there is no ambiguity.

The radar determination of the Geminid radiant yielded a maximum elevation of 76° at 0208 hours EST, giving a position at right ascension 112°, declination +31°, in good agreement with that determined by Whipple. The probable error of the radar radiant is between 2° and 3°. One can thus determine the position of the radiant of a strong shower by statistical analysis of radar echoes, using an antenna which is effectively nondirectional and without moving the antenna or measuring angles. The method is applicable to daytime showers as well, but might be difficult to apply where several radiants are simultaneously active.

Correlation of the Visual Observations with the Radar Data

The list of visual shower meteors was compared with the radar film with a view to establishing coincidences in time. It was quickly found that such coincidences had most significance in the cases where the visual meteor was relatively bright and/or the radar trace was several seconds long. In fact, for the vast bulk of the short-duration radar echoes which coincided with visual meteors within an arbitrary time interval (e.g., plus or minus 4 seconds), it was found that the correlation was very nearly what one might expect if one assumed that the visual and radar meteors were completely independent of each other. Preliminary correlation criteria, based on time coincidence, radar-echo duration, and visual brightness of the meteor, were applied to isolate rather more than one hundred examples from 3,700 radar echoes and 1,100 visual meteors recorded during the Perseid shower. A further test was applied to the selected list. The slant range of the radar echo was used with the elevation angle of the associated visual meteor to calculate the height of the midpoint of the meteor path above a curved earth. If the height seemed reasonable, on the preliminary assumption that most meteor echoes occur between 80 and 120 km, the correlation was accepted. A list of 101 such visual-radar meteors was eventually produced from the Perseid data, and statistical considerations indicated that not more than three of this list were likely to be chance coincidences.

The frequency distribution of α for the 101 selected Perseid meteors has been investigated and is summarized in Table I. It is evident that the observed distribution with respect to α extends over a wide range of values.

TABLE I

Distribution of Meteors with Respect to α

α	0°		10°		20°		30°		40°		50°		60°		70°
Number		1		7		6		4		10		12		17	

α	70°		80°		90°		100°		110°		120°
Number		10		12		14		2		6	

Let us refer to Fig. 1(a) (plan). The value $h=45°$ was chosen as an average value for the Perseid radiant. The frequency of shower meteors to be expected in the interval α_1 to α_2 is proportional to the area between the corresponding curves. The actual number seen by the radar should be found by reducing this frequency by factors involving the range, the antenna pattern, and the angle α. The range factor is the subject of considerable discussion in the literature at present, and, while a reasonably fair estimate of the correction could be applied here, the scatter of the values in Table I does not seem to warrant it. A rough comparison between Fig. 1 and Table I does indicate a fair correspondence between numbers and areas and demonstrates that, on 32.7 Mc, the normal reflection law has little effect on the radar detectability of the brighter meteors.

The Geminid and Lyrid data have not been fully reduced as yet, but so far there has been nothing to con-

[12] The *apparent radiant* is defined by the direction of the observed meteor paths and is, hence, the observed radiant uncorrected for the effect of the earth's gravity; i.e. zenith attraction.

[13] F. L. Whipple, "Photographic meteor studies IV. The Geminid shower," *Proc. Amer. Phil. Soc.*, vol. 91, pp. 189–200; April, 1947.

tradict the conclusions we propose to draw from the Perseid data. On the other hand, the supporting evidence of the Geminid and Lyrid observations fills in some gaps in the Perseid material; partly because the slower velocities of the Geminids and Lyrids, 35 km per second and 48 km per second, respectively, compared to 61 km per second for the Perseids, provided us with different values of one of the parameters, and partly because the improved performance of the radar system brought out some detail on the radar display that was not available in the earlier run. Further significant information is available from dual-station observations made during the Lyrid shower, where two independent radar stations 57 km apart (Ottawa and Arnprior) were used in conjunction with visual and photographic observations.

The selected Perseid echoes were originally classified into six groups based on their general appearance.[4] These were lettered A to F, and it was found that there was a correlation between the form of the echo and the values of α. For example, echoes showing a range decreasing with time had α's less than 90°, and those with a range increasing with time had α's greater than 90°. These correlations were confirmed in the results obtained from the Geminid and Lyrid observations, and the original classification of echo types has more recently been somewhat altered and expanded to include a wider range of observational material.[5] While there is a great variety of echo forms, and many strong echoes exhibit an individuality of their own, it is possible to systematize most of the observed characteristics. The designations assigned in the revised classification are listed in Table II, and illustrated with examples in Fig. 7.

TABLE II

CLASSIFICATION OF METEOR RADAR ECHOES

Basic type—dependent on α	
"A"—indication of decreasing range or approach	($\alpha < 90°$)
"D"—indication of increasing range or recession	($\alpha > 90°$)
"U"—indication of both approach and recession	($\alpha \doteq 90°$)
"E"—two or more discrete ranges, no motion evident	($\alpha \neq 90°$)
"F"—one discrete range, no motion evident	($\alpha = 0°$ to 180°)
Secondary characteristics—indicated by small letters and numbers following letter for basic type	
"h" = instantaneous head echo apparently moving with velocity of meteor	
"b" = moving echo with measurable and continuously variable duration	
"e"—two or more echoes at discrete ranges	
"f"—one echo at a discrete range	
"n"—nebulous or diffuse echo	
"s"—duration under 1 second	
2, 3, 4, etc.—number of components.	

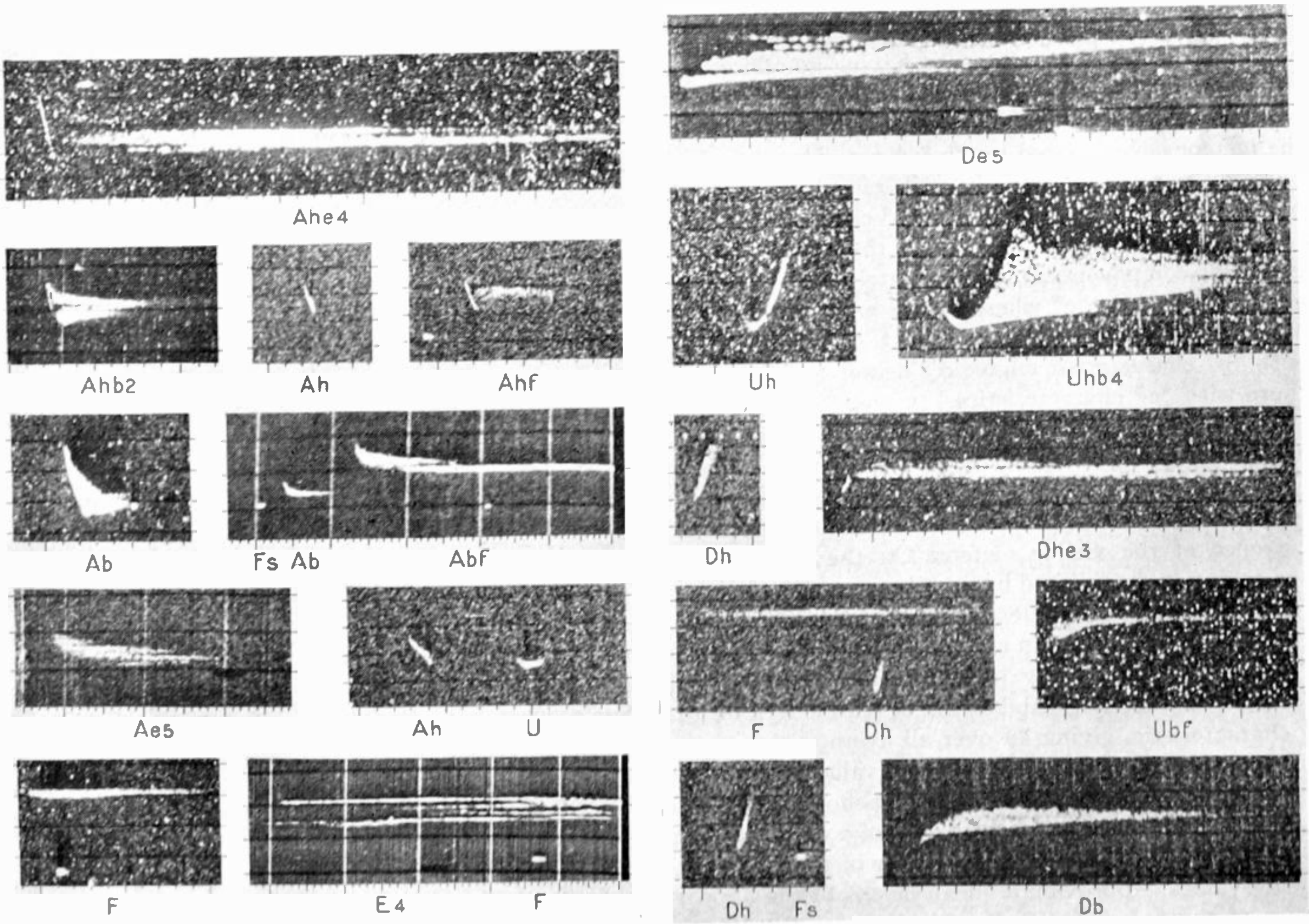

Fig. 7—Typical meteor echoes photographed on an intensity-modulated range-time display; range as ordinate, marked at 20-km intervals, time as abcissa, marked at one-second intervals.

The salient echo forms are described in more detail below.

(i) A large percentage of the bright and long-duration echoes shows evidence of a systematic change of range with time. As noted above, this property has been correlated with the observed α's. Echoes showing a decreasing range (i.e., approach) have been placed in type "A," while those with an increasing range, or recession, have been classified type "D." A less common form, exhibiting both approach and recession, has been called type "U".

(ii) Enduring echoes which do not fall into any of the above three types consist of one or more discrete components, each at an approximately constant but separate range. Those with two or more components at different ranges are called type "E," a numeral following the letter being used to indicate the total number of components observed. Enduring echoes at one range are type "F." The average range separation of the individual components of type "E" echoes is 3 to 4 km. Meteors producing echoes of type "E" average considerably brighter than those with type "F" echoes. Type "E" echoes, in general, correspond to meteors with α's differing considerably from 90°, but type "F" echoes may have α's of any value.

(iii) In considering compound echoes of type "A," "U," or "D," various classes of their component parts have been noted. A moving echo having no appreciable enduring characteristics, and with a range-time motion apparently corresponding to the geocentric velocity of the meteor is designated by "h." A moving echo whose duration varies continuously with range is designated by "b." The leading edge of the "b" characteristic may correspond to a velocity less than that of the meteor. Enduring components at approximately constant ranges are designated by "e" where two or more are observed, and by "f" where only one is present. In general, echoes with "b" characteristics have α's nearer to 90° than do those with "e" characteristics.

Apart from the general appearance and type of the echo, the two principal correlations with α involve the spread in slant range covered by the echo, and the mean delay in the appearance of the echo after the occurrence of the visible meteor. On the average, the spread in range exhibited by an echo is greater the more the corresponding α differs from 90°. For the Perseid echoes, where a spread in range could be measured, this corresponded to roughly 15 km of meteor path length for the *e* characteristic and 30 km of path length for the *b* characteristic, giving an over-all average radar path length of about 20 km. The average values of the time delay in the appearance of the echoes showed a definite increase as the corresponding values of α deviated from 90°, in spite of a considerable scatter of the individual delay values. Approximate mean delays of the appearance of the "e" and "f" characteristic for the Perseid meteors are listed in Table III.

For the selected Perseid echoes, the mean duration

TABLE III

α	Mean Delay
0°– 39°	10 seconds
40°– 59°	8
60°– 79°	6
80°– 99°	1
100°–119°	greater than 2

varied directly with apparent meteor brightness. There appears to be a straight-line relation between the logarithm of the echo duration and the apparent visual magnitude of the meteor. Echoes of 35 seconds duration correspond to Perseids of apparent magnitude zero. Examination of the Geminid and Lyrid observations indicates that the mean echo duration also varies with the geocentric velocity of the meteor; the durations of the Lyrid and Geminid echoes being shorter, corresponding to their lower velocities. The majority of the long-duration echoes ("e" and "f" characteristics) exhibit a slow drift in slant range. These drifts are of the order of 1 km in 25 seconds and seem to be predominantly away from rather than toward the observer.

In general, the visual meteor path length is longer than the path length observed by radar (by a factor of the order of 2 for the Perseids) but there is considerable evidence to show that the great majority of radar echoes occur somewhere along the visual meteor path. Making this assumption, and by combining visual and radar data, possible limits in height were computed for the selected list of Perseid echoes. The frequency distribution of all heights between these limits is shown in Fig. 8. The dotted curve is drawn for the meteors with the

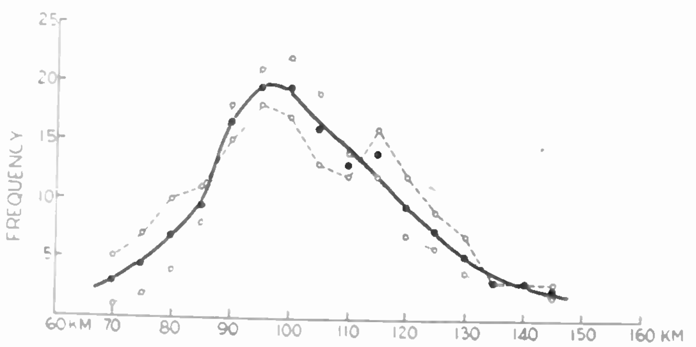

Fig. 8—Heights of enduring Perseid echoes. Solid curve—all selected echoes; dotted curve—echoes with more accurate visual observations; dashed curve—echoes with less accurate visual observations.

more accurate visual observations (about half the total number), while the dashed curve is drawn for the meteors less accurately plotted. The solid curve is the mean distribution for all the meteors. It will be noted that the maximum of the curve is considerably sharper for the better visual plots, indicating that a significant portion of the spread of this curve is due to observational error. By combining the known elevation of the radiant with the mean radar path length quoted earlier, we find that a bright meteor quite frequently produces enduring echoes extending over a height spread of 15 km. Let us assume a region, at least 15 km thick, in which enduring meteor echoes are most likely to be produced. Each of the three curves in Fig. 8 places this region at heights between 90 and 105 km above sea level, with a maximum just under 100 km. Heights for fifteen Lyrid meteors indicate values of the same order but with the maximum somewhat lower. This agrees with earlier observations by Hey and Stewart.[8]

A Theory of Meteor Radar Echoes

We shall first outline a phenomenological theory of the mechanics of radar reflections from meteoric ionization, and then examine whether the assumptions made explain the observations satisfactorily. The following postulates will be used:

1. There is an M region (M for meteor), in the upper atmosphere, which is slightly below but overlapping the recognized E region, and in which the great majority of the radar echoes from meteors occur. The M region is considered to have physical properties which sustain the ionization caused by the passage of a meteor. These properties have a maximum effect over a height spread of 15 to 20 km, centered somewhere between 90 and 100 km above sea level. The region may have diurnal and annual fluctuations in height, thickness, and ionizing ability.

2. A meteoroid will produce a long, thin column of dense ionization in its passage through the M region. Two mechanisms for the production of this ionization are envisaged. The first is a kinetic energy transfer through collisions involving both meteoric and air particles, the action occurring in the immediate vicinity of the meteoroid. The second is a radiation energy transfer produced by ultraviolet light from the meteor, which may be immediately effective at a considerable distance from the meteoroid.

3. In the M region are striae, or patches, which constitute a fine structure such that, within the striae, the physical properties of the atmosphere emphasize the formation and maintenance of meteoric ionization. No specific shape, thickness, extent, or orientation is assigned to these striae. They may exist as localized patches only, or they may extend throughout the M region in the form of horizontal, vertical, or inclined layers or rays. Their effective thickness is of the order of 1 km or less and the spacing between striae is of the order of 5 km.

4. The rate of expansion of any ionization cloud produced by the meteor is greatest at the top of the M region and decreases toward the bottom.

Fig. 9 is intended to show, diagrammatically, the successive stages in the growth of the ionized column, at times $t=1$, 3, and 10 seconds, say.

Immediately after the passage of the meteor, which normally takes less than 1 second to pass through the whole M region, the cloud of ionization is in the form of a long, thin column. This column reflects the radar wave most efficiently when viewed at an angle $\alpha=90°$. The received signal power is proportional to R^{-3}, which can be deduced from purely geometrical considerations.[14] The effective reradiation pattern (drawn in the plane containing the meteor path and the observer) of the column at $t=1$ second is such that the echoes from most of the meteors will not be detected at all for α's varying more than a few degrees from 90°.

Now, suppose we consider a meteor with $\alpha=70°$ at the midpoint, and which has produced a radar echo for 10 km on either side of the midpoint. This echo will appear to have a typical leading edge curl ("b" charac-

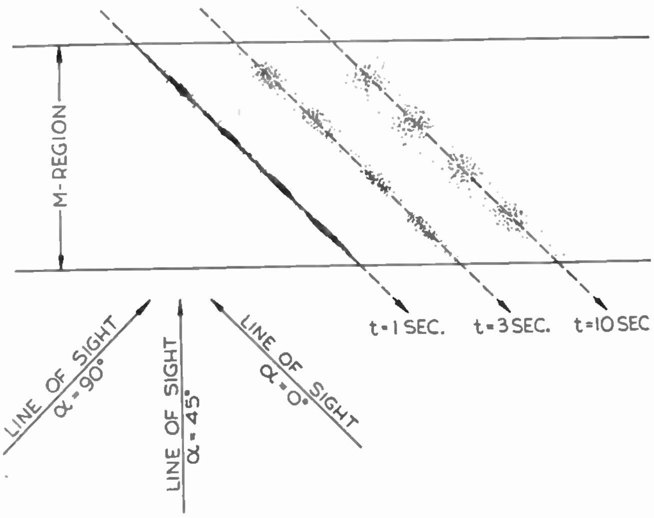

Fig. 9—Change of meteoric ionization clouds with time.

teristic), which generally represents an effective velocity somewhat less than the actual meteor velocity. On the basis of our assumptions, this retarded appearance is due to the slower growth of the ion cloud in the lower part of the M region. When viewed at a slight angle away from $\alpha=90°$ the radiation pattern is initially quite narrow, but quickly broadens out, and the broadening occurs first for the higher parts of the path. Thus there will be a slight but nevertheless real lag in the appearance of this echo, even for meteors with α near the 90° value. These lags might be measured in hundredths of seconds for the higher parts of the echo and tenths for the lower parts, and increase as α diverges from 90°.

In Fig. 9 the same meteor path is represented at time $t=3$ seconds. The ionization continues to expand as described in postulate 4, but the densities are rapidly thinning out. At time $t=10$ seconds, the densities between the striae may be so thin that a radar wave will not be reflected with detectable amplitude, despite the broadening pattern, but the striae patches may still be dense enough. Viewed from $\alpha=90°$, these patches may not be resolvable, since they may overlap in slant range. However, viewed from a low α direction, these striae patches may eventually build up to detectable and discrete target areas, provided their densities remain adequate.

[14] J. P. M. Prentice, A. C. B. Lovell, and C. F. Banwell, "Radio echo observations of meteors," Monthly Notices of the Royal Astronomical Society, vol. 107, no. 2, pp. 155–163; 1947.

The delays in appearance of the echoes will be successively greater for the lower patches and may be measured in seconds. The criterion for detectability is a combination of patch ionization density and solid angle of the patch as viewed from the radar. The received signal power in this case would probably be proportional to R^{-4}, as in the usual radar equation,[15] rather than R^{-3} for a thin cylinder.

Thus, only the brighter meteors viewed from low α's can cause echoes with "e" characteristics because several seconds are needed to blow up the patches to detectable targets, densities are rapidly thinning out in the process, and the effective range attenuation factor is increasing from R^{-3} to R^{-4}. The "Ae" echoes are considered to be the well-behaved ones and the processes described in the above paragraph account for them satisfactorily. The "E" echoes are thought to be cases in which the ionization is unevenly distributed through the various striae. The "F" echoes will occur when only one patch becomes detectable, or when the more complex characteristics of an echo near $\alpha = 90°$ are lost because of lack of range resolution.

Most of the visual meteors produce visible light that rises and falls in a continuous manner from start to finish of the trail. None of the meteors considered here exhibited luminosity bursts, which might explain the appearance of the "e" characteristic.

By assuming an arbitrary rate of diffusion of the ionized clouds in all directions, and then evaluating the solid angles of the clouds as viewed from different α's and at increasing time intervals after the passage of the meteor, it is possible to show that a time will come when, if a patch is sufficiently dense to be detectable at all, it will be detectable from any α. Examination of Fig. 9 will confirm qualitatively that the patches eventually should become roughly spherical. The dual-station results have confirmed this point; practically every echo that endured for more than five seconds at the Arnprior station was also displayed on the Ottawa radar film; the echo times and characteristics corresponded, though the ranges and "e" component separations varied as would be expected.

In any discussion of the origin and maintenance of the striae in the M region, the possible effect of the upper-atmosphere winds must not be forgotten.[8] The existence of such winds is evident from the motions of enduring meteor trains observed visually, the most recent compilation of observational data having been made by Olivier.[16] Motion in practically any direction, often turbulent in nature, and with velocities of the order of 200 km per hour (55 meters per second), is common. The original thin column of meteoric ionization is undoubtedly deformed by these winds, but that this deformation is entirely responsible for a number of discrete echoes appearing at such short range intervals, sometimes within a few tenths of a second, and maintaining identity over appreciable periods of time (even when viewed from two stations as in the Lyrid observations) seems hardly likely. It should also be pointed out that the frequent appearance of the F-type echo for meteors with low α's and visual path lengths in the neighborhood of 45 km indicates strongly the existence of sharper discontinuities in the physical conditions favorable to enduring echo production than would seem possible on the basis of wind motion alone.

In the case of the "h" characteristic (first noted by Hey and Stewart[5] for the Giacobinids) there is usually no measurable endurance of the echo, and for a very bright meteor this echo may appear well outside the M region. Since the motion of the "h" echo corresponds to the geocentric velocity of the meteor, it must be closely associated with the moving meteoroid. It is probable that the diffusion of ionization produced by collision is too slow to produce immediately a cloud big enough to reflect the 9-meter radio waves. We believe that instantaneous ionization by intense ultra-violet light at some distance from the meteor head may well explain the "h" characteristic echo.

It will be of interest now to examine, in more detail, the predicted appearance of the various echo forms. We shall assume that the echoes appear on a range-time display with a scale ratio of 10 km = 1 second (form factor = 10). Assuming a constant velocity for the meteor over most of its visible path, an assumption justified by photographic determinations of meteor velocity, the motion of the meteor on the range-time diagram will trace out a hyperbola. The form of this curve will vary with the geocentric meteor velocity v, and the perpendicular distance of the observer from the meteor path R_0 at time t_0. This hyperbola, $R^2 = R_0^2 + v^2(t - t_0)^2$, is the framework upon which any meteor echo is built.

A series of theoretical echoes has been plotted in Fig. 10, using typical values of the significant parameters and applying delays in the appearance and disappearance of the echo similar to those observed. The plotted echo outline has also been modified to take account of instrumental effects such as defocus and overshoot[17] of a bright trace, and the consequent masking of detail which would otherwise appear.

General Discussion

Assuming all the radar echoes to be due to meteors and to obey the normal reflection law, a satisfactory position of the Geminid radiant has been determined.

[15] 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.

[16] C. P. Olivier, "Long enduring meteor trains," *Proc. Amer. Phil. Soc.*, vol. 85, pp. 93–135; January, 1942; and vol. 91, pp. 315–327; October, 1947.

[17] Video overshoot was introduced in an attempt to obtain a three-dimensional graph of the echo. The length of the black shadow above the echo is a measure of the echo signal strength in excess of the clipping level.

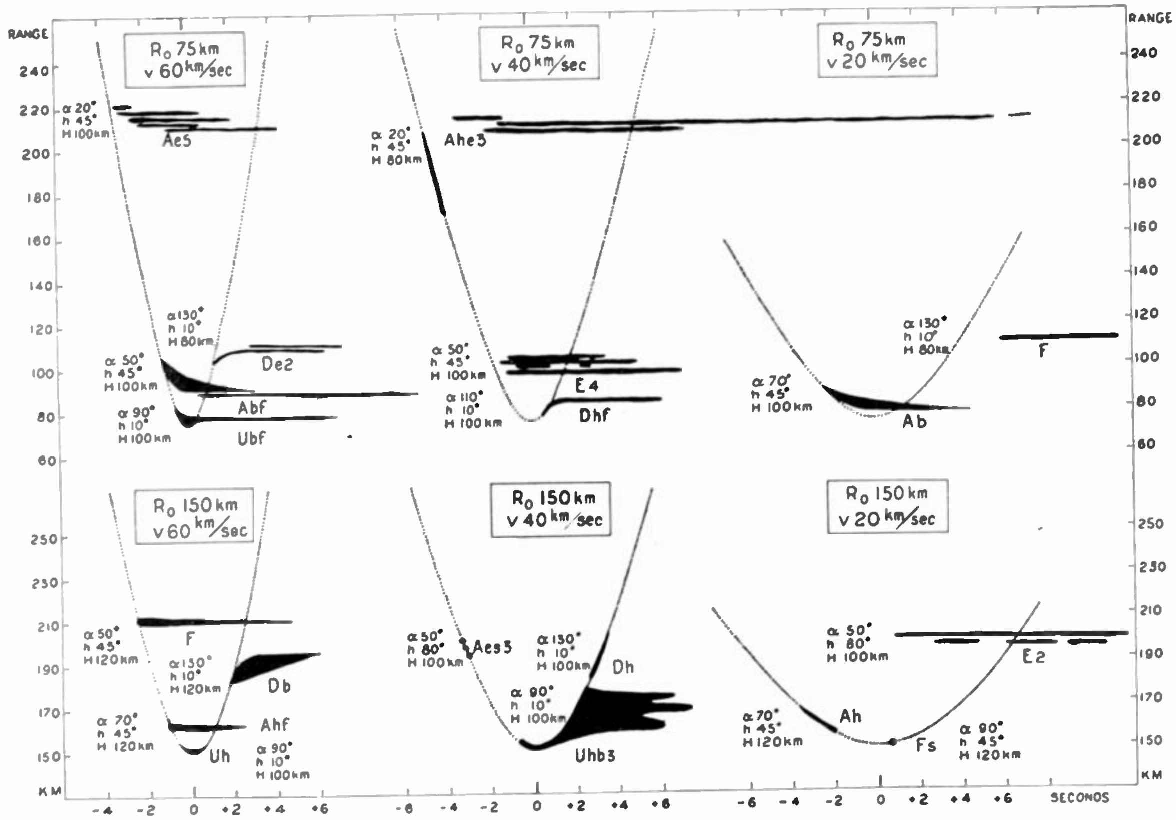

Fig. 10—Typical meteor echo forms. Representative echo types drawn by assuming various values of the following parameters, as indicated on the diagram:
R_0 = perpendicular distance from observer to meteor path
v = geocentric velocity of meteor.
α = angle between visual meteor path and line of sight.
h = angular elevation of the meteor radiant.
H = height of visual meteor path above sea level.

Basic hyperbolas have been shown as dotted curves. A distance on these hyperbolas corresponding to the assumed visual path length of 45 km can be drawn at any point by marking off time intervals of 0.75 second for v = 60 km per second, 1.12 seconds for v = 40 km per second, and 2.25 seconds for v = 20 km per second.

Doubt has been expressed in the past that all the transient echoes seen on frequencies below 50 Mc are due to meteors. The possibility of this cannot be denied, but by treating all the echoes as meteor echoes we have reached the conclusion that they all behave statistically as the known meteor echoes do. The myriads of small echoes may be due to telescopic meteors, or to nonluminous particles passing through one or more of the ionized regions.

As will be apparent from the hourly-rate curves of Fig. 4, the maximum ratio of shower echoes to sporadic echoes during the Geminid shower was of the order of 1 to 1. On the other hand, the ratio of visual shower meteors to sporadic meteors during the same night was between 2 to 1 and 3 to 1. The radar set can see echoes from meteors well below the visual limits, provided that the meteor paths are favorably oriented. The radar ratio should at least be equal to, if not greater than, the visual ratio during the periods when the radiant was at a favorable elevation, if one assumed that the size distributions of the shower particles and the sporadic particles were similar. Another observed fact is that the percentage of radar echoes of duration 1 second and longer during the shower periods was much higher than during typical nonshower periods (15 to 20 per cent, as compared to 8 to 10 per cent), indicating the presence of a greater percentage of large particles in the shower. One concludes, therefore, that there actually is a smaller percentage of faint meteors in the shower stream than in the sporadic background. This conclusion lends weight to the suggestion[18] that the established annual meteor streams, which have meteoric particles well distributed around their orbits and which originated thousands or even millions of years ago, contain few small particles. A younger shower, such as the Giacobinids which appeared in great strength in October, 1946, should accordingly yield a radar echo ratio more in line with the visual ratio.

Deep fluctuations in the signal strengths of the long-enduring echoes are frequently observed. These fluctuations are usually random in period and amplitude and are presumed to be due to the destructive interference effects of the signal received from various parts of the ionized cloud as it grows. Distortion by upper-atmosphere winds could contribute to this effect. Occasionally, a fairly regular amplitude flutter, of the order

[18] F. G. Watson "Between the Planets," Blakiston Co., Philadelphia, Pa., 1941; pp. 134–137.

of 5 to 10 cps, is superposed on the random fluctuations. As this is observed seconds after the meteor has occurred, it cannot be due to doppler effects from the moving meteor itself, athough it might be due to the doppler effect produced by the expansion of the ionized column. Successive rf pulses emitted by the radar transmitter are not coherent, so, if there is a true doppler effect, the reference phase must come from the ionized cloud itself. If one were to imagine that the central part of the cloud remains somewhat denser than the rest, and that the signal returned from this central part provides the reference phase, then, on the assumption that the interfering signal comes from near the expanding edge, one deduces that the rate of expansion is at least of the order of 50 to 100 meters per second. This analysis is, admittedly, highly speculative at this stage. Interference effects from the *moving* meteor have been detected by Ellyett and Davies,[19] using the already-formed part of the meteor ionization to supply the reference phase for the signal received from the latter portions as the meteor moves across the line of sight.

A possible explanation of the observed fact that most of the enduring echoes increase slightly in range before they fade out might be that the ionization density decreases below the critical density for the radio frequency employed, thus producing an apparent increase in range similar to the increase in virtual height of ionospheric records as the frequency is increased. We think that the critical-density hypothesis explains why workers using higher frequencies observe very few enduring echoes. The critical density of the cloud increases as the square of the frequency, so that a cloud detectable on 32.7 Mc will certainly remain visible much longer than on 72 Mc. As a consequence, we would expect fewer and fewer "e" characteristic echoes as the radio frequency is increased. Another way of stating this is that fewer echoes with α's diverging widely from 90° will be seen. On the other hand, as the frequency is lowered below 32.7 Mc, we would expect the "e" characteristic echoes to increase in number and duration. In particular, a very bright meteor should be capable of producing a large cloud of ions which might linger for hours with densities sufficient to reflect radio waves in the 2- to 25-Mc band. Such a cloud would drift with the winds, and, as it approached and receded from a vertical-incidence ionospheric station, the characteristic "abnormal-E" echoes would be obtained. We agree with Pierce[20] and with Appleton and Naismith[21] that meteoric ionization alone could thus be responsible for many of the "abnormal-E" effects observed in the 2- to 25-Mc band.

[19] C. D. Ellyett and J. G. Davies, "Velocity of meteors measured by diffraction of radio waves from trails during formation," *Nature*, vol. 161, pp. 596–597; April 17, 1948.

[20] J. A. Pierce, "Abnormal ionization in the E-region of the ionosphere," Proc. I.R.E., vol. 26, p. 892; July, 1938.

[21] Sir Edward Appleton and R. Naismith, "The radio detection of meteor trails and allied phenomena," *Proc. Phys. Soc.*, vol. 49, pp. 461–473; May, 1947.

Doppler Whistles

A remark on the doppler whistles from meteors, observed by many workers, seems in order here. A simple geometrical analysis shows that a whistle decreasing in pitch corresponds to an approaching meteor, and a whistle increasing in pitch to a receding meteor.[21] In some instances, it is probable that the whistle does represent the velocity associated with the moving meteor, but we feel that often the velocity deduced from the whistles will represent an apparent velocity[22] which is actually a measure of the leading-edge envelope ("b" characteristic) of the growth of the ionized clouds. Thus the whistle velocities would, in general, be somewhat less than the actual meteor velocities, and might even indicate fictitious decelerations. It should be emphasized, however, that these remarks are intended to apply to moving radar echoes and doppler whistles which extend over an appreciable time interval of the order of seconds, or which are comparatively remote from the t_0 point. The range resolution of the radar echoes obtained over a short time interval about the t_0 point is insufficient to predict the reliability of the corresponding doppler whistles. The doppler whistles might well yield accurate readings of velocities in the immediate vicinity of this point because, as has been shown above, the delay in echo appearance is least where $\alpha = 90°$, i.e., at t_0.

Some of the stronger meteors produced signals on the range-time display independently of the transmitter pulse, as was indicated by the appearance of noise on the film from zero to maximum range during the radar echo time. In some cases, the noise was strong enough to depress the inherent receiver noise by virtue of the video overshoot effect. Whether this noise actually emanated from the ionized cloud itself or was a reflection of some distant transmitter signal is not clear at present. The noise does appear to start 1 second or so *after* the passage of the meteor and to last as long as the enduring radar echo. The length of the noise pulses appears to be 2 or 3 microseconds, which is the limit of resolution of the receiver (bandwidth 300 kc). Similar bands of noise have appeared at times when no corresponding radar echo was seen, and these bands might have been due to meteors beyond the limits of the range display.

Conclusion

This paper is admittedly a preliminary analysis of the general problem, made chiefly from the observational viewpoint. An indication has been given of the value of the study of echo forms and their statistics in the detailed investigations of daytime meteors. The observational techniques employed are capable of yielding further information concerning the microscopic

[22] J. A. Pierce, "Ionization by meteoric bombardment," *Phys. Rev.* vol. 71, pp. 88–92; January 15, 1947.

factors involved (e.g., electron and ion densities) which have been discussed only superficially here, but this material will be left for a future paper.

ACKNOWLEDGMENT

It should be pointed out that there is a lengthy list of those who assisted in the construction and operation of the equipment, in the visual observations, and in the reduction of the data, and who through their enthusiasm contributed materially to the success of the over-all program. We should like, in particular, to acknowledge the assistance of Miss M. S. Burland, who directed the visual observers, and of E. L. R. Webb, who was in charge of the construction of the radar equipment.

A Spiral-Beam Method for the Amplitude Modulation of Magnetrons*

J. S. DONAL, JR.†, SENIOR MEMBER, IRE, AND R. R. BUSH‡

Summary—**A new method is described for the amplitude modulation of magnetrons. Although the method has been applied to the modulation of an existing cw magnetron at 900 Mc, scaling to higher frequencies should prove to be perfectly feasible. In principle, a beam of electrons spiraling in a longitudinal magnetic field varies the conductance presented by a resonant cavity coupled to the magnetron, and so varies the power delivered to the load. The peak power output of the system is at present 400 to 500 watts, while the necessary modulating power is only about 1/2 watt. The linearity of the system is reasonably good, and the bandwidth is at least 20 Mc. The depth of voltage modulation realized is 85 to 90 per cent, while the frequency variation during the amplitude-modulation cycle is only ±15 kc. A television picture free from obvious defects has for the first time, it is believed, been produced by the electronic amplitude modulation of a magnetron.**

I. INTRODUCTION

IT IS WELL KNOWN that the modern cw magnetron is an efficient oscillator capable of producing high rf powers. Thus, a kilowatt or more has been obtained at 10,000 Mc, several kilowatts at 3,000 Mc and tens of kilowatts at 1,000 Mc or below, all at efficiencies of 50 to 80 per cent. Modulation of magnetrons by means other than pulsing has been difficult, however. For example, variation of anode voltage for obtaining either amplitude or frequency modulation usually results in an undesirable mixture of the two and, in addition, requires high modulator power.

The method of frequency-modulating magnetrons proposed by Smith and Shulman[1] is particularly adaptable to wide-band services such as television; it has been employed in at least two developmental tubes.[2,3] It is the purpose of this paper to describe a similar electronic method for amplitude modulation, particularly adaptable to magnetrons, although not limited in its application to their use.

The method described has been applied to a 900-Mc cw magnetron, the rather low frequency being chosen largely because of the availability of the oscillator. About 500 watts of amplitude-modulated power were obtained in this investigation, but the method should be applicable up to several kilowatts. Rather than as a competitor of triodes or tetrodes, however, the chief utility of the method should be at higher frequencies, since it will be shown below that the power-handling capabilities of the spiral beam employed go up with the frequency.

An outstanding advantage of the system at any frequency is the low modulating power required, less than 1 watt to modulate 1/2 kilowatt with a bandwidth of 10 Mc.

II. THEORY

Gutton[4] has employed a magnetron as a variable impedance to modulate a klystron. The depth of modulation appears to be excellent, although the modulating voltage is rather high. Triodes have been used to amplitude-modulate other triodes at lower frequencies,[5] employing a circuit similar to that to be described. It appeared, however, that a beam of electrons spiralling in a

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[1] Lloyd P. Smith and Carl I. Shulman, "Frequency modulation and control by electron beams," PROC. I.R.E., vol. 35, pp. 644–657; July, 1947.

[2] G. R. Kilgore, C. I. Shulman, and J. Kurshan, "A frequency-modulated magnetron for super-high frequencies," PROC. I.R.E., vol. 35, pp. 657–664; July, 1947.

[3] J. S. Donal, Jr., R. R. Bush, C. L. Cuccia, and H. R. Hegbar, "A 1-kilowatt frequency-modulated magnetron for 900 megacycles," PROC. I.R.E., vol. 35, pp. 664–669; July, 1947.

[4] H. Gutton and J. A. Ortusi, "Ultra-high-frequency modulation on waveguides," *Jour. Brit. Inst. Rad. Eng.*, vol. VII, pp. 205–210; October, 1947.

[5] William N. Parker, "A unique method of modulation for high-fidelity television transmitters," PROC. I.R.E., vol. 26, pp. 946–962; August 1938.

longitudinal magnetic field might prove more efficient than a triode for such modulation.

Smith and Shulman[1] pointed out that, when an electron beam is subjected to a transverse rf electric field in a resonant cavity, and to a longitudinal magnetic field satisfying the relation

$$\omega = He/m \tag{1}$$

where ω is 2π times the operating frequency, the admittance presented by the beam is real. (In (1) and in all subsequent expressions, centimeter-gram-second electrostatic units are used.) This conductance is given by

$$G_m = e/m \frac{I_0}{d^2} \cdot \frac{\tau^2}{4} \tag{2}$$

where I, τ, and d are the beam current, the electron-transit time, and the distance between the plates, respectively.

If one or more electron beams were placed within the cavities of the magnetron, the low magnetic field needed to satisfy (1) would be present in the interaction space of the oscillator, and would reduce its efficiency. More significant, however, is the fact that the conductance of (2) would be effectively in shunt with the magnetron and would load it more heavily, increasing its total power output. Thus the diminution of load power would be less than desired.

By placing the modulating beam in a separate cavity, it is possible to decrease, rather than to increase, the magnetron efficiency when the load is shunted by the conductance of the beam. The circuit used to accomplish this is shown in Fig. 1. The resonant cavity containing the grid-controlled modulating beam is coupled by a loop to a coaxial line, and presents a conductance

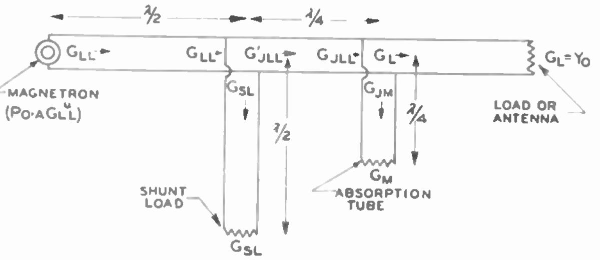

Fig. 1—Circuit used to vary the antenna power by grid control of the beam current in the modulation tube. The shunt load is used to prevent complete unloading of the magnetron, possibly resulting in poor spectrum.

G_M to this line. The shunt load is employed to prevent almost complete unloading of the cw magnetron, with possible harmful effects upon the spectrum and bandwidth. A shunt load conductance (G_{SL}) of 0.0025 mhos was found to be sufficiently high, resulting in only a slight decrease in the system power output.

Qualitatively, when the beam current in the absorption tube is zero, G_M is very low (about 0.0008 mhos), limited by losses in the cavity and tube. This results in what is substantially a short circuit at the right-hand junction, reducing the power in the load not only by shorting it but also by presenting a very low conductance to the magnetron and reducing its efficiency. Thus good depth of modulation is obtained. When the modulating beam is biased on, G_M increases but G_{JM} decreases. If the beam were infinite, the load power would be approximately that of the magnetron into a matched load.

If the line to the absorption tube were a half-wavelength long, rather than the quarter-wavelength shown, the maximum power into the load, occurring at zero beam current, would be nearly that of the magnetron into a matched load, but for practical values of beam current the depth of modulation would be very poor, possibly only 50 or 60 per cent in voltage.

Before considering the relation between the beam current and the conductance presented by the beam, the expressions for the powers in the components of the circuit external to the modulating tube will be derived in terms of G_M, the conductance presented to the line in the circuit of Fig. 1. It is convenient to assume that in the region of operation the power output of the magnetron may be expressed by the relation

$$P_0 = AG_{LL}^u \tag{3}$$

where A is a constant, G_{LL} is the conductance seen by the generator, and u is positive and less than unity. The validity of this assumption has been confirmed by experimental data on several tubes of different types. An average value of u of 0.33 was found and has been employed here. Rather wide variations from this value for an individual tube would make no serious changes in the conclusions reached later concerning the modulation characteristics.

It is further assumed that

$$G_M = \frac{G_m}{k^2} = aG_L = aY_0 \tag{4}$$

and

$$G_{SL} = \frac{G_L}{n} \tag{5}$$

where a and n are constants, G_L is the conductance of the useful load, Y_0 is the characteristic admittance of the lines, and k is the coefficient of coupling between the resonant cavity and the line.

From (3), (4), and (5),

$$P_0 = A\left[\frac{1}{n} + \frac{a}{1+a}\right]^u G_L^u. \tag{6}$$

If the powers in the shunt load, the useful load, and the modulation tube are represented by P_{SL}, P_L, and P_M, respectively, the powers in the absorption tube and loads can be expressed in terms of P_0 by the following relations:

$$P_{SL} = P_0 \frac{1+a}{1+a(n+1)}, \tag{7}$$

$$P_M = P_0 \frac{na}{(1+a)[1+a(n+1)]}, \tag{8}$$

$$P_L = P_0 \frac{na^2}{(1+a)[1+a(n+1)]}. \quad (9)$$

It is convenient to plot the ratios of the powers to the power output of the magnetron into a matched load, for the ordinates of the curves are then independent of the latter power. Since the output into a matched load is

$$P_{00} = AG_{LL}{}^u = AG_L{}^u, \quad (10)$$

(6) becomes

$$\frac{P_0}{P_{00}} = \left[\frac{1}{n} + \frac{a}{1+a}\right]^u. \quad (11)$$

Using (6), (7), (8), and (10), the ratios of P_0, P_M, P_L, and P_{SL} to P_{00} may be computed and are shown plotted on Fig. 2 for a value of n of 8. It will be noted that, when the modulating-tube conductance is zero, the relative output of the magnetron is 0.5. The output of the magnetron is increased for all values of a by the presence of the shunt load (for very high values of a it is greater than P_{00}), with the result that the useful power in the load is not decreased, by the presence of the shunt load, as much as would otherwise be the case.

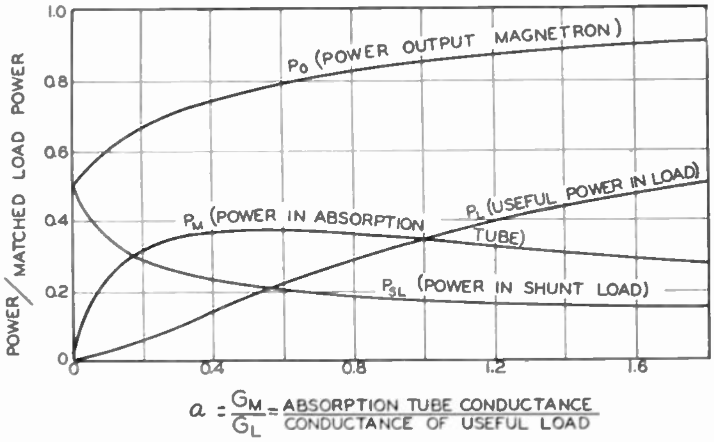

Fig. 2—Variations of the ratios of the rf powers to the power output of the magnetron into a matched load, as a function of the conductance presented to the line by the modulation tube.

Modulation Characteristics

The modulation characteristic is usually defined as the relation between the rf voltage across the useful load to the control voltage on the modulating element, here the voltage on the control grid of the modulation tube. Since the conductance is expressed in (2) in terms of the beam current, the modulation characteristic will be derived in terms of this current, which in turn is assumed to vary as $Eg^{3/2}$. It is necessary to determine the characteristic in two domains, namely, that in which electrons strike the pole faces of the modulation tube as well as that in which they do not.

In the following simplified derivation of the modulation characteristic, space charge and the finite size of the beam are neglected. It is further assumed that the poles of the modulation tube are parallel planes spaced apart by $2R$, twice the maximum radius of the electron spiral.

The amplitude of the spiral which the electrons execute between parallel-plane pole faces has been found by Smith and Shulman[1] to be

$$r = \frac{e}{m} \cdot \frac{V_m}{4\omega R} \cdot \tau \quad (12)$$

where V_m is the amplitude of the rf voltage between the pole faces, and R is half the distance between them. If the spiral is sufficiently small and the electrons do not strike the pole faces, τ is constant and dependent only upon the beam voltage. If the rf field increases until the electrons strike the pole, however, r is constant for increasing V_m, and τ becomes a function of V_m. Thus the conductance, as well, becomes a function of V_m, from (2). The relation of the power in the load to the conductance, shown in Fig. 2, remains the same, but the variation of load power with beam current, and hence the modulation characteristic, would be expected to alter when the electrons start to strike the pole faces. At low beam currents, with low losses in the tube and resonant cavity, the rf voltage between the poles would be very high, approaching infinity as the losses approach zero, and the electrons would be expected to strike the poles. It is necessary, therefore, to determine the modulation characteristic, as a function of system power output, losses, and coupling coefficient, in some detail. It is possible that a sharp change of slope of the characteristic, when the electrons cease to strike the poles, might make the method unusable.

Let the critical voltage V_c be the value of V_m at which the electrons cease to strike the poles. From (12),

$$\tau = \frac{4\omega R^2}{V_c \cdot e/m} = \frac{l}{\sqrt{2e/mV_b}} \quad (V_m \leqq V_c), \quad (13)$$

$$\tau = \frac{4\omega R^2}{V_m \cdot e/m} \quad (V_m \geqq V_c). \quad (14)$$

From (2), (13), and (14),

$$G_m = \frac{(\omega R)^2 I_0}{V_c{}^2 \cdot e/m} \quad (V_m \leqq V_c), \quad (15)$$

$$G_m = \frac{(\omega R)^2 I_0}{V_m{}^2 \cdot e/m} \quad (V_m \geqq V_c). \quad (16)$$

$V_m < V_c$ (electrons not striking poles)

Since P_L, the power in the load, is known in terms of a from Fig. 2, it is desirable to express the beam current I_0 in terms of a to determine the relation between $\sqrt{P_L}$ and I_0. Taking account of losses,

$$a = \frac{G_M}{G_L} = \frac{(G_{loss} + G_{MI})}{G_L} = a_0 + \frac{G_m}{k^2 G_L} \quad (17)$$

where the conductance presented outside the coupling loop is divided into that due to losses and that (G_{MI}) due to the beam. G_{loss}/G_L is represented by a_0, the minimum value of a, obtained when the beam is off. Using (13) and (15),

$$I_0 = (a - a_0) \frac{32k^2 G_L V_b R^2}{l^2} . \quad (18)$$

In order to calculate the modulation characteristic for the case in which the electrons do not strike the poles, a value of the coupling coefficient k is assumed, together with values of a greater than a reasonable experimental value of a_0. For each a, P_L/P_{00} is read from Fig. 2 and plotted against $I_0^{2/3}$ obtained from (18). Assuming V_b and l to be 1,000 volts and 15 cm respectively, the solid curves of Fig. 3 were calculated for the indicated coefficients of coupling. The value of a_0, which determines the intercept on the axis of ordinates, was obtained by assuming that, when P_L is half of P_0 (Fig. 2), the losses permit P_L to be modulated 85 per cent in voltage. These values are representative of experimental results.

$V_m \geqq V_c$ (electrons striking poles)

Employing a procedure analogous to that used in the derivation of (18), and defining P_{MI} as the power absorbed by the beam alone,

$$a = \frac{G_M}{G_L} = a_0 + \frac{G_{MI}}{G_L} = a_0 + a\frac{G_{MI}}{G_M} = a_0 + a\frac{P_{MI}}{P_M} \cdot$$

$$= a_0 + a\frac{\frac{1}{2}V_m{}^2 G_m}{P_M} \cdot \quad (19)$$

Substituting for G_m from (16) and solving for I_0,

$$I_0 = \left(\frac{a - a_0}{a}\right) \frac{2e/m}{(\omega R)^2} P_M. \quad (20)$$

Therefore, to calculate the modulation characteristic for the case when electrons are striking the pole faces, increasing values of a are assumed beginning with the value of a_0 used in (18). For each value of a, P_L/P_{00} and P_M/P_{00} are read from Fig. 2. For an assumed P_{00}, I_0 is calculated from (20) and $\sqrt{P_L/P_{00}}$ is plotted as a function of $I_0^{2/3}$ to give the modulation characteristic. Using 900 Mc and a value of R of 2 cm, the dashed lines of Fig. 3 were so calculated from the values of P_{00} indicated.

It is to be noted that in Fig. 3 the curves for $V_m \geqq V_c$ are independent of coupling coefficient, but dependent upon magnetron power output, while the reverse is the case for the solid curves for which $V_m < V_c$. For a given combination of power output and coupling, the dashed curve is followed, if it lies below the solid curve, until it intersects the latter, at which point the voltage within the modulation tube is V_c, the voltage at which the electrons are just grazing the poles as they are collected at the end of the tube. Above the intersection of the dashed and solid curves, the latter gives the modulation characteristic.

Fig. 3 shows that, for magnetron matched-output powers up to 500 watts (system power outputs of 250 to 300 watts), there is no bulge in the modulation characteristic for a coupling coefficient of 0.05 or higher. At this coupling coefficient, 50 ma would be expected to yield 50 per cent of the matched-load magnetron output. At a magnetron output of 1,000 watts, however, it would be necessary to increase the coupling coefficient to 0.075 to remove the discontinuity in slope. At this coupling coefficient, about 100 ma would be expected to yield the 50

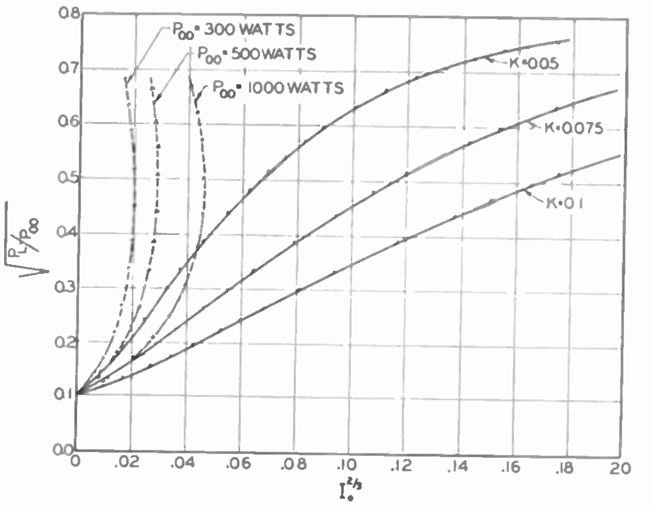

Fig. 3—Theoretical modulation characteristics for the case (dashed curves) when electrons are striking the pole faces of the modulation tube, and for the case (solid curves) when the rf voltage is reduced until the electrons do not strike the poles.

per cent relative output. Thus the theory predicts that, as the power is increased, the beam current must also increase in order to maintain a good modulation characteristic. In practice, the changes in slope would not be as abrupt as indicated in Fig. 3, since the finite diameter of the beam would have a smoothing effect.

Should the system power output be pushed to such high values that the curvature of the modulation characteristic is troublesome, and should it be impractical to increase further the coupling and the beam current, an increase in modulating-tube diameter would result in great improvement. Thus, from (18), if R is doubled, k may be halved and a given current would still yield the powers indicated by the original solid curve of Fig. 3. However, doubling R would reduce the current in (20) to one-fourth, which results in the dashed curves of Fig. 3 moving to the left, in this case until the 1,000-watt curve assumes approximately the position of the 300-watt curve shown. The net result is that the dashed curves would not intersect the solid curves unless the power were increased several fold or k reduced in a like ratio.

Scaling to Higher Frequencies

The energy absorbed by each electron in the beam is

$$E = \tfrac{1}{2}mv^2 = \tfrac{1}{2}m(\omega r)^2 \quad (21)$$

where v is the tangential velocity, and r is the radius of rotation. For N electrons per second, the total power absorbed is

$$P = NE = \frac{I_0}{e} E = \frac{I_0 \omega^2 r^2}{2e/m} . \quad (22)$$

When the electrons strike the poles just before collection, r has the maximum value of R, so that, for fixed I_0

and R, the power absorbed will vary as the square of the frequency. Thus, for the same geometry the performance would be greatly improved at higher frequencies..

It can be shown that the modulation characteristics of Fig. 3 will improve with increasing carrier frequency, in that their curved portions (dashed) will recede to still lower grid voltages. In plotting the solid curves of Fig. 3, $\sqrt{P_L/P_{00}}$ is obtained from assumed values of a. From (18), however, the I_0 calculated for these assumed values of a is independent of frequency; hence, the solid curves are unchanged with increasing frequency. In calculating the dashed curves, however, the currents from (20) vary inversely as the square of the frequency. Therefore, all ordinates will move to the left, the ratio of the abscissae being in the inverse ratio of the four-thirds power of the frequency. At only 1,800 Mc, the 1,000-watt curve of Fig. 3 would move into the position of that labeled 300 watts.

III. Tube Construction

If the object of this investigation had been to produce a final engineering design affording the maximum power from the system, the electron beam and the resonant cavity would have been placed within a metal evacuated envelope, eliminating all possibility of structural failures due to high rf fields within the cavity. However, to afford great simplicity of tube construction and to provide ease of adjustment of resonant frequency and coupling coefficient during the developmental work, the electron beam was confined within a glass capsule inserted into the cavity. Fig. 4 shows a typical tube, the wall of which is of low-loss glass. Electrodes, of heavy silver paste with leads extending through the walls, form internal pole pieces between which the rf field is applied. These electrodes, at dc ground potential, prevent changes in the electron optics of the device which might result from charging of the walls by the beam. The disk visible in Fig. 4 is an accelerating aperture held at about 600 volts below the collector and pole-face potentials, while the cathode is about 1,000 volts below collector potential.

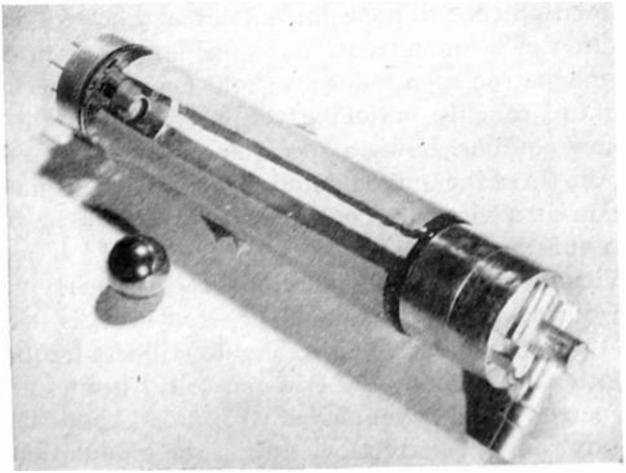

Fig. 4—A glass-capsule modulation tube, When the tube is inserted into the resonant cavity, the silver-paste pole faces make contact with the pole faces of the cavity. The 1-inch-diameter ball bearing indicates the scale.

IV. Resonant Cavity and Magnet

The resonant cavity (Fig. 5) is effectively a half-wavelength section of coaxial line, shorted at the ends. The cavity pole pieces, one of which may be thought of as the center conductor of the coaxial line, extend from one lid to the other, as is best seen in Fig. 6, with the corresponding wall coatings within the tube extending from the position of the accelerating aperture to just short of the collector. Since the rf field between the pole faces falls to zero at the shorted ends of the coaxial line, there are no disturbing rf fields parallel to the beam between the ends of the pole faces and the lids, as there would be in a cylindrical cavity with axially short pole faces projecting toward the axis.

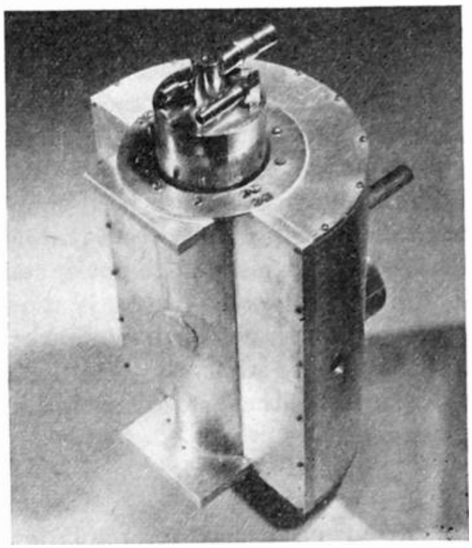

Fig. 5—Coaxial resonant cavity showing the glass modulation tube in position.

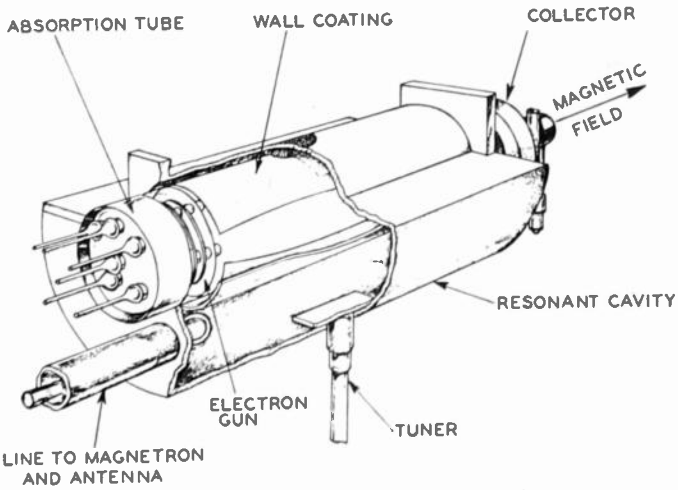

Fig. 6—A perspective view of coaxial resonant cavity, cut away to show the capsule modulation tube. The central half-cylindrical pole piece is connected to the ends of the cavity.

The heavy leads through the envelope make contact with the midpoints of the cavity pole faces. The coupling between the cavity and the coaxial line, to which the loop is attached, is adjustable by rotation of the line. The cavity may be tuned by adding lumped capacitance (Fig. 6) at the midpoint of the length of the coaxial cavity.

The required magnetic field of 325 gausses at 900 Mc, from (1), is supplied by the solenoidal type magnet with an external iron shell which is shown in Fig. 9.

V. Performance

The control-grid capacitance of the modulating tube is about 30 $\mu\mu$f, and the total grid swing, from cutoff to maximum beam current, is about 40 volts. If the output of the modulator is considered to be a simple *RC* circuit, the resistance necessary to give a bandwidth of 10 Mc is 500 ohms, so that the average power consumed by this circuit, for the grid swing stated, is about one-half watt. The grid current drawn at maximum grid voltage is less than 10 ma, so that the additional average power consumed due to the conduction current is of the order of 1/10 watt.

Power Output

The beam current necessary for modulation is a function of the coupling between the resonant cavity and the line, as shown in the solid curves of Fig. 3. With the coupling usually used, 50 to 60 per cent of the power output of the magnetron into a matched load is developed across the system load for a beam current of 75 ma. Higher currents result in undue curvature of the modulation characteristics.

The maximum power output of the system is limited at present to between 300 and 500 watts by possibility of failure of the glass envelope of the absorption tube, because of overheating in the high rf field. Increasing the coupling would, of course, reduce this danger, but would require higher beam currents. While enclosing the beam and resonant cavity within an evacuated metal envelope would remove this cause of failure, a low-loss ceramic envelope is under development in order to retain the advantages of the separate capsule tube.

Depth of Modulation

The depth of voltage modulation is of importance, not only because of standards imposed on some systems, but because the signal in the receiver is greater as the depth of modulation is increased, for the same power swing. The minimum power (Fig. 3) in the system under discussion is determined by the losses in the resonant cavity and absorption tube. These losses are at present largely in the glass envelope, in the pole-face coatings of the modulation tube, and in the leads, which absorb some rf power from the cavity. Thus, per 100 watts of system output, the losses in the empty cavity are about 0.2 watt, while successive additions of the glass envelope, the wall coatings, and the leads add about 0.5, 0.6, and 0.7 watt, respectively. The total of 2 watts corresponds to the approximately 85 per cent voltage modulation obtained for the system. Further development work should prove fruitful in reduction of the losses, and should increase the obtainable depth of modulation.

Frequency Stability

Since the magnetron oscillator[3] used in this system is equipped with grid-controlled frequency-modulation and control beams, maintaining the average carrier frequency constant by comparison with a standard cavity or a crystal-controlled oscillator would be comparatively simple. Of more interest is the frequency change encountered during the amplitude-modulation cycle. Since the electron beam presents a pure conductance, the circuit of Fig. 1 presents a pure conductance to the magnetron. If the Rieke diagram of the magnetron is ideal, in that a frequency contour lies along the zero-reactance line, no frequency change occurs during the modulation cycle. In practical cases, the Rieke diagram is not usually ideal, however, and the frequency will shift, although the shunt load decreases the range of impedances seen by the magnetron, and hence decreases the frequency change to be corrected. In order to minimize the frequency variation, the lines between the right-hand junction of Fig. 1 and the magnetron are first adjusted, with the beam of the modulator tube off, until the impedance presented to the magnetron lies on the same frequency contour as the impedance presented with the beam of its maximum value. This easy adjustment gives the same frequency at the end points of the modulation cycle. It is not necessary to alter the line lengths when modulation tubes are replaced.

The second adjustment for the correction of frequency changes is to vary the magnetic field of the absorption tube very slightly (with negligible change in power output of the system) from the value of (1). This causes the modulation tube to present a small reactance which is a function of beam current. The impedance presented to the magnetron then follows a slight curve on the Rieke diagram, roughly matching the curvature of the frequency contour. Using a spectrum analyzer, for example, the tip of the unmodulated spectrum can be made to remain at a virtually constant frequency as the beam current is varied.

With the adjustments described, the frequency variation during the amplitude-modulation cycle is usually less than ± 15 kc at 900 Mc. Should still less frequency variation be desired, the grid-controlled beams in the magnetron could be employed with broad-band control circuits to make corrections during the modulation cycle, by comparison of the system frequency with a standard frequency.

Bandwidth

The bandwidth of the system was examined by impressing on the grid a (known-amplitude) signal of frequency variable from 100 kc to 20 Mc, amplitude-modulated at 400 cps. The output of a diode detector inserted into the rf line at the load was passed through a second detector and viewed on an oscilloscope. The response of the system was found to be substantially independent of frequency up to about 20 Mc; this result was independent of the depth of voltage modulation of the rf carrier over the range from 5 to 85 per cent.

Linearity

The degree of linearity of the system, that is, the linearity of the square root of the power output versus the control-grid voltage, has been determined for manual control of the bias, for 60-cps modulation, and for several modulation frequencies in the range from 1 to 20 Mc. A representative result at the higher frequencies is shown in Fig. 7, where the depth of modulation, for a constant dc grid bias, is plotted against the rms grid volts. Within the probable error of the results, the characteristic is linear up to about 85 per cent voltage modulation; this is the maximum obtainable modulation, limited by losses as explained earlier.

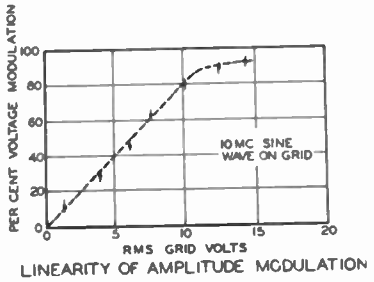

Fig. 7—The modulation characteristic at high frequencies, in terms of depth of voltage modulation versus control-grid voltage.

Manual control of the bias and 60-cps modulation gave substantially identical results. Under either of these conditions all quantities could be measured more accurately, and the results could be compared directly with the theoretical characteristics of Fig. 3. In the 60-cps case, for example, for each measured peak-to-peak swing in grid voltage and peak beam current, a voltage proportional to the maximum rf voltage across the load was measured, using a diode coupled to the rf line. Using the same diode, the relative voltage developed by the magnetron when coupled to a 52-ohm load was measured at each magnetron input power for which a modulation characteristic was obtained. The ratios of the rf voltage during modulation of the beam to the rf voltage across a 52-ohm load are plotted as a function of grid voltage in Fig. 8 for power outputs of 400 and 700 watts from the magnetron into a matched load. The maximum values of the peak system power outputs are thus about 200 and 400 watts for the maximum rf voltages indicated in the figure; the depths of voltage modulation are about 85 and 86 per cent, respectively.

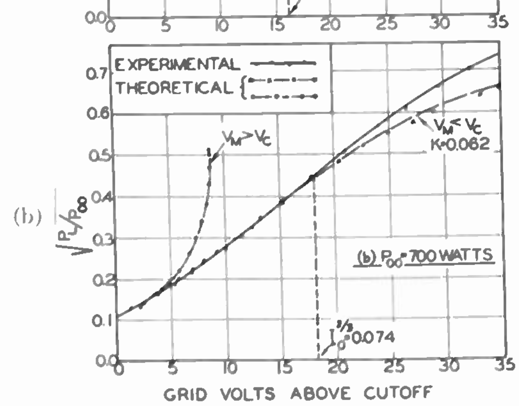

Fig. 8—A comparison between experimental and theoretical modulation characteristics at two power levels.

For comparison with the theoretical curves of Fig. 3, the rf experimental data would ordinarily be plotted as a function of the measured peak currents. However, due to secondary emission, the measured beam currents were not those effective in the tube. Therefore, the beam current measured at the mean value of peak grid voltage was used to compare the grid-current scale with the grid-voltage scale. At higher and lower grid voltages the current was assumed to vary as $Eg^{3/2}$. Thus, in Fig. 8(a), the measured beam current was 17.5 ma at the grid voltage of 15.9 volts, and the scale of $I^{2/3}$ is assumed to be linear either side of this point. From the relative rf voltage measured at $Eg=15.9$ volts, a was determined from Fig. 2 and k calculated from (18). Using this value of k and values of a from Fig. 2, the beam currents were then calculated from (18) as a function of the measured rf voltages, yielding the long-dashed curve of Fig. 8(a). The long-dashed curve of Fig. 8(b) was calculated in the same manner.

The short-dashed curves of Fig. 8 were calculated from (20) and Fig. 2 for the case in which the electrons strike the pole faces. For each curve, as was done in Fig. 3, the output of the magnetron into a matched load was used to determine P_M, from Fig. 2, for use in (20). Since the theoretical curves do not intersect in Fig. 8, no deformation of the experimental characteristics at the low grid voltages would be expected; none is apparent in the figure. In general, then, the experimental characteristics are a reasonable reproduction of those calculated, except that the former show slightly less evidence of saturation.

Television Resolution Pattern

As an over-all qualitative check of the measured bandwidth and linearity, a television resolution pattern was reproduced by the system. The monitor derived its signal from a diode inserted in the rf line near the load. The picture was free from noticeable defects.

Fig. 9 shows the final developmental equipment, contained, except for the power supplies, in one standard short rack. The solenoidal magnet and water loads are mounted on top. A cable feeds the input signal to the modulator, which has a bandwidth of 20 Mc, recessed into the front panel.

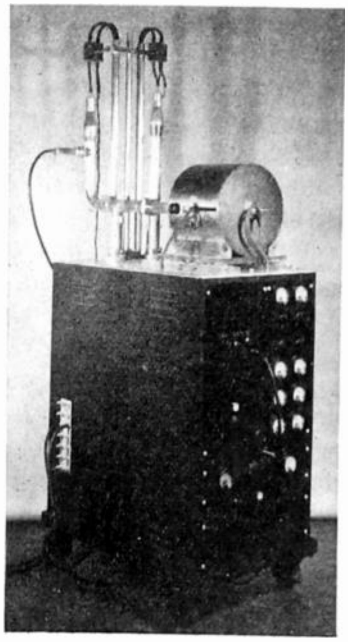

Fig. 9—A front view of the rack containing the amplitude-modulation system. All additional power supplies may be contained in another rack of the same size.

VI. Conclusions

This new electronic method for the amplitude modulation of magnetrons makes use of a beam of electrons which presents a pure conductance to a resonant cavity. The cavity is, in turn, coupled to the magnetron and the load in such a manner that grid control of the beam current varies the system power output by shunting the load and by altering the efficiency of the magnetron. The bandwidth of the system is at least 20 Mc, the depth of voltage modulation is 85 per cent, the linearity is reasonably good, and the frequency variation during the modulation cycle is about ±15 kc at the carrier frequency of 900 Mc. The average carrier frequency could be controlled by use of frequency-modulation beams in the oscillator used; the frequency variation during the modulation cycle could be decreased in the same manner.

The peak power output obtained from the present developmental system is 400 to 500 watts, and is limited by failure of the envelope of the type of modulation tube so far employed. Proposed tubes of other designs should raise the power output to several kilowatts. Scaling of the modulation tubes to higher frequencies would offer no serious difficulties, since, for a given geometry, the power absorbed by the beam is proportional to the square of the frequency.

A television picture, free from obvious defects, has been produced, it is believed for the first time, by the electronic amplitude modulation of a magnetron.

VII. Acknowledgments

The authors acknowledge their indebtedness to J. J. Stickler and J. J. Thomas, who assisted with the experimental work during a portion of this project; to M. A. Colacello, who helped with the construction of the modulation tubes; and particularly to L. S. Nergaard, who proposed the type of resonant cavity employed, and who contributed much to the solution of the nonlinear problem presented by the theoretical modulation characteristic.

Correction

The following error in the paper, "Square-wave analysis of compensated amplifiers," by Philip M. Seal, which appeared in the January, 1949, issue of the Proceedings of the I.R.E., has been called to the attention of the editors by the author.

Equation (30) on page 52 should read as follows:

$$\text{relative } e = \frac{A}{\cosh \frac{\pi}{2}\frac{f_2}{f}} \epsilon^{(\pi/2)(f_2/f)(1-2r)} - \frac{A-1}{\cosh \frac{\pi}{2}\frac{f_3}{f}} \epsilon^{(\pi/2)(f_3/f)(1-2r)}. \qquad (30)$$

Measured Noise Characteristics at Long* Transit Angles

NORMAN T. LAVOO†, MEMBER, IRE

Summary—Tests made on diodes at 3,000 Mc and transit times of over one radio-frequency cycle indicate that the space-charge reduction of noise is of the order of 10 to 1. This was true of diodes having tungsten, thoriated-tungsten, and oxide emitters. The magnitude of the noise, as well as the variation with transit time, was found to agree qualitatively with the present theory.

I. Introduction

THIS PAPER presents several experimental investigations concerning the effects of space charge and transit time on noise in microwave tubes. One of the questions which it was hoped would be answered by these investigations was the relative importance of the noise contribution due to electrons circulating in the region between the cathode and the potential minimum. Should the reduction of noise due to space charge prove to be inferior at long transit angles, it would be necessary to differentiate between the reduced efficiency of the mechanism of the potential minimum smoothing of noise at long transit angles and any new noise sources that may arise within the potential minimum. On the basis of the results of this investigation it does not seem worth while to separate these effects, since only a small difference was found between the reduction factor due to the space charge normally obtained at low frequencies and that measured on several diodes at 3,000 Mc with transit angles of over one cycle.

If the output of a 3,000-Mc grounded-grid rf amplifier using a parallel-plane triode is fed into a sensitive receiver, considerable noise is indicated at the output of the receiver. The order of magnitude of this noise referred to the input of the amplifier is 20 to 25 db above $KT\Delta F$ for a lighthouse triode of the 2C40 type. Further, if the input circuit of the rf amplifier is detuned, very little change is ordinarily noticed in the noise output. This would indicate that most of the noise measured between the grid and plate of the grounded-grid amplifier at this frequency is the result of noise currents inherent in the electron stream as it leaves the potential minimum and of a law-input impedance to the amplifier. This has been observed on triodes and tetrodes with a noise figure as low as 15 db/$KT\Delta F$.

On some of these tubes the noise output versus the filament voltage was noted; therefore, the effect of space charge could be observed. On the tubes with large cathode areas, such as the 2C39, sizable reduction factors were noted, but they indicated a decided frequency dependence. For instance, at 1,000, 3,000, and 3,600 Mc the factors were 25 to 1, 6 to 1, and 2.4 to 1, respectively. Tubes with considerably smaller cathodes and with very close spacing and fine grids indicated reduction factors of the order of 3 to 1 at 3,000 Mc.

These measurements of reductions in noise caused by varying the cathode temperature do not give a clear picture of the various factors that make up the reduction factor. For instance, as one varies the cathode temperature, the potential minimum moves over a very considerable fraction of the distance between the cathode and the grid. In addition, the field at the cathode is far from uniform because of the nonuniformity of the emitter and the fact that the grid is not infinitely fine. Finally, even though the output circuit was deliberately detuned to present a fairly low impedance to the electron stream, the changing admittance due to the electron stream is another source of complication. For these reasons it was deemed advisable to build a tube that would eliminate as many of these undesirable variables as possible. The choice of structure was that of a diode built in the form of a coaxial transmission line. This form is similar to one used by Kompfner et al.[1] They report measurements, made on a tungsten-filament diode, of the effect of transit time and space charge. In this report, verification of their results will be presented, as well as an extension to the cases where thoriated-tungsten and oxide emitters are used.

II. The Diode and Measurement Apparatus

The form of the diode is shown in Fig. 1. The filament was 0.004 inch in diameter and approximately 4

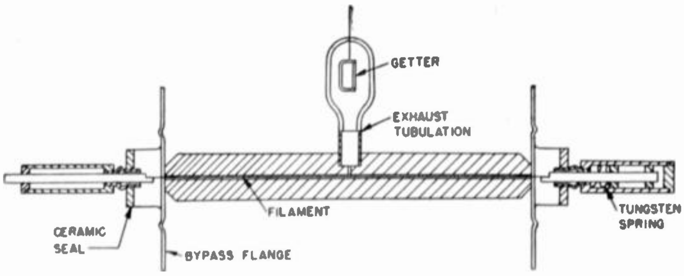

Fig. 1—Cylindrical, transmission-type noise diode.

inches in length. The anode was formed by cutting a 0.080-inch hole through the ½-inch diameter anode

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[1] K. Kompfner, J. Hatton, E. E. Schneider, and L. Dresel, "The transmission-line diode as a noise source at centimetre wavelengths," *Jour. I.E.E.* (London), vol. 93, pt. 3A, no. 9; March–May, 1946.

block. The hole through the anode was accurately drilled and then swaged to the proper size. The filament was supported at the ends by a low-loss ceramic seal and was held taut by means of a tungsten spring in one end. All tubes were given a twenty-four-hour bake-out at 450°C, and were activated while still on the pump. The vacuum was of the order of 10^{-8} millimeter of mercury under these conditions. After activation a Batalum getter was flashed and the tube sealed off.

Each end of the tube was matched by means of a tapered section and a double stub. Fig. 2 shows the

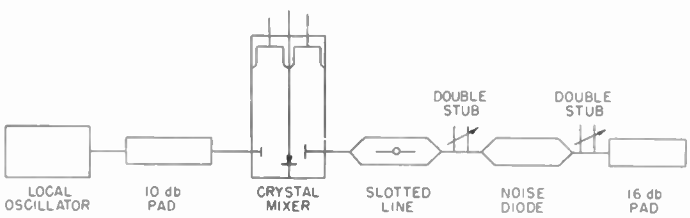

Fig. 2—A block diagram of the noise measurement setup.

apparatus involved in the noise measurements. One end of the tube is fed into a slotted section and then into the crystal mixer. The other end is matched into a 16-db pad. The diode in the matched condition was found to be broad enough to give full output at intermediate frequency either side of the local-oscillator frequency. The local oscillator was tuned to 10 cm, and the receiver responded to signals at approximately 9.9 and 10.1 cm.

Fig. 3 shows the bandwidth of the crystal mixer and diode. These standing-wave ratios were taken with

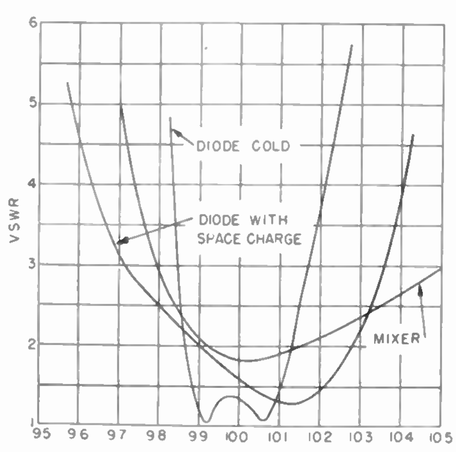

Fig. 3—Bandwidths of the oxide-emitter diode and crystal mixer.

the slotted line as indicated, looking first into the mixer and then into one end of the diode. The diode bandwidths shown are for an oxide emitter. As indicated, the bandwidth continually changes as one goes from diode cold to full space-charge limited with these two conditions as extremes. Depending on the matching, one bandwidth could be made less or more than the other. The attempt in each case was to keep the standing-wave ratio at a minimum over the widest band under all conditions. It was felt that a standing-wave ratio less than two or three in voltage would not give rise to enough reflection to invalidate the results. The condition for minimum standing-wave ratio over the widest band corresponds to minimum standing-wave ratio within the coaxial line part of the diode. The settings of the matching elements were fairly broad, and were very symmetrical at each end of the diode. The condition of match at both ends was found to be superior to the condition of match on one end and complete reflection on the other. With the match condition on both ends, the crystal-mixer current was much more constant under all operating conditions of the diode. This, in addition to the fact that the attenuation correction is much simpler, dictated the use of the fully matched condition.

III. Attenuation Through the Diode

Fig. 4 shows the insertion loss through the oxide emitter diode at 3,000 Mc. It also indicates the static

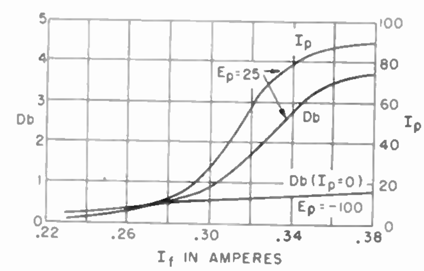

Fig. 4—Insertion loss of the oxide-emitter diode.

characteristic and the insertion loss measured with the filament at various temperatures, but with the anode biased to a high negative voltage. The plate voltage is indicated as 25 volts; but in this, as well as the subsequent characteristics to be described when voltage is to be a fixed parameter, it is given a slight modification to compensate for the filament drop. For instance, the anode voltage measured with respect to one end of the filament is adjusted so that the same current would be drawn, assuming space-charge limitation of current, as the desired voltage applied to a diode with an unipotential cathode.[2] This works out so that the voltage is very nearly the voltage of the anode, with respect to the center of the filament. The filament drop is small in the oxide-emitter diode, but it becomes more important in the case of the tungsten emitter.

Fig. 5 presents the attenuation for various fixed filament temperatures, with anode voltage as the variable. In this curve, the constant loss at a given temperature due to the filament being hot has been subtracted.

[2] W. G. Dow, "Fundamentals of Engineering Electronics," John Wiley and Sons, Inc., New York, N. Y., 1937; p. 112.

Hence, the curves present only the loss due to electron flow across the diode. The curves depicted in Figs. 4 and 5 are for an oxide emitter, but the curves for thoriated tungsten and pure tungsten are quite similar, except for the required filament currents.

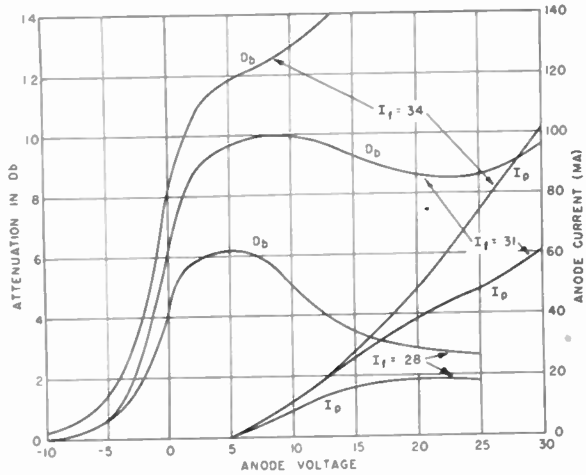

Fig. 5—The loss in the oxide-emitter diode with dissipation due to hot conductor subtracted.

IV. Effect of Transit Time on Noise

A number of papers have been written on the analysis of noise in the presence of transit time.[3-5] Any of three or four approaches to this problem show that the reduction due to transit time between parallel planes is given by the expression

$$F = \frac{4}{\theta^4}\left[2 + \theta^2 - 2(\cos\theta + \theta\sin\theta)\right]$$

where θ is the transit angle in radians. This expression has been shown by Rack to apply in the presence of space charge, as well as to the case of temperature-limited emission.

For the reduction factor due to transit time between coaxial cylinders, the mathematical expression cannot be integrated in closed form, so that one is forced to numerical integration or a series-type solution.[4] The result to be integrated is

$$F = \frac{1}{(\ln a/b)^2}\left[\left\{\int_1^{a/b}\cos\theta_r\frac{d(r/b)}{r/b}\right\}^2 + \left\{\int_1^{a/b}\sin\theta_r\frac{d(r/b)}{r/b}\right\}^2\right]$$

where θ_r is the transit angle required for an electron to transit from the inner cylinder, $r=b$, to the radius r which is part way out to the outer cylinder, $r=a$.

$$\theta_r = \frac{\omega(a-b)\sqrt{\ln a/b}}{\sqrt{\frac{2eV}{m}}(a/b-1)}\int_1^{r/b}\frac{d(r/b)}{\sqrt{\ln r/b}}.$$

This reduction factor for the particular value of $a/b=20$, which is the case of the diode described in this report, is plotted in Fig. 6. It is plotted versus the

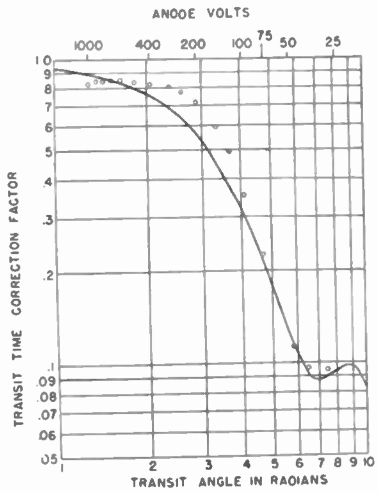

Fig. 6—The variation of noise with transit angle from a temperature-limited diode.

transit angle required for transit from the inner to the outer cylinder. Also shown in Fig. 6 is the experimental curve taken on the diode with 5 ma plate current and a frequency of 3,000 Mc. The noise plotted is just that due to the flow of electrons across to the anode; that part due to the filament temperature was subtracted. It is necessary to subtract the thermal noise of the filament, in order that the noise be directly proportional to the plate current. The experimental and theoretical curves agree fairly well, except for the region of low-transit angles. To obtain these low angles the voltage necessary was quite high, and it is the effects of this high voltage that are believed to be responsible for the discrepancy.

V. Space-Charge Cushioning of Noise at Long Transit Angles

At the lower frequencies, accurate calculations and measurements have been made of the influence of space charge on noise. Although the calculations were made for the parallel-plane case, experiments indicate that even for the cylindrical case the agreement is excellent.[6] The factor by which shot noise is reduced at low frequencies is of the order of 15 to 1, varying more or less depending upon the operating conditions. A value

[3] S. Ballantine, "Schrot-effect in high frequency circuits," *Jour. Frank. Inst.*, vol. 206, p. 159; August, 1928.

[4] Eberhard Spenke, "Die Frequenzabhangigkeit der Schroteffektes," *Wissenschaftliche Veroffentlichungen aus den Siemens-Werken*, vol. 16, pp. 127–136; October, 1937.

[5] A. J. Rack, "Effect of space charge and transit time on the shot noise in diodes," *Bell Sys. Tech. Jour.*, vol. 17, pp. 592–619; October, 1938.

[6] D. O. North, "Fluctuations in space charge limited currents at moderately high frequencies," *RCA Rev.*, vol. 5, pp. 106–124; July, 1940.

very similar to this one has been reported on a tungsten-filament diode at long transit angles.[1] The reduction of noise, as a result of space charge, for a tungsten, thoriated-tungsten, and a double-carbonate emitter are presented in Fig. 7. In each case, the reduction factor is quite sizable and almost comparable with representative values for low frequencies.

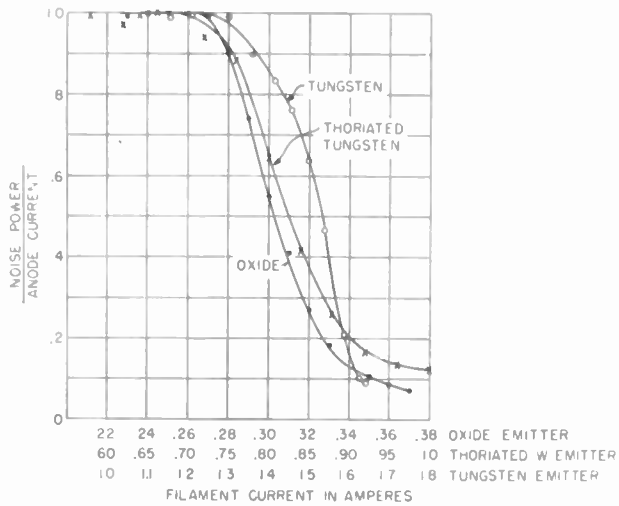

Fig. 7—Cushioning of noise due to space charge in tungsten, thoriated-tungsten, and double-carbonate type of emitters. Anode voltage-held approximately constant at 25 volts.

Unfortunately, space charge, through its variations of the potential minimum, is not the only influence affecting the noise in Fig. 7. Because of the voltage drop along the filament, the transit angle will be different for each end of the diode, as compared with the middle section. In addition to the transit angle differing for different portions of the diode, at a given part of the diode the changing space charge in itself results in a change in transit angle which, in turn, influences the magnitude of the noise. In going from temperature-limited to complete space-charge-limited emission, the transit angle is increased by a factor of slightly more than 30 per cent for the particular diode described in this report.[7] The average transit angle in the temperature-limited-emission region shown in Fig. 7 amounts to about 7 radians, and increases about 30 per cent as the space charge is increased. Fortunately, as can be noted from Fig. 6, this transit-time increase causes very little charge in the noise because of the relatively flat portion of the transit-time correction curve in the region of 7 to 10 radians. Another factor that must be considered is the attenuation to a signal at the frequency of measurement, 3,000 Mc, offered by the diode. From the experimental curves for the magnitude of the attenuation, of which Fig. 4 is an example, it is possible to calculate the noise power in the absence of attenuation. It is the noise power in the absence of attenuation that is of interest, since a comparison with the low frequencies is desired. Since the attenuation is greatest at full space charge, if a correction for attenuation were not made the reduction factors would appear considerably greater.

The factor by which the noise power coming from one end of the matched diode is reduced because of attenuation is given by

$$H = \frac{1 - e^{-0.23L}}{0.23L}$$

where L is the insertion loss of the diode in decibels measured in the matched condition. For each of the diodes, H is about 0.5 for full space charge, and practically 1 for temperature-limited operation. These corrections have been applied in the plotting of Fig. 7.

Conclusions

Because of the nonuniformity of emission across the surface of an emitter, it was felt that there probably was a deterioration in the amount of space-charge cushioning with frequency. If this were true, an oxide or thoriated-tungsten emitter would probably be much worse than a pure-tungsten emitter in this respect. However, such is not the case, as the experiments show. For all three types of emitters operating at 3,000 Mc, with the time of transit well over a complete rf cycle, the reduction due to space charge is of the order of 10 to 1.

The magnitude of noise energy in the temperature-limited condition and for low transit angles was measured and found to agree with the calculated value. Further, as the transit angle was increased, by lowering the anode voltage, the noise reduced in a way which agrees qualitatively with theory.

Finally, it should be pointed out that the results apply only to diodes. There is considerable experimental evidence indicating that a close-spaced grid reduces the magnitude of the space-charge reduction of noise at the ultra-high frequencies.

[7] F. B. Llewellyn, "Electron Inertia Effects," Cambridge University Press, New York, N. Y., 1939.

The Response of Frequency Discriminators to Pulses*

EUGENE F. GRANT†, ASSOCIATE, IRE

Summary—**An analysis is made of the time response of a simple shunt resonant circuit and the results applied to the behavior of the Round-Travis and Foster-Seeley frequency discriminators. The relationship between the parameters of the circuit are derived such that the discriminator will have only one crossover frequency in the desired frequency band for the application of pulsed signals.**

Introduction

THERE ARE MANY applications where a frequency discriminator of any of the popular types—Round-Travis or Foster-Seeley—is used for the frequency discrimination of pulses of short duration. An application of this use is the automatic frequency control of pulsed oscillators; in particular, the control of microwave oscillators with respect to maintaining a constant difference frequency with a pulse oscillator. As might be suspected, when the frequency spectrum of the pulse to be passed through the discriminator has a width which is of the same order as the pass-band of the discriminator, an anomalous behavior is presented. Specifically, if one observes the response of the discriminator to pulses of short duration as a function of the center frequency of the pulse, one may find that there are several crossover points. This type of behavior is confusing to an automatic-frequency-control circuit.

By making a time analysis of the frequency-sensitive elements of a discriminator, design formula may be derived such that the response of the discriminators will have only the required crossover for a particular pulse length within a specified band of frequencies.

There are two types of discriminators in common use: the Round-Travis, and the Foster-Seeley (Figs. 1 and 2).

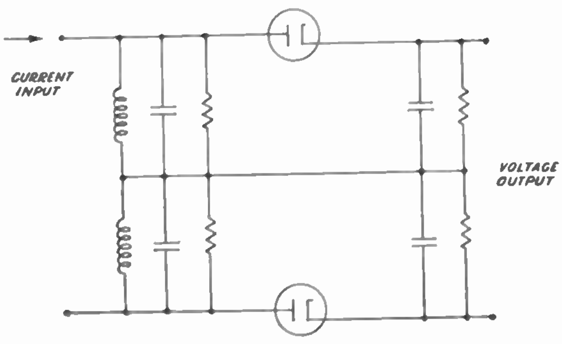

Fig. 1—Round-Travis discriminator circuit.

In the Round-Travis discriminator, use is made of the envelopes of the output of two simple resonant circuits displaced in frequency. The rectified outputs are added in opposition, so that, at a frequency equal to the mean of the individual resonant frequencies, the output is zero. On the application of an input on either side of this frequency, one side dominates, and there is an output. Since, then, the output is a superposition of two identical circuits displaced slightly in frequency, it will be sufficient to make a time analysis of only one of them.

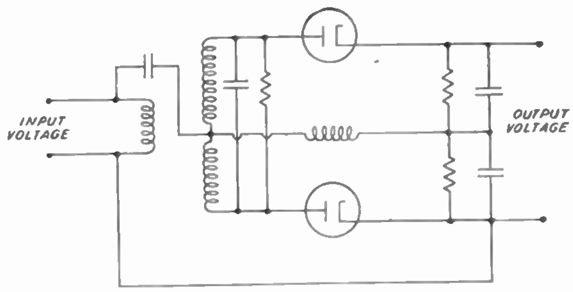

Fig. 2—Foster-Seeley discriminator circuit.

In the Foster-Seeley discriminator, use is made of the variation of the phase shift of the response of a resonant circuit as a function of frequency. The resonant circuit has a coil which is loosely coupled to a source and is center-tapped and fed from the same source. At the resonant frequency, the voltage across the coil is 90° out of phase with the voltage injected at the tap. (This is true since there are losses in the circuit.) The difference in the rectified voltages from the ends of the coil with respect to the return circuit is zero at the resonant frequency. For other frequencies, the phase shift from 90°, causing a small component of voltage to be added to one side and to be subtracted from the other, results in an output.

Glossary

$p = j\omega$

E_1 = value of voltage independent of time (or frequency)

$E(t)$ = output voltage as a function of time

$E_v(t)$ = the envelope of $E(t)$

$E(p)$ = output voltage in steady state as a function of frequency

I = value of current independent of time

$I(t)$ = applied current as a function of time

$I(p)$ = applied current in steady state as a function of frequency

ω = radian frequency $= 2\pi f$

ω_a = crossover frequency of the discriminator

ω_2 = resonant radian frequency of the high-frequency circuit of the Round-Travis discriminator

* Decimal classification: R361.217. Original manuscript received by the Institute, September 25, 1947; revised manuscript received, April 1, 1948.

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ω_3 = resonant radian frequency of the low-frequency circuit of the Round-Travis discriminator
ω_0 = radian frequency of the applied current
A = the peak amplitude of the envelope of $E(t)$
H = the peak amplitude of the envelope of $E(t)$ at *resonance*
Br = the Bromwich contour of integration. It is a path in the p plane extending from $-j\infty$ to $+j\infty$ in such a manner as to pass all the poles of the integrand on the positive real side
α = the time duration of the applied pulse
β, γ = roots of the equation $p^2+\omega_1 p/Q+\omega_1^2=0$
V = the total rectified output of the discriminator
L = the inductance of the generic circuit
C = the capeitance of the generic circuit
ω_1 = resonant radian frequency of the generic circuit $=1/\sqrt{LC}$
Q = "Q" of the generic circuit $= R/(\omega_1 L) = RC/L$
$C(p)$ = the frequency spectrum of the input pulse.
$\Delta\omega = \omega_0 - \omega_1\sqrt{1-1/4Q^2}$
$x = 2Q\Delta\omega/\omega_1$
$y = \alpha\omega_1/2Q$
$\Delta\omega_b$ = the minimum frequency deviation from ω_1 for which a departure from monotoneity exists
$\Delta\omega_c$ = the minimum frequency deviation for which an additional crossover exists.

Analysis

It will be assumed in the following analysis that the effect of the loading of the rectifiers on the resonant circuits will be that of a shunt resistance, and that the output of the diodes is a voltage which is the envelope of the ac applied to them. In general, this will be approximately true. At least for an ideal linear diode, the output is proportional to the input. Specifically, then, the analysis will amount to finding the envelope of the voltage across a shunt resonant circuit resulting from the application of a pulse of ac.

In terms of the steady-state behavior of a shunt-resonant circuit, the relationship between the driving current and the voltage across the circuit in terms of its resonant radian frequency and Q is

$$E(p) = \frac{RI(p)}{1 + Q(p/\omega_1 + \omega_1/p)}. \tag{1}$$

By the method of the Laplace transform, the instantaneous voltage resulting from the application of a pulse of current of duration α may be derived. (See Appendix I.) In general, the phase of the alternating wave within the pulse will affect the shape of the resulting voltage. However, the parameters of discriminators generally are such that this effect is small enough to be neglected. (The magnitude of this discrepancy is noted in Appendix I.) With this approximation, then, the peak amplitude of the envelope of the response reduces to a tractable expression. The peak of the response is chosen for discussion for two reasons. For automatic-frequency-control work, the peak response of the discriminator is usually chosen to operate the control mechanism. For any other use, since the behavior is assumed linear, any other important quantity will be proportional to the peak amplitude. (As, for example, the integral of the response with respect to time.)

Then, subject to the restriction that $1/(4Q)$ is much smaller than unity and the region of frequency deviation is such that $\Delta\omega/\omega_1$ is smaller than unity, the following expression for the peak voltage out of the rectifier is valid:

$$E_s(\alpha) = \sqrt{\frac{1 - 2e^{-\alpha\omega_1/2Q}\cos\alpha\Delta\omega + e^{-\alpha\omega_1/Q}}{1 + \left(\frac{2Q\Delta\omega}{\omega_1}\right)^2}}. \tag{2}$$

This expression has some rather interesting ramifications. As one would expect, if the pulse length were very long compared to the proper circuit parameter, then the response of the system should be that of a simple resonant circuit. This is easily seen from the above expression. As the quantity $\alpha\omega/2Q$ is made very large, the, exponentials disappear, leaving the expression

$$\sqrt{\frac{1}{1 + \left(\frac{2Q\Delta\omega}{\omega_1}\right)^2}} \tag{3}$$

which is the approximation to a resonance curve in steady state for frequencies near resonance.

If, on the other hand, the circuit Q were made very large, one might expect that the resulting peak response as a function of frequency would be proportional to amplitude of the frequency spectrum of the pulse. This is seen to be the case by taking the limit of the expression for the peak voltage as Q gets indefinitely large, yielding

$$\left(\frac{\alpha\omega_1}{2Q}\right)\left|\frac{\sin(\alpha\Delta\omega/2)}{(\alpha\Delta\omega/2)}\right|. \tag{4}$$

There is a factor in the expression which vanishes with a large Q. It is seen, then, that as the circuit Q gets very low, the frequency behavior of the circuit approaches that of the steady-state case. On the other hand, if the circuit Q is very high, the resulting output depends on the time behavior within the pulse itself, and the output is of the form of its frequency spectrum. This latter characteristic is not a desirable one to have for the elements of the frequency discriminator, since the form of the responses is independent of the resonant frequency. The fundamental problem is to arrive at the appropriate circuit Q such that the peak response of the circuit is nearly that of the steady-state resonant response and shall be monotonically decreasing from the peak output (at resonance) over the desired band of operation. If there should be a response which has a succession of maxima and minima, then it could be expected that there will be several crossover points as a function of frequency when two such circuits are used in the discriminator system. (Fig. 3 portrays a discriminator that has this difficulty.)

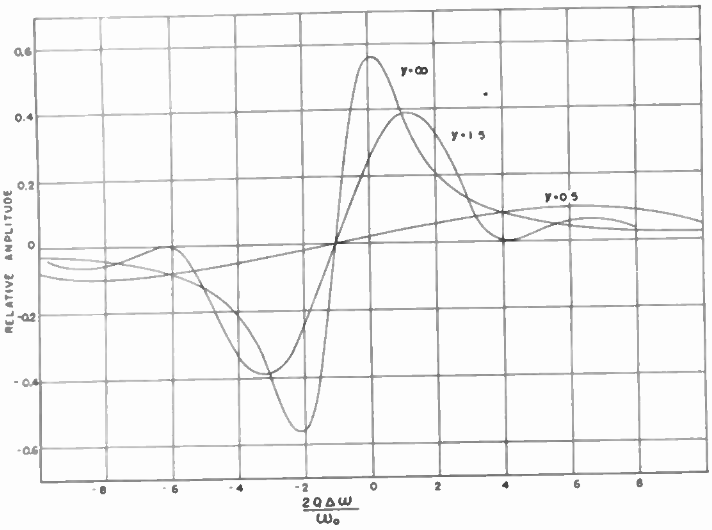

Fig. 3—Response of a discriminator to pulses; Round-Travis type with resonant-circuit spacing of $2Q\Delta\omega/\omega\alpha = \alpha$.

There are at least two approaches to the design of a proper circuit given the conditions under which it must operate. One is to plot the value of (1) for various values of the parameters involved and then select the appropriate curve that would provide the proper operation. This process can be simplified with the substitution of

$$x = \frac{2Q\Delta\omega}{\omega_1} \quad \text{and} \quad y = \frac{\alpha\omega_1}{2Q}. \tag{5}$$

x then has the character of a frequency-deviation parameter, and y is a dimensionless quantity involving the bandwidth of the circuit and the pulse length.

Rewriting (1), it follows that

$$E_v(\alpha) = \sqrt{\frac{1 - 2e^{-y}\cos xy + e^{-2y}}{1 + x^2}}. \tag{6}$$

A discussion of this equation in terms of the x's and y's then will apply to all circuits with these parameters. Figs. 4, 5, and 6 plot $E_v(\alpha)$ over a convenient range of x with several well-chosen values of y as a parameter. Fig. 7 is the curve of a discriminator designed by this method.

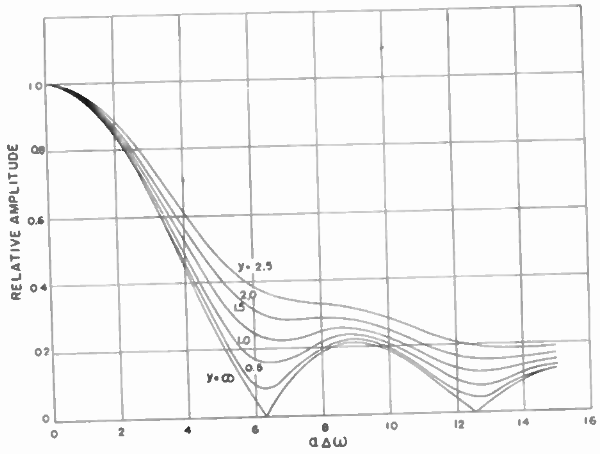

Fig. 5—Normalized resonant-circuit response to pulses.

The other method is perhaps more erudite: That is, differentiate the function $E_v(\alpha)$ with respect to $\Delta\omega$ and deduce the circuit Q for the condition that the derivative shall be nonzero in the desired range of operation (other than at the peak of the response at resonance). A nonzero derivative insures that the function shall be monotonic.

Let, then, $\Delta\omega_b$ be the point at which it is permissible that the slope of the curve be zero. Under the condition

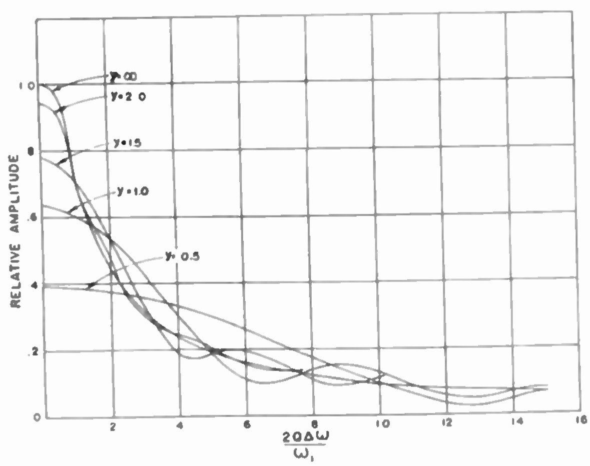

Fig. 4—Resonant-circuit response to pulses.

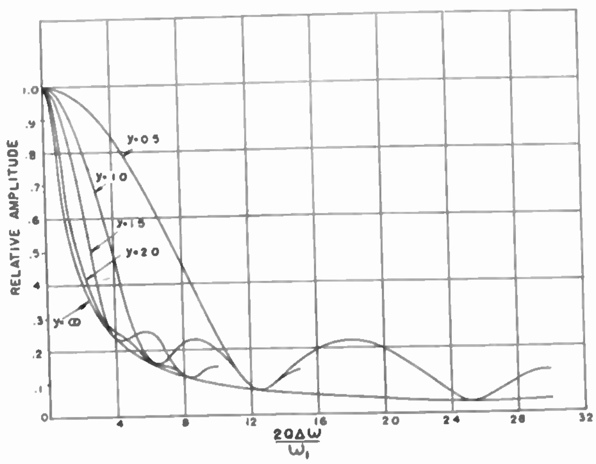

Fig. 6—Normalized resonant circuit response to pulses.

that this point is removed from the 3-db point of the resonance curve, an approximation may be made which leads to a rather simple solution for the appropriate Q. If the conditions are such that this approximation may not be made, the equation to be solved remains transcendental with its inherent difficulties. Then, given the pulse lengths, center frequency, and the frequency over which monotoneity is desired, the following formula gives the circuit Q which is necessary. Equation (7) is subject, of course, to the restriction that the quantity x corresponding to $\Delta\omega_b$ must be greater than unity. This fact must be determined after the Q is calculated.

$$Q \cong \frac{\alpha\omega_1}{2 \operatorname{arc\,cosh}\left[\frac{\alpha\Delta\omega_b}{2}\sin\alpha\Delta\omega_b + \cos\alpha\Delta\omega_b\right]}. \qquad (7)$$

Although, for discriminator action, it is not essential that the individual resonant curves be monotonically decreasing from the resonant point, it is usually more convenient to make them that way, for the slope of the over-all discriminator curve may be maximized or made to meet other conditions arbitrarily.

The slope of the output versus frequency in the Round-Travis discriminator may be maximized rather easily at the crossover point in steady state for a fixed Q and a variable ω_2 and ω_3. For pulsed operation, the Q has been established subject to other requirements. While the rectifiers are connected so that the voltage outputs subtract, the rates of change of the voltage in the two circuits add in the region near the crossover point. For maximum slope, the spacing between the resonant frequencies should be such that the points of maximum slope coincide. It is easily shown by differentiation of the resonance curve that the frequency spacing should be $2/Q$ multiplied by the crossover frequency.

The previous results may be heuristically applied to the Foster-Seeley frequency discriminator as well. In this type of discriminator there is only one resonant circuit, and the frequency-discriminator action results from the addition of a voltage which is 90° out of phase with the voltage of the circuit at resonance. The over-all behavior is nearly that of the Round-Travis discriminator. (See Appendix II for the mathematical discussion.) In this case, it is easier to treat the additional crossover points directly, yielding the following formula for the Q of the coil subject to the approximation that the resulting x_c shall be large compared to unity:

$$Q = \frac{\alpha\omega_a}{2 \log_e \cos \alpha\Delta\omega_c}. \qquad (8)$$

Conclusions

A discussion has been given and curves portrayed of the pulsed behavior of a simple resonant circuit with a view to finding the limitations of the parameters so that the discriminator would approximate steady-state behavior upon the application of pulsed signals. It was shown that, when the Q of the circuit was very low, the circuit had an output which varied in the same manner as if it were in steady state to pulses. When, on the other hand, the circuit Q was very high, the simple resonant circuit had an output as a function of frequency which was proportional to the spectrum of the signal pulse. The entire question then resolved itself to a determination of parameters such that the resonant circuit does not exhibit any behavior as a function of frequency which would cause additional crossover points when assembled into a discriminator. The simplest approach to this is to plot the behavior of the circuit as a function of frequency with a parameter and then select the curve

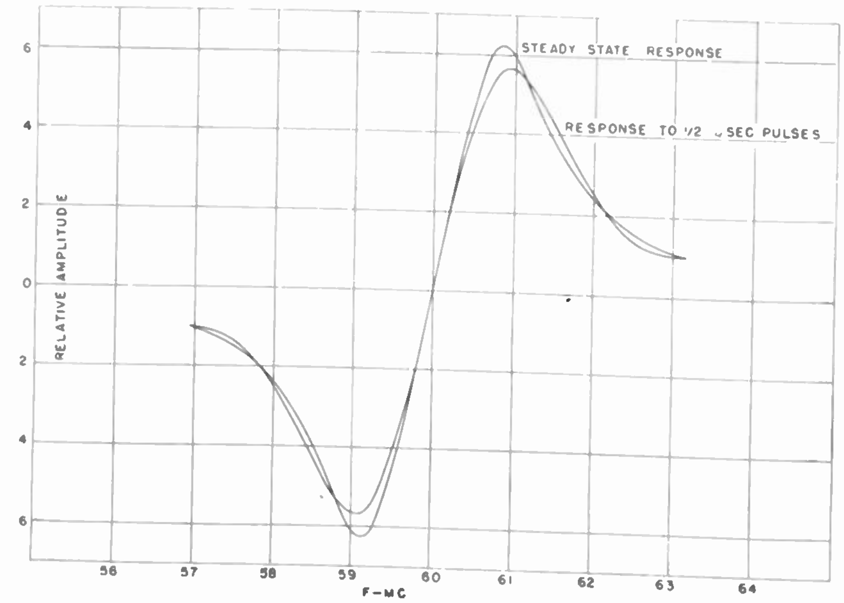

Fig. 7—Response of a discriminator to pulses; Round-Travis type with resonant-circuit frequency spacing of 1.75 Mc and a Q of 60.

which would give the most satisfactory results. For most purposes, this is a sufficiently accurate method. If this method is not accurate, a more accurate formulation of the slope of the response can be represented analytically, and under certain approximations the circuit Q can be solved for which will give a monotonic-decreasing response in the desired range of operation.

Appendix I

The response of a simple resonant circuit to a single current pulse of frequency ω_0 will now be derived. It will be assumed that all of the dissipative loading on the circuit can be considered as an equivalent resistance R.

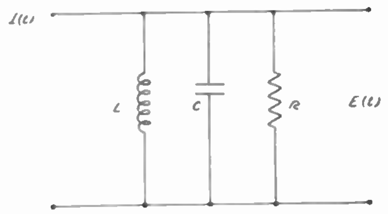

Fig. 8—The generic circuit.

It is easily seen from Fig. 8 that the relationship between the input current $I(p)$ and the output voltage $E(p)$ in steady state is

$$\frac{E(p)}{I(p)} = \frac{1}{1/R + Cp + 1/Lp}. \tag{9}$$

However, it is more convenient to discuss the behavior in terms of R, Q, and ω_1, instead of R, L, and C; therefore, by substitution, it follows that

$$\frac{E(p)}{I(p)} = \frac{R}{1 + Q(p/\omega_1 + \omega_1/p)}. \tag{10}$$

The methods of the Laplace transform may be applied to derive the response of the network to a pulse of current. For a current pulse of amplitude $1/R$, duration α, and center frequency ω_0, the frequency spectrum $G(p)$ is

$$G(p) = \frac{1}{R}\int_0^\alpha e^{p_0 t} e^{-pt} dt. \tag{11}$$

By integration, it is seen that

$$G(p) = \frac{1}{R}\,\frac{1 - e^{\alpha(p-p_0)}}{p - p_0}. \tag{12}$$

The response of the network is, then,

$$E(t) = \frac{1}{2\pi j}\int_{Br} \frac{1}{1+Q(p/\omega_1+\omega_1/p)}\,\frac{1-e^{\alpha(p-p_0)}}{p-p_0}\, e^{pt}dp. \tag{13}$$

It is to be noted that the real part of $E(t)$ will be the response to a cosine wave of current input, and the imaginary part is the response to a sine wave of current input. Simplifying (5) further,

$$E(t) = \frac{\omega_1}{Q2\pi j}\,\frac{p[1 - e^{\alpha(p-p_0)}]e^{pt}dp}{(p - p_0)(p - \gamma)(p - \beta)} \tag{14}$$

where

$$\beta = \frac{-\omega_1}{2Q} + j\omega_1\sqrt{1 - 1/(2Q)^2}$$

and

$$\gamma = \frac{-\omega_1}{2Q} - j\omega_1\sqrt{1 - 1/(2Q)^2}. \tag{15}$$

This integral may be evaluated by Cauchy's integral theorem, or it may be found in the literature.[1]

$$E(t) = \frac{\omega_1}{Q(\gamma - \beta)}\left\{\frac{\beta[1 - e^{\alpha(p_0-\beta)}]e^{\beta t}}{p_0 - \beta} - \frac{[1 - e^{\alpha(p_0-\beta)}]e^{\gamma t}}{p_0 - \gamma}\right\}. \tag{16}$$

Equation (16) is valid for $t \geqq \alpha$. Simplifying further, letting $\Delta\omega = \omega_0 - \omega_1\sqrt{1 - 1/4Q^2}$,

$$E(t) = e^{-\omega_1 t/2Q}\left[1 + \frac{j}{\sqrt{4Q^2 - 1}}\right]\left\{\frac{1 - e^{\omega_1\alpha/2Q}e^{j\alpha\Delta\omega}}{1 + j\dfrac{2Q\Delta\omega}{\omega_1}}e^{j\omega_1 t} - \frac{\sqrt{4Q^2 - 1} - j}{\sqrt{4Q^2 - 1} + j}\,\frac{1 - e^{\omega_1\alpha/2Q}e^{j\alpha 2\omega_0 - \alpha\Delta\omega}}{1 + \dfrac{j2Q(2\omega_0 - \Delta\omega_1}{\omega_1}}e^{-j\omega_1 t}\right\}. \tag{17}$$

This expression is difficult to interpret as it stands. If approximations are made that Q is large and that $\Delta\omega/\omega_1$ is small, it reduces to a more tractable expression.

$$E(t) \cong \frac{1 - e^{\omega_1\alpha/2\alpha}e^{j\alpha\Delta\omega}}{1 + j\dfrac{2Q\Delta\omega}{\omega_1}}\, e^{j\omega_1 t}e^{-\omega_1 t/2Q}. \tag{18}$$

Since the rectifiers may be thought of as providing an output equal to the time envelope (of course, the pass band of the output filters must be sufficiently great), the direct-voltage output will then be approximately equal to

$$E_v(t) \cong \left|\frac{1 - e^{\omega_1\alpha/2Q}e^{j\alpha\Delta\omega}}{1 + j\dfrac{2Q\Delta\omega}{\omega_1}}\right| e^{-\omega_1 t/2Q}. \tag{19}$$

It may be shown that this expression is roughly within $1/4Q$ of the actual envelope of (19) for frequencies near the resonance of the circuit.

It is interesting to note that by this process the result is now independent of the use of a cosine wave or a sine wave for the pulse. The difference in the responses to the cosine and sine wave was contained in the term which was dropped.

[1] G. A. Campbell and R. M. Foster, "Fourier integrals for practical applications," Bell System Monograph, pairs 453 and 207; 1932.

Equation (19) is valid for $t \geqq \alpha$, but it is a simple exponential function of time, so it may as well be discussed at its maximum amplitude of $t=\alpha$. Usually, it is this amplitude which is the desired one. Simplifying further,

$$E_v(\alpha) = \sqrt{\frac{1 - 2e^{-\alpha\omega_1/2Q}\cos\alpha\Delta\omega + e^{-\alpha\omega_1/Q}}{1 + \left(\frac{2Q\Delta\omega}{\omega_1}\right)^2}}. \quad (20)$$

$E_v(\alpha)$ may also be expressed in terms of the hyperbolic functions.

$$E_v(\alpha) = \sqrt{\frac{2e^{-\alpha\omega_1/2Q}\left(\cosh\frac{\alpha\omega_1}{2Q} - \cos\alpha\Delta\omega\right)}{1 + \left(\frac{2Q\Delta\omega}{\omega_1}\right)^2}}. \quad (21)$$

Appendix II

The problem is to find the Q which will cause the frequency response to be monotonically decreasing out to $\Delta\omega_b$. To do this it is expedient to find the first zero of the derivative of the resonance curve with respect to frequency. Of course, there will be a zero at the center frequency and at infinite frequency, but these will be suppressed. Differentiating the response with respect to $\Delta\omega$ by conventional methods results in

$$\frac{dE}{d\omega} = \frac{Q}{\omega_0} E_v(\alpha)\left\{\frac{y \sin xy}{\cosh y - \cos xy} - \frac{2x}{1+x^2}\right\}. \quad (22)$$

Setting $dE/d\omega = 0$ results in

$$\frac{y \sin xy}{\cosh y - \cos xy} = \frac{2x}{1+x^2}. \quad (23)$$

For most applications, the band of operation will be large compared to the 3-db bandwidth of the resonant circuit. (The 3-db bandwidth corresponds to x equal unity.) Hence, the approximation may be made that x^2 is much greater than unity. This simplification greatly simplifies (23). Ignoring unity as compared to x^2, it is seen that

$$\frac{\sin xy}{\cosh y - \cos xy} \cong \frac{2}{xy} \quad (24)$$

or

$$\cosh\frac{\alpha\omega}{2Q} \cong \frac{\alpha\Delta\omega_b}{2}\sin\alpha\Delta\omega_b + \cos\alpha\Delta\omega_b. \quad (25)$$

Hence,

$$Q \cong \frac{\alpha\omega_1}{2\ \mathrm{arc\ cosh}\left[\frac{\alpha\Delta\omega_b}{2}\sin\alpha\Delta\omega_b + \cos\alpha\Delta\omega_b\right]}. \quad (26)$$

In the event that the approximation cannot be made, the original equation would have to be solved in complete form.

Appendix III

The Foster-Seeley circuit may be treated in a somewhat similar manner to that of the Round-Travis discriminator. The resonant circuit may be considered to be fed from a loosely coupled transformer (in order to apply the preceding theory) with the injection from a frequency-insensitive circuit. Due to the nature of the coupling, the resonant circuit will be 90° out of phase with the injection (in steady state at resonance). Let P be the amplitude of the injection compared to one-half the resonant circuit voltage. Then, since the rectifiers supply the difference of the absolute magnitudes,

$$E_{out} = \left|\frac{e^{-y} - e^{jxy}}{1 + jx} + jP\right| - \left|\frac{e^{-y} - e^{jxy}}{1 + jx} - jP\right|. \quad (27)$$

E_{out} will have a zero at $x=0$. It will have many other zeros. It will be desirable to adjust the parameter Q such that these zeros will fall outside the usable range.

$$E_{out} = \sqrt{\frac{(e^{-y} - \cos xy - Px)^2 + (P - \sin xy)^2}{1 + x^2}} - \sqrt{\frac{(e^{-y} - \cos xy + Px)^2 + (P + \sin xy)^2}{1 + x^2}}. \quad (28)$$

Equating to zero, transposing and squaring,

$$\frac{(e^{-y} - \cos xy - Px)^2 + (P - \sin xy)^2}{1 + x^2} = \frac{(e^{-y} - \cos xy + Px)^2 + (P + \sin xy)^2}{1 + x^2}. \quad (29)$$

Remove parenthesis, multiply by $(1+x^2)$, transpose and factor:

$$0 = 4Px\left(\frac{\sin xy}{x}\right) + \left(e^{-y} - \cos xy\right). \quad (30)$$

It is seen that one zero occurs at x equal zero. The location of the zeros is independent of P since it enters only as a factor. It remains to find the value of Q given α, ω_a and $\Delta\omega_c$.

$$0 = \left[\frac{\sin\alpha\Delta\omega_c}{\frac{2Q\Delta\omega_c}{\omega_a}} + e^{-\alpha\omega_a/2Q} - \cos\alpha\Delta\omega_c\right]. \quad (31)$$

If the quantity $x_c = (2Q\Delta\omega_c/\omega_a)$ is large compared to unity, then the first term may be ignored, and

$$Q = \frac{\alpha\omega_a}{2\log_e\cos\alpha\Delta\omega_c}. \quad (32)$$

A Consideration of Directivity in Waveguide Directional Couplers*

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Summary—A hypothesis is developed for the directivity characteristic of a type of waveguide directional coupler. Formulas and design curves are presented, and supporting experimental data are given.

INTRODUCTION

THE PURPOSE of this paper is to indicate the existence of mathematical relations for the directivity of a coupler, to show that these relations have a plausible physical explanation, to give supporting experimental data, and to offer generalized curves useful to the practicing designer. Although the work is based on a careful mathematical foundation (outlined in the Appendix), the emphasis is on a physical interpretation and the presentation of design technique. For a more extensive discussion of the underlying mathematical theory, reference should be made to an excellent paper by Riblet.[1]

A directional coupler is a circuit component having the distinctive characteristic of directional transmission, which is the ability to separate waves traveling in opposite directions. Since this characteristic has found application in many circuit arrangements, directional couplers are widely used and have achieved considerable importance. To minimize design effort, a useful hypothesis for the directional characteristic of directional couplers has been evolved. This hypothesis, which was found to correlate well with experimental data, is presented here as a step toward a better understanding of the operation of directional couplers in general.

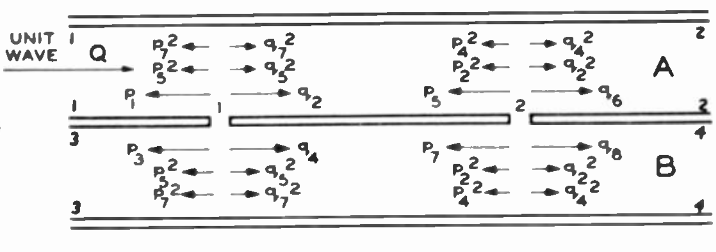

Fig. 1—Explanation of the mechanism of two-hole coupling.

The type of coupler to be considered is shown in Fig. 1, and consists of two equal rectangular guides so arranged that one of the narrow sides forms a common wall between them. Coupling between the two guides is accomplished by circular holes in the common wall.

For ease of discussion, two conventions are adopted, as follows: Waves traveling in the waveguide in the same direction as the applied wave are called *forward;* while those traveling in the opposite direction are called *backward.* The term wavelength is understood to mean wavelength within the waveguide.

To illustrate the basic principles involved, a simplified theory of operation will first be described.

Simplified Theory

Referring to Fig. 1, let a forward wave Q of unit magnitude be introduced in guide A. At orifice 1 a small portion of the wave will be coupled into guide B, where it will then produce a forward wave q_4 and a backward wave p_3. The initial unit forward wave, diminished by a negligible amount, travels on in guide A until it reaches orifice 2, where a small fraction of the wave is coupled into guide B where it again produces a forward wave q_8 and backward wave p_7. If the two orifices are of equal size and very small, the magnitudes of p_3 and p_7 as well as q_4 and q_8 can be considered as equal. If the spacing of the hole centers is one-quarter wavelength of the impressed unit wave or an odd multiple thereof, the resultant wave in guide B at any transverse plane to the left of orifice 1, will be zero, since the wave p_7 will be one-half wavelength out of phase with the wave p_3. Alternatively, the resultant wave in guide B at any transverse plane to the right of orifice 2 will be twice the magnitude of either q_4 or q_8, since q_8 will be exactly in phase with q_4.

Hence, it can be seen that a unit forward wave in guide A will produce an attenuated, resultant forward wave in guide B, but the resultant backward wave in guide B will be zero. A more comprehensive analysis of the operation of directional couplers based on the simplified theory has been given by Mumford.[2]

Definitions

For purposes of formulating definitions of important terms needed at this point, it is assumed, in the coupler shown in Fig. 1, that perfect terminations exist at the reference planes *1-1*, *2-2*, *3-3*, and *4-4*.

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[1] H. J. Riblet, "A mathematical theory of directional couplers," PROC. I.R.E., vol. 35, pp. 1307–1313; November, 1947.

[2] W. W. Mumford, "Directional couplers," PROC. I.R.E., vol. 35, pp. 160–166; February, 1947.

The following terms are now defined:

Transmission factor $T=$ amplitude ratio of transmission through an orifice.

$$\text{In Fig. 1, } T_1 = \frac{q_4}{Q} \quad \text{and} \quad T_2 = \frac{q_8}{Q}$$

$$\text{direct loss ratio} = \left[\frac{\text{power received at plane (4-4)}}{\text{power introduced at plane (1-1)}}\right]^{1/2}$$

$$\text{direct loss in db} = 20 \log_{10}\left[\frac{1}{\text{direct loss ratio}}\right]$$

$$\text{directivity ratio } D = \left[\frac{\text{power received at plane (3-3)}}{\text{power received at plane (4-4)}}\right]^{1/2} \text{ when power is introduced at plane (1-1)}$$

$$\text{directivity in db} = 20 \log_{10}\left[\frac{1}{\text{directivity ratio}}\right].$$

More Accurate Theory of Operation

In the material that follows, the discussion will deal almost exclusively with the directivity ratio of the coupler. The simplified theory that has been given is useful for purposes of explanation, but is not sufficiently accurate for exacting designs. A more complex working hypothesis is presented from which the derived directivity characteristics of directional couplers were found to agree quite closely with the observed evidence.

The equations for the directivity characteristics which have been derived are based on H. A. Bethe's theory of orifice coupling. Bethe's theory postulates a single orifice of very small diameter in a very thin wall. Physically, these conditions cannot be realized. Also, in the derivations several very small terms are neglected to arrive at results without excessively cumbersome computations. In spite of these departures from idealized conditions, there is still reasonably good agreement between theory and experimental data.

The Two-Hole Directional Coupler

When a wave passes an orifice in the common wall between two waveguides, it produces a reaction at the orifice, so that, for simplicity, the orifice can then be considered as a generator which produces equi-vector fields that travel in both directions in each guide. Referring to Fig. 1, let a forward wave of unit magnitude be introduced into guide A. It excites orifice 1 and orifice 2. As a result, orifice 1 generates forward waves q_2 and q_4 and backward waves p_1 and p_3. Likewise, orifice 2 generates forward waves q_6 and q_8 and backward waves p_6 and p_7. Since the p and q waves are identical in nature with the unit applied wave, they also will excite an orifice when passing it. Therefore, forward wave q_2 will excite forward and backward waves q_2^2 and p_2^2 in both guides A and B at orifice 2. Forward wave q_4 will excite forward and backward waves q_4^2 and p_4^2 in both guides A and B at orifice 2. Similarly, p_6 and p_7 excite corresponding q_6^2 and p_6^2 and q_7^2 and p_7^2 waves in both guides at orifice 1. This process of interaction continues indefinitely, and the resultant field at an orifice is the limit of the summation.

The magnitude of the individual waves is dependent on the number and arrangement of the holes, so that, mathematically, the general expression for each wave must be expressed implicitly in a series of simultaneous equations.

By a method given in the Appendix, the following equation for the directivity ratio D is derived:

$$D = \frac{1 + e^{j\theta}(1 + j4T)}{2}.$$

It can be seen that D is a function of two parameters: the orifice transmission factor T, and the propagation phase-shift angle θ.

Phase shift angle $\theta=$propagation phase shift of a wave traveling a distance $2S$

$$\theta = \frac{4\pi S}{\lambda g} \text{ radians} = \frac{720S}{\lambda g} \text{ degrees}$$

where

$\lambda g=$waveguide wavelength
$S=$distance from reference plane to center of hole.

The parameter θ is introduced for the purpose of deriving general expressions.

Using the above expression for directivity, and assuming values for the transmission factor T, curves of directivity ratio D versus θ can be plotted as shown on Fig. 2.

In the simplified theory previously outlined for a two-hole coupler, a point of zero directivity ratio would obviously occur at $\theta=(2n-1)$ 180° where $n=1, 2, \cdots$. However, it can be observed from the curves that actually the directivity ratio does not go to a minimum value of zero and the minimum point does not occur at

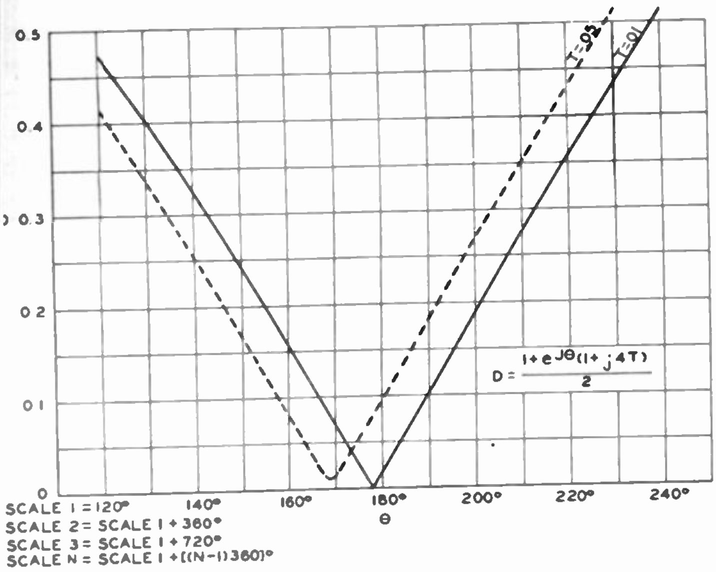

Fig. 2—D versus θ characteristics of the two-hole coupler as a function of the transmission factor.

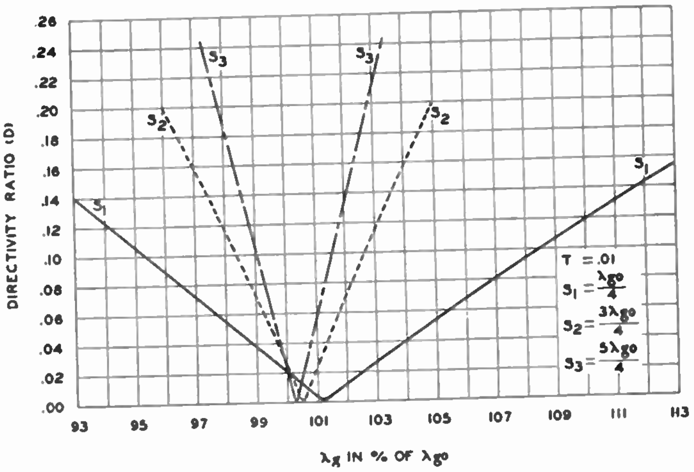

Fig. 3—D versus λg characteristics of the two-hole coupler as a function of the intra-pair spacing.

$\theta=(2n-1)$ 180°. The reason for this phenomenon will be called interaction effect.

Interaction Effect

The effect of interaction between the holes of a two-hole coupler can be determined by examining the expression for directivity of a two-hole coupler, where it is seen that, as soon as the hole size is fixed, the value of T is determined, and the expression can be rewritten as follows

$$2D = 1 + Ae^{j(\theta+\phi)}$$

where

$$\phi = \arctan 4T$$

$$A = \sqrt{16T^2+1}.$$

The directivity ratio can now be considered as the sum of a stationary unit vector and another slightly larger vector which rotates about it. Obviously, the value of D will be a minimum but not zero when the two vectors are in phase opposition.

Discussion of Curves

The curves in Fig. 2 demonstrate that the magnitude of the displacement of the minimum from 180° and the minimum directivity ratio become larger when the hole diameter, and, consequently, T, is increased.

To illustrate how a generalized D versus θ curve can be used in a specific case, the curves on Fig. 3 are derived from the curve of $T=0.01$ (Fig. 2). Three curves are derived, S_1, S_2, and S_3, for which the distances between the centers of the holes (intra-pair spacing) are assigned respective values of 1/4, 3/4, and 5/4 of a reference waveguide wavelength λg_0. Since the reference plane is assumed to be at the center of the first hole, the intra-pair spacing is also the distance S.

The important facts to note in Fig. 3 are as follows:

1. As the holes are spaced farther and farther apart in incremental odd quarter wavelengths, the effective wavelength bandwidth becomes narrower and narrower.

2. According to the simplified theory, the minimum directivity ratio should occur at λg_0, and the distance S should correspond exactly to an odd multiple of $\lambda g_0/4$. It can be seen from the curves that, as the holes are spaced farther apart in increments of odd quarter wavelengths, the point of minimum directivity ratio moves toward λg_0. From the directivity equation it can be deduced that the idealized condition is approached as an asymptote and is realized at infinite hole spacing. The phase-shift angle ϕ due to interaction becomes relatively less appreciable as the hole spacing and the resultant phase-shift angle θ become larger.

In order to correlate the theory with experimental results, refer to Fig. 4. Each point on the curves repre-

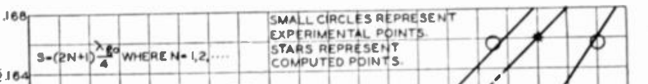

Fig. 4—Experimental data on the two-hole couplers showing interaction and proximity effects.

sented by a small circle was determined experimentally by testing a two-hole coupler of known center spacing to find the frequency of optimum directivity. The points designated by stars were computed from the equation. It can be seen that excellent correlation was obtained for hole spacings of $S=3\lambda g_0/4$ (scale B). As previously explained, the optimum directivity occurs at a longer wavelength than λg_0, which is to say that the apparent electrical centers of the holes are farther apart than the geometric centers.

An exception to the general statement above occurs when the hole centers are only a single quarter-wavelength apart. Due to an effect which, for convenience, will be termed "proximity effect," the optimum directivity was found to occur at a shorter wavelength than λg_0, which is to say that the apparent electrical centers of the holes are closer together than the geometric centers. This is shown by the experimental curve, scale A on Fig. 4.

Proximity Effect

It is evident from Fig. 4 that there is a good correlation between theory and experimental result when the holes are three quarter-wavelengths apart. However, when the holes are drilled only one quarter-wavelength apart, the theory becomes inadequate, since the experimental results show that the holes electrically appear to be somewhat closer together than they are mechanically. This discrepancy is called the "proximity effect," and arises because the theory assumes the holes to be far enough apart so that the field at one hole is undistorted by the other. Apparently, three quarters of a wavelength is a sufficient distance for the local disturbance at a hole to become negligible, but one-quarter of a wavelength is not. Since, so far as is known, no theoretical attack has yet been made on the very complex problem of describing the field conditions when two holes are closely spaced, the only alternative is to gather experimental data.

It can be seen that Fig. 4 contains the design information necessary to produce a two-hole coupler with quarter-wavelength spacing whose maximum selectivity occurs at a given frequency.

The Four-Hole Directional Coupler

It is now evident that it is not possible to achieve a low directivity ratio over a wide frequency band with the simple structure consisting of two equal-diameter round holes. The next type which has been considered is the four-hole coupler consisting of two pairs of equal-diameter holes arranged in sequence.

In order to proceed, it is necessary to define three new terms which are adopted for convenience. Refer to Fig. 5.

Intra-pair Spacing

B is the linear distance between the centers of the holes comprising a pair.

Displacement

A is the linear distance between the centers of adjacent holes of adjoining pairs.

Displacement Factor

R is the ratio of the displacement to twice the intra-pair spacing.

The following equation for the directivity ratio of a four-hole coupler is derived in the Appendix.

$$2D = \cos\theta_a + \cos\theta_b - 2T(\sin\theta_a + 3\sin\theta_b) + j6T(\cos\theta_a + \cos\theta_b)$$

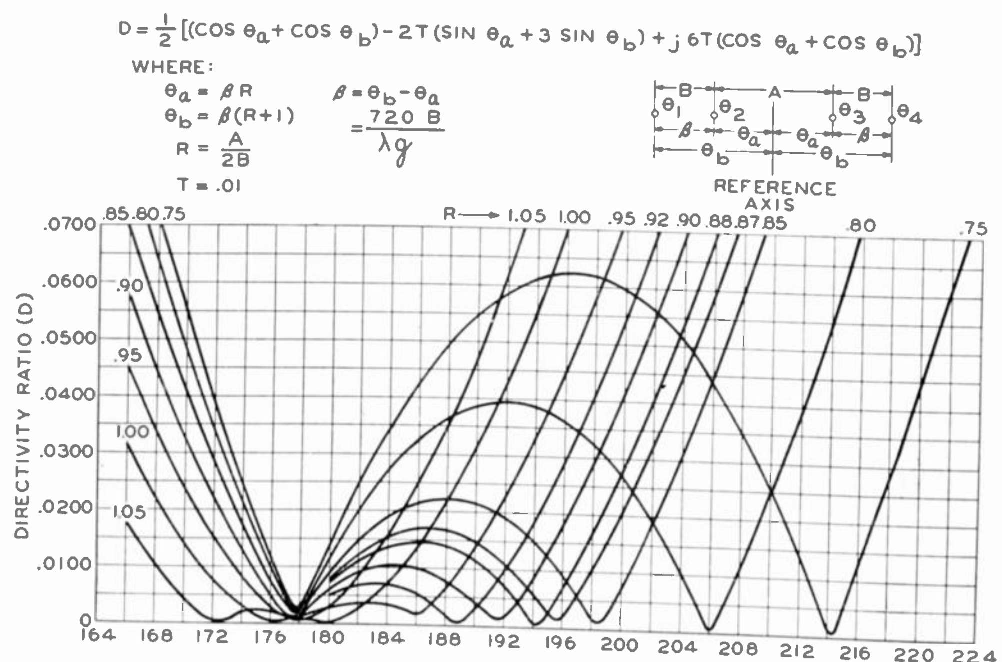

Fig. 5—D versus β characteristic of the four-hole coupler as a function of the displacement factor.

vhere θ_a and θ_b are the phase-shift angles corresponding :o the distance from the respective holes to the symnetrically located reference plane. With the relationships between A, B, R, and β given on Fig. 6, the

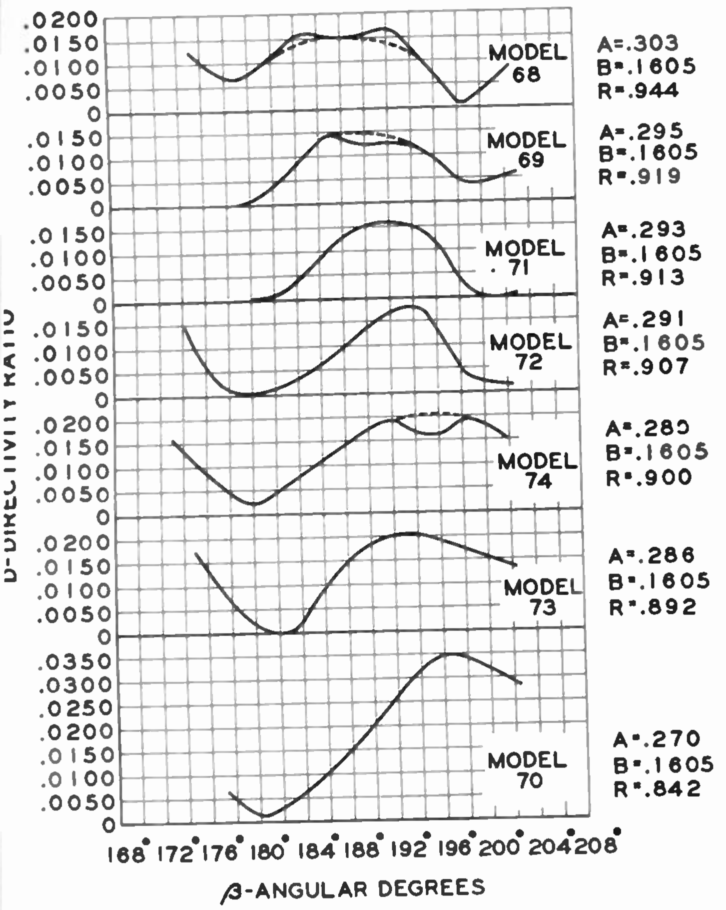

Fig. 6—Experimental D versus β characteristics of four-hole couplers as a function of the displacement factor.

directivity can be expressed in terms of three parameters, the orifice transmission factor T, the intrapair phase-shift angle β, and the displacement factor R.

If a hole size is now assumed, a family of curves can be plotted depicting directivity ratio D as a function of the intra-pair phase-shift angle β, with displacement factor R as the third parameter.

These curves contain a number of interesting facts which can be seen by referring to Fig. 5.

In the region of interest, $164° < \beta < 224°$, when $R = 1$ two points of minimum directivity ratio occur, one at 176°, the other at 180°, and the angular bandwidth for any given maximum directivity ratio is relatively narrow. As the value of the displacement factor is reduced from unity in successive steps, three significant facts should be noted.

The left minimum, after an initial shift, stays almost fixed at 178°, while the right minimum moves steadily to the right and, simultaneously, the intervening maximum, which occurs about midway between the two minima, increases in height. Any D versus β curve can be converted to a directivity versus frequency curve by assuming a value of intra-pair spacing. The resulting curve will have the same general shape as the original, and the position of the low-frequency minimum will be a direct function of the intra-pair spacing.

Fig. 6 shows the experimental results of tests on seven models of four-hole directional couplers in which the intra-pair spacing was held constant and the displacement varied.

By using the actual values of displacement A and intra-pair spacing B of the seven couplers, the displacement factors R can be calculated and listed in column I of Table I. Then the theoretical curve which corresponds most closely to each experimental curve is selected from Fig. 5 and the value of displacement factor listed in column II of Table I. This correspondence was based on the relative positions of the left minimum point and the maximum point. It is interesting to note that the right minima of the experimental curves consistently occurred at a higher angle than predicted by theoretical curves. In both four- and eight-hole couplers (discussed later), the experimental characteristics were found to be flatter and more extended (in other words, better) than expected from the theory. By dividing column I by column II, it can be seen that the quotients are very nearly constant at the value 1.05, which indicates that, on this basis, the theory falls short of agreement with experimental results by about 5 per cent.

TABLE I

$B=0.1605$	A	Column I R computed	Column II R estimated from Fig. 5	Column I ÷ column II
	0.303	0.944	0.900	1.049
	0.295	0.919	0.875	1.050
	0.293	0.913	0.870	1.050
	0.291	0.907	0.865	1.049
	0.289	0.900	0.860	1.048
	0.286	0.892	0.850	1.049
	0.270	0.842	0.810	1.040

Within the limits of experimental error, the main features of the theory are substantiated. Reading the curves from top to bottom, the left minimum remains fixed, the right minimum moves to the right, and the intermediate maximum increases in height. Therefore, it is apparent that, with a specified allowable maximum directivity ratio, a suitable displacement factor R can be selected from Fig. 5 to make the angular bandwidth a maximum. Since the left minimum of the experimental curves occurs at approximately 178° when an unmodified value of intra-pair spacing is used, it appears that the proximity effect which was so noticeable in two-hole couplers is now quite negligible. Presumably, the complicated proximity effect reactions occurring in a single pair tend to cancel out when two or more pairs are used.

Vector Representation of Four-Hole Coupler

An interesting explanation of the D versus β characteristic of a four-hole coupler is based on a vector method, and proceeds as follows: If $\theta_3=\theta_a$ and $\theta_4=\theta_b$, the directivity of a four-hole coupler is

$$4D = e^{-j\theta_b} + e^{-j\theta_a}(1 + j4T) + e^{j\theta_a}(1 + j8T) + e^{j\theta_b}(1 + j12T).$$

Letting $T=0.01$ and $R=0.8$ and multiplying through by $e^{j1.8\beta}$

$$4De^{j1.8\beta} = 1.0 + 1.0008e^{j(\beta+2°17')} + 1.032e^{j(2.6\beta+4°34')} + 1.0072e^{j(3.6\beta+6°50')}.$$

In the above equation, it can be seen that the directivity is expressed as the sum of four vectors, and the magnitude of the resultant is a measure of the directivity ratio.

Fig. 7 shows the vectors with all magnitudes and angles drawn to scale, and indicates what happens when β is increased across the band. Beginning with $\beta=166°$ in (A), it can be seen that the gap between vector 1 and vector 4 is fairly large and represents a large directivity ratio. As β increases to 172° in (B) the gap becomes smaller, and in (C) the condition representing the left minimum has been reached. Here the vectors from the first orifice pair are almost canceling each other and the vectors from the second orifice pair are also nearly canceling each other, so that the gap is approximately zero.

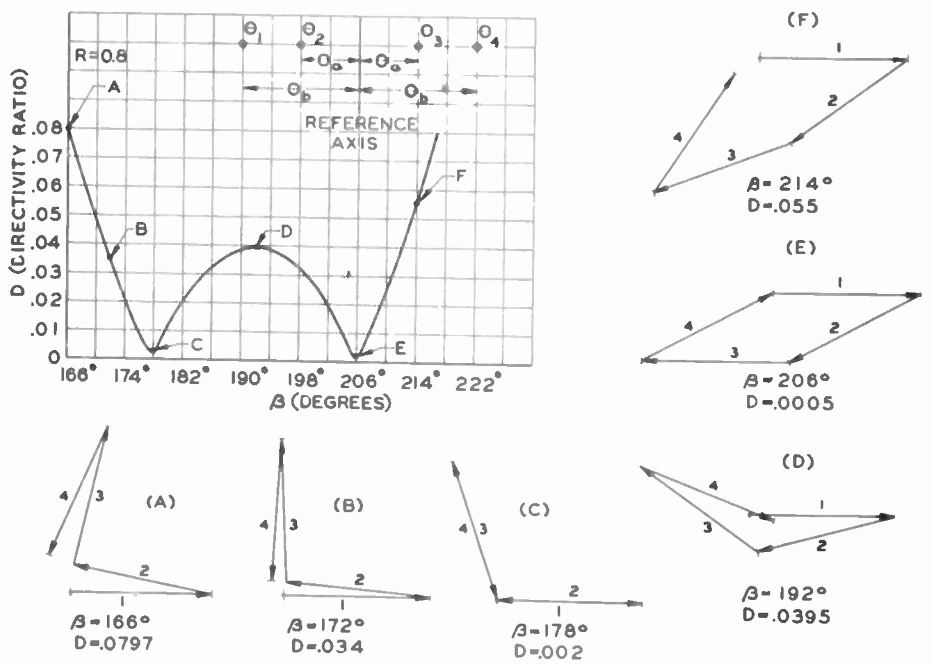

Fig. 7—Vector diagrams showing the D versus β characteristic of the four-hole coupler.

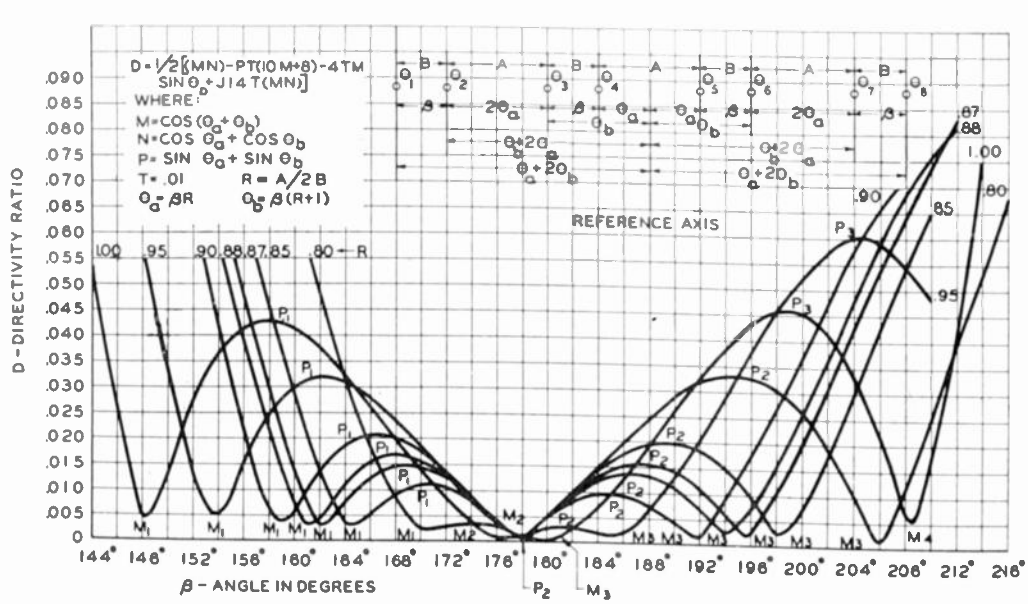

Fig. 8—D versus β characteristic of the eight-hole coupler as a function of the displacement factor.

The large gap appearing in (D) depicts the condition for the intermediate maximum which occurs between the two minima. (E) illustrates the vector condition for the right minimum. Here an interesting condition prevails. Although the vectors of each pair do not cancel each other, the resultant of the first orifice pair is almost exactly equal and opposite to the resultant of the second orifice pair, so that the over-all gap is practically zero. (F) shows the gap opening once again.

The Eight-Hole Directional Coupler

To provide a low directivity ratio over a still wider band of frequencies, the next simple configuration to be considered is the eight-hole coupler, consisting of four pairs of holes arranged in sequence. The directivity ratio is given in the Appendix. On Fig. 8 the directivity ratio is plotted against intra-pair phase-shift angle β with the displacement factor as the third parameter. The region $144° < \beta < 216°$ is of particular interest. For convenience, let the minimum points be numbered from left to right such as M_1, M_2, $\cdots$ and the intervening peaks numbered P_1, P_2, $\cdots$.

As the displacement factor is reduced from unity, several significant changes occur in the D versus β curves.

1. M_1 moves to the right.
2. M_2 moves to approximately 178° and remains fixed thereafter.
3. M_3 moves to the right.
4. M_4 moves to the right and out of the region of interest.
5. P_1 moves to the right and decreases in height.
6. P_2 moves to the right and increases in height.

The minimum M_2 is considered to be the basic minimum, and can be adjusted to fall at a desired frequency by proper selection of the intra-pair spacing.

Effect of Transmission Factor on D versus β Characteristics

Since one of the basic assumptions in Bethe's theory of orifice coupling is that the holes are extremely small, it is only natural to wonder how much accuracy is lost when the theory is applied to cases where hole diameters approach a quarter wavelength. Fig. 9 provides at least a partial answer to this question. Using a constant displacement factor of unity, D versus β characteristics are shown for several values of transmission factor T. It is evident that there is relatively little change as T is increased from a minimum of 0.002 through 0.005 to 0.01. Doubling the value of T to 0.02 begins to produce an appreciable effect, and a final step to 0.04 materially changes the curve. Thus it may be argued that, as T is reduced from 0.01 and approaches zero as a limit, the curve does not change greatly; but if larger transmission factors are involved, new curves should be derived.

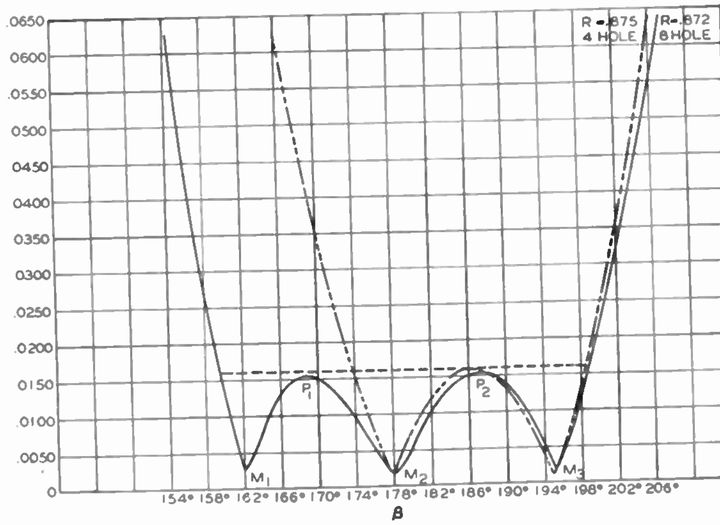

Fig. 10—Comparison of D versus β characteristic between the four- and eight-hole couplers.

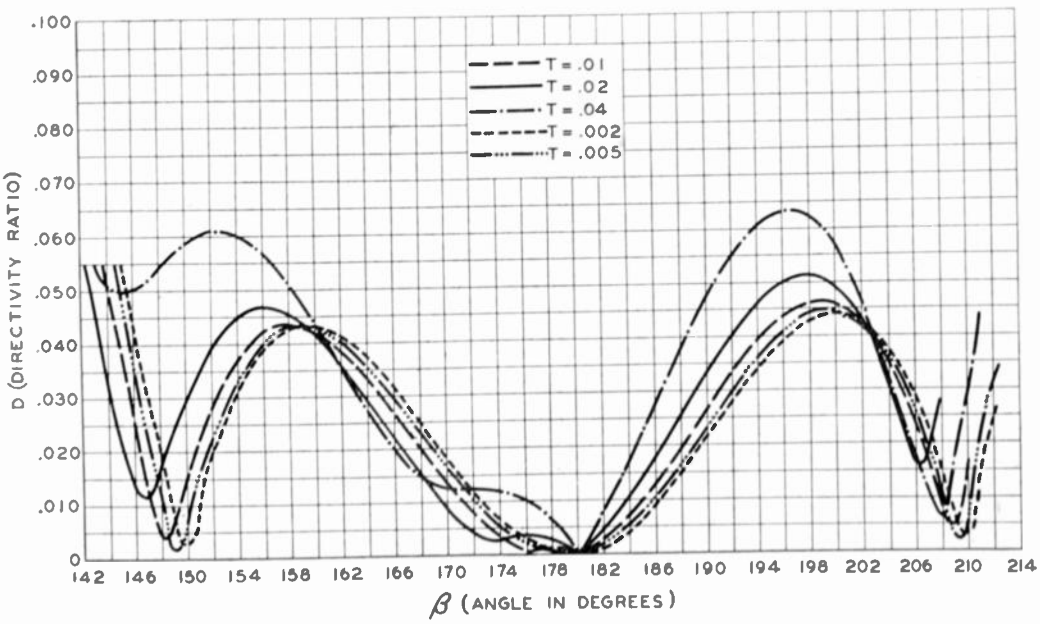

Fig. 9—D versus β characteristic of the eight-hole coupler as a function of the transmission factor ($R = 1.0$).

Comparison of Four- and Eight-Hole Couplers

Fig. 10 presents a simple comparison of the D versus β characteristic of a four-hole coupler with the D versus β characteristic of an eight-hole coupler. A displacement factor of 0.872 was selected in the latter case, because the two maxima, P_1 and P_2, are equal in height at this value. In the event that the maximum allowable directivity ratio was equal to the height of the maxima (0.016), the angular bandwidth would be about 40°. If a four-hole coupler was selected with the same tolerance, the displacement factor would be about 0.875, and the angular bandwidth is only 25°. Therefore, in this example, doubling the number of holes increased the bandwidth by 60 per cent. There does not appear to be any valid reason for the very close agreement between the displacement factors. Since the right portions of the two characteristics very nearly coincide, it can readily be seen that doubling the number of holes added another loop to the D versus β characteristic.

Conclusions

1. Mathematical expressions can be evolved for the directivity ratio of a directional coupler.
2. Based on these expressions, a fairly simple and plausible physical explanation has been developed and confirmed by measurement.
3. The equations have been reduced to curves which enable the practicing engineer to design a coupler with a minimum of effort.

Acknowledgment

The authors are indebted to many associates at the Bell Telephone Laboratories for contributions to the theoretical and experimental work, and in particular to J. P. Kinzer, who interpreted and extended the theory of orifice coupling.

Appendix

Derivation of Directivity Expressions

To evaluate the directivity of a coupler, it is necessary first to have quantitative knowledge of the field conditions existing at an orifice. This information is given by Bethe[3] in his theory of thin-wall orifice coupling. The next step requires the writing of a set of equations which sum the wave contributions of all orifices in the coupler. After carrying through the straightforward but lengthy algebra involved, the two following basic equations result:

$$2T_u \sum_{v=1}^{u-1} e^{-j\theta_v} p_v + je^{-j\theta_u} p_u + 2T_u e^{-j\theta_u} \sum_{v=u+1}^{n} p_v = -T_u \quad (1)$$

$$q_u = p_u e^{-j\theta_u} \qquad (u = 1, 2, \cdots, n) \quad (2)$$

where

θ_u = twice the angular phase shift of a wave traveling from the orifice center to a reference axis
u, v = indices of summation
p = backward wave
q = forward wave
T = transmission factor of orifice
n = number of orifices.

Two-Hole Coupler

From (1), the backward waves are

$$je^{-j\theta_1}p_1 + 2T_1 e^{-j\theta_1} p_2 = -T_1 \quad \text{and}$$
$$2T_2 e^{-j\theta_1} p_1 + je^{-j\theta_2} p_2 = -T_2.$$

Solving for p_1 and p_2,

$$p_1 = \frac{T_1\{je^{-j\theta_2} - 2T_2 e^{-j\theta_1}\}}{e^{-j\theta_1}\{4T_1T_2 e^{-j\theta_1} + e^{-j\theta_2}\}} \quad \text{and}$$

$$p_2 = \frac{T_2(j - 2T_1)}{4T_1T_2e^{-j\theta_1} + e^{-j\theta_2}}.$$

From (2), the forward waves are

$$q_1 = \frac{T_1\{je^{-j\theta_2} - 2T_2e^{-j\theta_1}\}}{4T_1T_2e^{-j\theta_1} + e^{-j\theta_2}} \quad \text{and}$$

$$q_2 = \frac{jT_2e^{-j\theta_2} - 2T_1T_2e^{-j\theta_2}}{4T_1T_2e^{-j\theta_1} + e^{-j\theta_2}}.$$

The directivity ratio

$$D = \frac{p_1 + p_2}{q_1 + q_2}$$

$$= \frac{jT_1e^{-j\theta_2} + e^{-j\theta_1}(jT_2 - 4T_1T_2)}{e^{-j\theta_1}[-2T_1T_2e^{-j\theta_1} + e^{-j\theta_2}(jT_1 + jT_2 - 2T_1T_2)]}.$$

When

$$T_1 = T_2 = T, \qquad 2D = \frac{e^{j\theta_1} + e^{j\theta_2}(1 + j4T)}{jT[e^{j(\theta_2-\theta_1)} + 1] + 1}.$$

Let the reference axis pass through the center of hole one and let $T \ll 1$. Then

$$2D = 1 + e^{-j\theta}(1 + j4T). \quad (3)$$

By a similar procedure, the directivity of an n-hole coupler is

$$D = \frac{1}{n}\sum_{u=1}^{n} e^{j\theta_u}[1 + j4T(u - 1)]. \quad (4)$$

Four-Hole Coupler

From (4),

$$4D = e^{j\theta_1} + e^{j\theta_2}(1 + j4T) + e^{j\theta_3}(1 + j8T)$$
$$+ e^{j\theta_4}(1 + j12T).$$

If a symmetrically located reference axis is chosen as

[3] H. A. Bethe, MIT Radiation Laboratory Reports 43-22 and 43-27.

shown on Fig. 5, then $\theta_1 = -\theta_4 = -\theta_b$ and $\theta_2 = -\theta_3 = -\theta_a$, and the above can be reduced to

$$2D = \cos\theta_a + \cos\theta_b - 2T(\sin\theta_a + 3\sin\theta_b) + j6T(\cos\theta_a + \cos\theta_b).$$

Eight-Hole Coupler

From (4),

$$8D = e^{j\theta_1} + e^{j\theta_2}(1 + j4T) + e^{j\theta_3}(1 + j8T) + e^{j\theta_4}(1 + j12T) + e^{j\theta_5}(1 + j16T) + e^{j\theta_6}(1 + j20T) + e^{j\theta_7}(1 + j24T) + e^{j\theta_8}(1 + j28T).$$

A symmetrically located reference axis as shown on Fig. 8 is chosen:

$$\theta_8 = -\theta_1 = \theta_a + 2\theta_b \qquad \text{Let} \quad \cos(\theta_a + \theta_b) = M$$
$$\theta_7 = -\theta_2 = 2\theta_a + \theta_b \qquad \cos\theta_a + \cos\theta_b = N$$
$$\theta_6 = -\theta_3 = \theta_b \qquad \sin\theta_a + \sin\theta_b = P.$$
$$\theta_5 = -\theta_4 = \theta_a$$

After numerous expansions and substitutions,

$$2D = (MN) - PT(10M + 8) - 4TM\sin\theta_b + j14T(MN).$$

Discussion on

"Influence of Reproducing System on Tonal-Range Preferences"*

HOWARD A. CHINN AND PHILIP EISENBERG

Herbert G. Messer[1]: I have read with considerable interest the paper by Chinn and Eisenberg. I want not only to compliment the authors on an excellent paper, but also to express appreciation of the publication of the data with reference to the conditions under which the tests were conducted.

However, I feel that I must take exception not only to some phases of the test itself, but also with the conclusions that were drawn. While the conclusions appear quite in harmony with the tests, I feel certain from my own experiments that tests can be made in a manner that would have been more fair to "high fidelity" and would have produced different conclusions.

My first contention is directed at the subjects which were selected as a jury. The authors stated that they "were selected by means of spot announcements" and that these subjects "represented a cross section of radio listeners." I judge this to be mistake number one, althought it may not be construed as such from a strictly commercial viewpoint. If all industry was directed solely at the average man, then we would never have opera, ballets, classical music, nor even Cadillac automobiles. The listeners, in my opinion, should have been very carefully selected from a group of people who know good music and are capable of appreciating it. This factor will be discussed further, later in this letter .

My second, and equally important, contention is that the program material was wholly inadequate to the cause of high fidelity. Someone has said that Charlie McCarthy sounds as good on a ten-dollar table model radio as he does on the highest-priced console. The number of popular disk records that can be benefited by high-fidelity reproduction is practically negligible. I have many records of popular music, and, with the single exception of two or three rumbas recorded by Don Apiazu and his orchestra, all are improved by narrow-band reproduction, if only because of the reduction in needle scratch. Little, if any, popular music is written with a wide tonal range. Even more serious is the fact that what tonal range it does possess is not engraved on the record at the studio. I am not averse to popular music—quite the contrary. But I feel that it has no place in a test of this nature.

I feel that so-called classical music is the only true method of demonstrating the value of high fidelity. Furthermore, this does not mean all classical music and most emphatically does not mean all recordings of even suitable classical music. The full tonal frequency range must be recorded, and the recording must be free of all forms of distortion. Lamentably, the record manufacturers have not acquitted themselves too brilliantly in this respect. It can be easily demonstrated that a poor recording sounds much better on a narrow-band reproducing system. It can likewise be adequately demonstrated that a brilliant recording sounds infinitely better on a wide-range reproducing system. I have demonstrated this fact to the complete satisfaction of a wide range of listeners, although I know that my reproducing system can be materially improved.

I would suggest that the authors repeat their tests, perhaps with identical reproducing equipment, but with the following changes:

(a) The subjects to be carefully selected only from people who know and like classical music, but who have formed no preconceived conclusion as to the merits of narrow- and wide-range reproduction. The subjects

* Proc. I.R.E., vol. 36, pp. 472–580; May, 1948.

[1] Byington and Cira., Rio de Janeiro, Brazil.

should preferably be those who are more accustomed to hearing music at the source (i.e., live talent), rather than hearing it almost wholly from records and radio. Under the latter conditions, they are more apt to favor that to which they are accustomed (no matter how horrible), rather than judge the case solely on its merits.

(b) A better selection should be made of program material. The authors listed only two classical selections. I am not familiar with the Preludia by Jarnefelt, but I am familiar with Bizet's "Carmen." In any recording that I have heard of Carmen, I see very little that will demonstrate the value of wide-range reproduction. I suggest the following program material, although I feel sure there are many other selections as good or better:

Classical:

"La Gaite Parisienne," by Offenbach—Victor Vinylite

"Concerto in B minor for Cello and Orchestra" by Dvorak—Victor

"Rossini's Overtures,"—Victor Vinylite

Many *ffrr* Decca English recordings

"Till Eulenspiegel," by Richard Strauss—Victor Vinylite

"The Whistler and His Dog"—Victor Red Seal (10-inch).

Popular:

"Green Eyes" and "El Maniñero," both by Don Apiazu and his Orchestra—Victor.

With reference to the selection of the "jury," I find an interesting relation between Table VI, page 579, and Table I, page 574. By averaging the results of all four experiments in Table VI, we find the following percentages:

Musical training: more than 2 years	23 per cent
Musical preference: classical	22 per cent

Now, compare these figures with Table I:

Tonal-range preference: wide	
Symphonic, uncompensated	25 per cent
Symphonic, compensated	27 per cent

In my opinion, this is a very significant agreement.

With further regard to the commercial desideratum of appealing to the average listener, we must first overcome the preconceived preference for narrow-band reproduction which we in the engineering profession imposed upon him. I would suggest some procedure whereby the selected listeners hear only *wide-band recordings* for at least one hour daily for at least a week. During that period they should not hear any other form of reproduction. Then, at the end of the week, make the comparative tests. My tests have proved that the average listener, without musical education, turns enthusiastically to wide-range reproduction. It is the only conclusion that makes any sense if symphonic composers know their job.

H. A. Chinn and P. Eisenberg[2]: Colonel Messer's comments relative to our studies on tonal-range preferences seem to have been occasioned by a misunderstanding of the objectives of our studies and the conditions surrounding them. We are sorry that Colonel Messer has been confused by our report.

Studies such as we have undertaken can be performed with an endless number of possible boundary conditions. We described the particular circumstances of our tests in considerable detail and presented the results in a like manner, in order that all interested parties could draw their own conclusions, if they so chose.

Colonel Messer's first objection concerns the makeup of the subjects whose preferences were studied. We are interested in the preferences of broadcast listeners, and consequently chose them as subjects. The title of our paper specifically states this fact. We look forward to the results of others who may undertake similar studies with more specialized groups such as suggested by Colonel Messer.

The second objection has to do with program content. Colonel Messer prefers the use of classical music to popular music and infers that only the latter was used in our work. Actually, more than 70 per cent of the subjects took part in experiments involving classical music. The program passages themselves were not standard arrangements but were carefully selected, after considerable research, and, of course, were chosen to present the widest possible tonal range.

Colonel Messer's third objection pertains to the quality of commercial recordings. His list of records recommended for future tests suggests that he is of the opinion that our test recordings were of quality comparable (or perhaps inferior) to his selections. Such is far from the case! The recordings used for portions of the test were not commercial pressings or even excerpts from standard arrangements. As detailed, they were special "master" recordings made particularly for the tests. Furthermore, a given recording was used for only one session and then discarded. This was done in order to avoid any possible degradation because of record wear.

The comparison Colonel Messer makes between Tables I and VI, while interesting, has no significance. Analysis of the data shows no relation between those preferring wide range and those having musical training or a preference for classical music.

In his efforts to reconcile the results of our work with his own convictions, Colonel Messer has overlooked the most important point of all; namely, that the conclusions relate to "single-channel listening—using present-day broadcasting pickup techniques." Furthermore, although we report the preferences of the majority, I am unaware that we have ever advocated narrow-band reproduction. In fact, quite the reverse is specifically recommended.

[2] Columbia Broadcasting System, Inc., New York, N. Y.

Discussion on

"The Application of Matrices to Vacuum-Tube Circuits"*

J. S. BROWN AND F. D. BENNETT

W. Buchholz[1]: In their paper, Brown and Bennett draw attention to the usefulness of matrix notation in the analysis of linear, unilateral, vacuum-tube circuits. Some of the results obtained (e.g., equation (25) of their paper) bear a close resemblance to Bode's generalized method of network analysis (footnote reference 20 of the paper); indeed, the determinants of the z and y matrices of this paper are Bode's "immittance" determinants Δ applied to a four-terminal network. Matrix methods, however, have the further advantage of permitting, by simple addition and multiplication of matrices, the analytical interconnection of elementary networks into more complex structures. The two methods of approach could easily be combined, a step one might recommend for future papers along this line.

In giving two examples to illustrate matrix techniques, the authors seem to imply that the direct application of Kirchhoff's equations is considerably more involved. They state that the second example, the derivation of the input impedance of the grounded-cathode triode amplifier with interelectrode capacitances (Fig. 9 of the paper), "would have required the solution of seven loop equations." The authors may not have intended this to mean that there are seven simultaneous equations to be solved; for there are actually only three, or at most four, independent loops. But their remarks give an erroneous impression; when properly set up, the direct approach is quite simple, and it is seen, as the authors finally mention at the end of their paper, that little if any time is saved in using matrix methods.

It is worth illustrating this by comparing the example cited above with the node method of analysis,[2,3] which is gradually achieving recognition in the literature. Only *two* equations are needed here.

Let E_1 = grid voltage, E_2 = plate voltage, I_1 = input current,

$$Y_1 = j\omega C_{gk}, \quad Y_2 = 1/r_p + 1/Z_L + j\omega C_{pk}, \quad Y_3 = j\omega C_g.$$

Replace the tube by a current generator,[3] $-g_mE_1$, in parallel with the plate resistance r_p. Then

$$-g_mE_1 = (Y_2 + Y_3)E_2 - Y_3E_1, \qquad (1)$$

$$I_1 = (Y_1 + Y_3)E_1 - Y_3E_2. \qquad (2)$$

(Note that $Y_1 + Y_3 = y_{11}$, $-Y_3 = y_{12}$, $Y_2 + Y_3 = y_{22}$ in the authors' matrix notation.)

Solve (2) for E_2 and substitute into (1). Collecting terms gives, omitting two short steps,

$$E_1[g_m - Y_3 + (Y_2 + Y_3)(1 + Y_1/Y_3)] = (1 + Y_2/Y_3)I_1. \qquad (3)$$

Substituting again for Y_1 and Y_3 gives the input *admittance*

$$Y_i = \frac{I_1}{E_1} = j\omega(C_{gk} + C_{gp}) + \frac{j\omega C_{gp}g_m + \omega^2 C_{gp}^2}{Y_2 + j\omega C_{gp}}. \qquad (4)$$

This expression is much more useful for calculations than the reciprocal expressions for z_{11} as given by the authors (on p. 851 of the paper), to which it is easily converted. For Y_2 is merely the parallel combination of Z_L, r_p, and C_{pk}, and frequently $Y_2 \approx 1/Z_L$. Also, if $|Y_2| \gg \omega C_{gp}$ and the small conductance $\omega^2 C_{gp}^2$ is neglected, there results immediately the well-known expression

$$Y_i \approx j\omega[C_{gk} + (1 + K)C_{gp}]$$

where $K = g_m/Y_2$, the tube gain. The term $(1+K)C_{gp}$ gives the so-called Miller effect.

The above analysis is considerably shorter. Thus matrix methods do not save labor and present few advantages in relatively simple cases, as has been pointed out before by several authors, especially if one starts "from scratch." But when one proceeds to combine individual stages, filters, and other bilateral networks into complex networks, the use of matrices can be of great value in organizing the work. The solution of equations is reduced to routine matrix manipulations which, while they are not shorter, may easily reduce the number of mistakes made in handling many complex equations. In numerical cases, matrix methods are amenable to machine treatment on modern high-speed computers. Incidentally, tensor methods (footnote reference 8 of the paper) are a further step in the same direction.

Much work is still to be done in compiling the matrices of component networks, such as the authors have done in the case of the basic triode amplifiers. If suitable tables were available one could indeed save time, for one would no longer start the analysis "from scratch."

I might also draw attention to an obvious misprint in the line below equation (1b), which should read $\eta = AD - BC$, as in the rest of the discussion.

* Proc. I. R. E., vol. 36, pp. 844–852; July, 1948.

[1] California Institute of Technology, Pasadena, Calif.

[2] Electrical Engineering Staff of the Massachusetts Institute of Technology, "Electric Circuits," John Wiley and Sons, New York, N. Y., 1943; chap. VI, secs. 14 and 15.

[3] J. H. Mulligan, Jr., and L. Mautner, "The steady-state and transient analysis of a feedback video amplifier," Proc. I.R.E., vol. 36, pp. 595–609; May, 1948. See particularly, Fig. 2.

F. D. Bennett[4] **and J. S. Brown**[5]: We would like to thank Mr. Buchholz for his comments on our paper. In particular, the statement on page 851 to which he refers, "but would have required the solution of seven loop equations," should be corrected by substituting "branch" for "loop." R. S. Glasgow (footnote reference 22 of the paper) actually makes use of seven branch equations derived by application of Kirchhoff's laws in his rather laborious derivation of the input impedance of the triode.

The node method of analysis seems to give access to the input admittance of the triode circuit with a minimum of effort and with a convenience not offered by either Kirchhoff's laws in their classical form or the matrix method "starting from scratch." If we admit the necessary basic structure of the matrix method, the power and elegance of which it is capable also show to better advantage in this example when we solve for the input admittance as Mr. Buchholz has done.

Rewriting for reference,

$$Y_{TC}\| = \begin{Vmatrix} j\omega(C_{gk}+C_{gp}) & -j\omega C_{gp} \\ (g_m - j\omega C_{gp}) & j\omega(C_{pk}+C_{gp}) + \frac{1}{r_p} + \frac{1}{Z_L} \end{Vmatrix}$$

[4] Ballistics Research Laboratory, Aberdeen Proving Ground, Aberdeen, Md.

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and adding the definition

$$Y_t = \frac{1}{Z_{11}},$$

we find, using the identities of Appendix I,

$$Y_i = \frac{|Y|}{Y_{22}} = Y_{11} - \frac{Y_{12}Y_{21}}{Y_{22}}.$$

Evaluating this expression from the admittance matrix, we are led to

$$Y_i = j\omega(C_{gk}+C_{gp}) + \frac{j\omega C_{gp}g_m + \omega^2 C_{gp}^2}{j\omega(C_{pk}+C_{gp}) + \frac{1}{r_p} + \frac{1}{Z_L}},$$

which is Mr. Buchholz's equation (4).

We would submit that, once one is familiar with the matrix technique, a definite decision as to which attack will lead to less work becomes to a great extent an individual matter depending on a preference between the fundamental approach or the use of a mechanical method where it applies.

Correspondence

Radar Reflections in the Lower Atmosphere*

Several letters and papers have appeared in these PROCEEDINGS on the subject of radar reflections in the lower atmosphere,[1-4] a phenomenon commonly known as "angel" reflections. These are sharp echoes of short duration which are observed most frequently below heights of about 3,000 feet. Some recent observations indicate rather convincingly that perhaps the chief, and possibly the only, sources of these reflections are flying insects and birds. We were led to this viewpoint when all attempts to synthesize "angels" by artificially producing boundaries of temperature, humidity, or turbulence failed completely, and when visual observation of insects coincided strikingly with the radar observations.

These tests were made in January, 1949, at the Gila Bend, Arizona, site of the Naval Electronics Laboratory of San Diego during some co-operative experiments of the Naval Electronics Laboratory and Bell Telephone Laboratories. The radar equipment was the Bell Laboratory vertical-incidence radar shown in the photograph, Fig. 1. Separate antennas consisting of conical horns fitted with molded polyethylene lenses, 30 inches in diameter, are used for transmitting and receiving; this eliminates the recovery time of a receiver disconnect switch, and, since the direct pickup from transmitting antenna to receiving antenna is very low, it is possible to observe reflections occurring at heights as low as the antennas themselves. Simultaneous observations at two wavelengths are made by transmitting interlaced pulses on wavelengths of 3.2 and 1.25 cm. The pulse widths are about 0.15 microsecond, the recurrence rate is 5,000 pulses per second, and the transmitted peak power is about 40 kw at 3.2 cm and 15 kw at 1.25 cm. The sensitivity is sufficient to permit detection of a low-velocity 22-calibre rifle bullet at its maximum vertical range of about 3,500 feet; the reflection from a high-velocity 22-calibre bullet can be detected moving in the noise at over 5,000 feet altitude. Fig. 2 is a photograph of the receiver cathode-ray tube showing the reflection from a 22-calibre rifle bullet at a range of about 1,400 feet. The photograph also shows an "angel" reflection at about 500 feet.

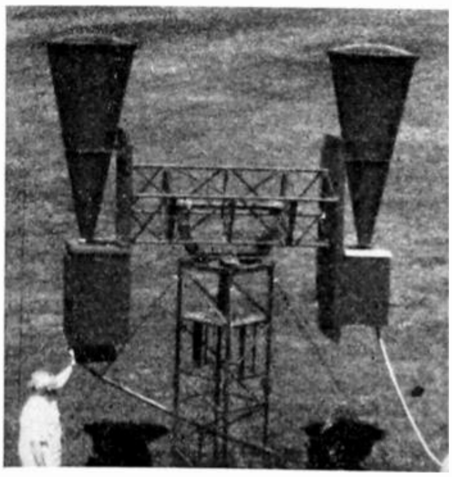

Fig. 1—Photograph of antennas used in the vertical-incidence radar.

In an attempt to synthesize "angel" reflections, a small charge of nitro-starch was exploded some 500 feet above the antennas and slightly off the beam; an airplane was flown over the radar at low levels and reflections from the exhaust gases were looked for after the plane had passed overhead; and bonfires were built up-wind, so that the hot combustion gases and steam clouds formed by pouring water on heated rocks were

* Received by the Institute, February 28, 1949.

[1] H. T. Friis, "Radar reflections from the lower atmosphere," PROC. I.R.E., vol. 35, p. 494; May, 1947.

[2] W. B. Gould, "Radar reflections from the lower atmosphere," PROC. I.R.E., vol. 35, p. 1105; October, 1947.

[3] M. W. Baldwin, "Radar reflections from the lower atmosphere," PROC. I.R.E., vol. 36, p. 363; March, 1948.

[4] W. E. Gordon, "A theory on radar reflections from the lower atmosphere," PROC. I.R.E., vol. 37, pp. 41–43; January, 1949.

carried into the beam of the antennas. In no case were any reflection effects observed which could be attributed to dielectric-constant boundaries. However, we had no independent means for determining whether or not sharp dielectric constant boundaries were actually produced.

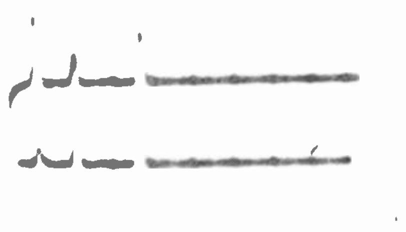

Fig. 2—Photograph showing an "angel" reflection at about 500 feet and the reflection from a 22-calibre rifle bullet at about 1,400 feet. The top trace is for 3.2-cm wavelength, the bottom trace for 1.25-cm wavelength. The pulses at the extreme left are the transmitted pulses. The dark spots on the base lines are 1-microsecond markers.

Visual observation of flying insects, even at low heights, is almost impossible by day, but even small insects become visible as bright glints in the beam of a searchlight at night. For a first trial, to get an order of magnitude of "bug density," a searchlight beam was directed vertically near the 200-foot tower used in NEL's propagation studies. Observers located at different levels on the tower counted flying insects, while the operator of the radar, located some half-mile distant, counted "angel" reflections occurring in the first 250 feet of height. During the half-hour observation period, 206 insects were counted; in the same time interval, 55 angel reflections were observed.

Although some insects may have been attracted by the searchlight, the observations showed them to be present in sufficient number to account for the "angel" reflections. The searchlight was then moved to the radar location and observers located on four sides of the beam called out when an insect was in the light beam. During a quarter-hour period, twenty "angel" reflections were observed, of which fifteen coincided with the sighting of an insect. There were thirteen insects sighted which did not correspond with a radar reflection, but this is readily explainable since the diffuse beam of the searchlight was much wider than the radar beam. An observer using a theodolite focused for a height of 400 feet observed an insect coincident with the appearance of an "angel" reflection at the same range.

The insect explanation is in keeping with most of the observed characteristics of "angel" reflections, such as their apparent small size; their movements at speeds comparable to wind velocity, usually with but sometimes against the wind; their presence in both the daytime and nighttime; and their appearance in much greater number in warm weather than in cold weather.

In some previous work, consisting of part-time, on-the-spot observations, we did not find a correlation between "angel" activity and microwave propagation effects; if the insect explanation is the correct and only explanation, such a correlation would not be expected. The vertical-incidence radar, however, may be a useful tool to the entomologist for observing how the density of flying insects varies with the season, time of day, weather conditions, and the like.

Acknowledgment

The assistance and suggestions of J. B. Smyth, L. J. Anderson, and other members of the Naval Electronics Laboratory group are gratefully acknowledged. L. R. Lowry and S. E. Reed of the Bell Telephone Laboratories also co-operated in the work.

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Pulse Radar History*

It is the purpose of this note to supply information on the early pulse radar work of the U. S. Navy. This is done to correct the literature,[1,2] wherein the early pulse radar development in this country is omitted, whereas the early pulse radar work done in England is described at some length. Fink[1] leaves the impression that early work on radar at the U. S. Naval Research Laboratory ended when A. H. Taylor detected aircraft by cw signals at 60 Mc, using the wave interference pattern as an indication. Ridenour[2] indicates that the earliest full success of pulse radar at the Naval Research Laboratory was in early 1939, when a radar set was given exhaustive tests during battle maneuvers on the U.S.S. *New York*. He refers to a book by H. E. Guerlac, covering the prior history. A check with Little, Brown and Co., Boston, Mass., has revealed that this book will not be published. It is evident that these errors by omission are the result of unavailability to the authors of Navy records on the subject.

It will be of interest to members of The Institute of Radio Engineers that, in March, 1934, R. M. Page recorded: "Work was resumed with Mr. (L. C.) Young on the airplane echo problem. It was decided to attack this problem in a manner similar to that by which supersonic depth finding is accomplished." This log-book entry is the written record that marks the beginning of active work on pulse radar. By December, 1934, a 60-Mc pulse system was operated and aircraft were detected with it. It is interesting to note that Fink and Ridenour set the date for similar activity by Sir Robert Watson-Watt of England as early in 1935. In April and May, 1936, according to log-book entries, Page and Robert C. Guthrie of the U. S. Naval Research Laboratory had built and demonstrated a 28-Mc, 6-μs pulse radar system giving ranges on airplanes to 24 nautical miles. By August, 1936, Page and A. A. Varela had demonstrated a successful 200-Mc radar system using a common antenna for transmitting and receiving. During 1936, demonstrations were made to many persons, including Admirals Bowen, Stark, and Standley.

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* Received by the Institute, September 27, 1948.

[1] D. G. Fink, "Radar Engineering," McGraw-Hill Book Co., New York, N. Y., 1947; p. 7.

[2] L. N. Ridenour, "Radar System Engineering," McGraw-Hill Book Co., New York, N. Y., 1947; p. 14.

Some Notes on Corona Static*

Listeners to radio broadcasting in urban areas ordinarily are not familiar with the natural static caused by corona discharge during an electrical storm. The crackling type of static, reminiscent of the noise produced in a receiver by a sparking Tesla coil, is easily recognized by a rural radio listener, however. This is especially true if his receiving location is equipped with one or more lightning rods.

The writer, being a city dweller, became quite intrigued by this phenomenon while working for the radiophoto receiving station of the Office of War Information, which was located at Slingerlands, N. Y., just outside of Albany.

During the course of a number of violent electrical storms sweeping down from the mountains during the summer of 1945, several peculiarities of the static were noted. For example, the corona static would start building up gradually until it finally drowned out the BBC station GWO, in London, which was being monitored on 9.625 Mc. The intensity of the static would keep mounting until it reached a steady roar. Then, at the stroke of lightning, it would cease abruptly, and after a while begin building up and repeat the operation all over again. This cycle might continue until the fury of the storm had passed.

Sometimes corona static would begin to develop in the receivers and continue at a moderate rate and never be interrupted, despite the fact that lightning had flashed and discharged. Generally, this discharge occurred at some distance from the receiving station, and probably for this reason did not affect the static locally. At other times the corona static would grow in intensity and then die away at a gradual rate, as though the charged cloud had approached, passed over the receiving station, and continued on out of range.

Still another type was very dramatic. No corona would be noticeable in the communications receivers during some portion of a particular storm. But, with the stroke of lightning, the corona static would burst with full force through the loudspeakers. It would continue at this high level and gradually wear itself off. Lightning would strike again, and once more the corona static would burst forth with maximum intensity. Then it would wear off again, as though the cloud had passed beyond the station, or the strain had reached an equilibrium.

As interesting as all these manifestations were, no study could be made due to the nature of operations at the receiving station.

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* Received by the Institute, November 5, 1948.

High-Impedance Cable*

I have read with some interest the paper by Kallmann[1] and with even more interest the discussion between Kallmann and Winkler.[2]

The disagreement between these gentlemen centered around the question of a correction factor to be applied to the calculation of the inductance of a long solenoid surrounded by a cylindrical shield. In view of the meager theoretical information available, it was difficult to arrive at a satisfactory conclusion. In fact, a partial conclusion was finally reached only by appeal to experiment.

There is another approach which does not seem to have been covered in the literature, which, although it has its weaknesses, is capable of yielding an indication as to the true behavior of the inductance as shield diameter and coil diameter approach each other. In fact, this analysis goes further and attempts to predict the behavior of the delay line when the electrical length of each coil turn may not be neglected.

This approach assumes that the coil is wound with very small wire and that the gap between coil and shield is very small compared to the coil diameter. One then imagines the coil and shield to be cut physically on a line parallel to the cable axis and the whole assembly unwrapped, so that the coil shield becomes a plane surface of width equal to its circumference and the coil turns become straight, parallel wires of equal height above this plane surface. This arrangement now constitutes a multiwire line of length π_a (in Kallmann's notation), and can be analyzed in terms of the theory of the *TEM* mode for such lines. (See Fig. 1.) For $\rho \ll D, g$ (Fig. 1), one obtains for the characteristic impedance

$$Z_0 = \frac{60}{\sqrt{k}} \ln \frac{2g}{\rho} + \frac{120}{\sqrt{k}} \sum_{\mu=1}^{\infty} \left\{ \ln \left[1 + \left(\frac{2g}{\mu D} \right)^2 \right]^{1/2} \right\} \cos u\theta_c. \quad (1)$$

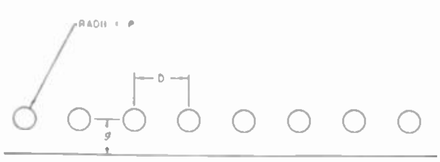

Fig. 1—System of parallel wires above a plane.

In case the electrical length per turn can be neglected, then $\cos u\theta_c = 1$ for terms in the series which are not negligible, and (1) becomes

$$Z_{00} = \frac{60}{\sqrt{k}} \ln \frac{2g}{\rho} + \frac{60}{\sqrt{k}} \sum_{\mu=1}^{\infty} \ln \left[1 + \left(\frac{2g}{\mu D} \right)^2 \right]. \quad (2)$$

The infinite series in (2) may be summed, and (2) becomes

$$Z_{00} = \frac{60}{\sqrt{k}} \left\{ \ln \left[\frac{D}{2\pi\rho} \sinh \frac{2\pi g}{D} \right] \right\}. \quad (3)$$

* Received by the Institute, October 1, 1948.

[1] Heinz E. Kallmann, "High-impedance cable," Proc. I.R.E., vol. 34, pp. 348–351; June, 1946.

[2] M. R. Winkler and Heinz E. Kallmann, discussion on "High-impedance cable," Proc. I.R.E., vol. 35, p. 1097; October, 1947.

From (3), several other quantities are immediately obtainable. Thus, the inductance per meter of wire is

$$L' = \frac{\sqrt{k}\, Z_{00}}{3(10)^8} \text{ henry/meter,}$$

so that the inductance per meter of delay line is

$$K = \frac{\pi a n \sqrt{k}\, Z_{00}}{3(10)^8} (10)^{-2}$$

$$= 2(10)^{-9} \pi a n \left\{ \ln \frac{\sinh 2\pi g n}{2\pi\rho n} \right\}$$

$$\text{henry/meter,} \quad (4)$$

since

$$D = \frac{1}{n}.$$

Similarly, the capacitance per meter of wire is

$$C' = \frac{\sqrt{k}}{3(10)^8 Z_{00}},$$

so that the capacitance per meter of delay line is

$$C = \frac{\pi a n \sqrt{k}}{3(10)^8 Z_{00}} (10)^{-2}$$

$$= \frac{\pi a n k}{18} (10)^{-9} \left\{ \ln \frac{\sinh 2\pi g n}{2\pi\rho n} \right\}^{-1}$$

$$\text{farad/meter.} \quad (5)$$

The velocity of propagation is

$$V = V' \frac{D}{\pi a} = \frac{3(10)^8}{\pi a n \sqrt{k}} \text{ meters/second,} \quad (6)$$

and the time delay is

$$T = \frac{1}{V} = \frac{\pi a n \sqrt{k}}{3(10)^8} \text{ seconds/meter.} \quad (7)$$

Referring now to (4), this is rewritten as

$$L = 2(10)^{-9} \pi a n \{ \ln \sinh (2\pi g n) - \ln 2\pi\rho n \}. \quad (8)$$

We shall now make an assumption roughly equivalent to the assumption of a uniform current sheet in the normal calculation of inductance. We shall assume that D approaches 2ρ in magnitude. Admittedly, this is contrary to an assumption previously made in computing the impedance coefficients and, strictly, is not permissible. However, the trend in the value of ($\ln 2\pi\rho n$) can certainly be observed, and it is a fair guess to say that this term can be made very small. At the same time, we impose the requirement that $g \gg D$, so that

$$\sinh (2\pi g n) \to \tfrac{1}{2} e^{2\pi g n}$$

and

$$\ln \sinh (2\pi g n) \to 2\pi g n - \ln 2 \approx 2\pi g n.$$

In that case, (8) becomes

$$L \approx 4\pi^2 (10)^{-11} a g n^2 \approx 4\pi^2 (10)^{-11} d g n^2 \quad (9)$$

where g is now measured in centimeters rather than meters.

Let us return now to the discussion between Kallmann and Winkler. The formula for inductance is taken as

$$L = 10^{-11} \pi^2 n^2 d^2 M \text{ henrys/meter} \quad (10)$$

where, according to Kallmann,

$$M_k = 1 - \left(\frac{d}{a} \right)^4 \quad (11)$$

and, according to Winkler,

$$M_w = 1 - \left(\frac{d}{a} \right)^2. \quad (12)$$

Our quantity g is given approximately by

$$a = d + 2g$$

or

$$\frac{d}{a} = \frac{1}{1 + 2\dfrac{g}{d}}.$$

Since we have assumed that $g/d \to 0$, we get

$$M_k \approx 8 \frac{g}{d}$$

$$M_w \approx 4 \frac{g}{d}.$$

Using M_k in (10),

$$L \approx 8(10)^{-11} \pi^2 d g n^2. \quad (13)$$

Using M_w in (10),

$$L \approx 4(10)^{-11} \pi^2 d g n^2. \quad (14)$$

Apparently the use of (12), rather than (11), gives agreement with (9).

Incidentally, what of the capacitance formulas? Under the assumptions made, (5) should certainly reduce to the result obtained for a coaxial line under the same conditions.

Under the same assumptions as made for L, (5) reduces to

$$C = \frac{ak}{36g} (10)^{-7} \text{ farad/meter} \quad (15)$$

where, again, g has been converted to centimeters. On the other hand, for the coaxial line,

$$C = \frac{k \times 10^{-9}}{18 \ln \dfrac{a}{d}}. \quad (16)$$

But, since

$$\frac{a}{d} = 1 + 2\frac{g}{d},$$

$$\ln \frac{a}{d} \approx 2\frac{g}{d}, \qquad \frac{g}{d} \ll 1,$$

and if g is converted to centimeters,

$$C \approx \frac{ak}{36g} (10)^{-7} \text{ farad/meter,}$$

in agreement with (15).

In conclusion, one can indicate that the velocity of propagation is approximated by a simple rule of thumb: It is simply the velocity of propagation of a plane wave in a medium of dielectric constant k, modified by a factor which is the ratio of the coil pitch to the coil circumference. This approximation has been observed a long time ago in the behavior of the self-resonant, single-layer-solenoid type of rf chokes, and more recently in the behavior of traveling-wave tubes. The time delay T is, of course, merely the reciprocal of this velocity.

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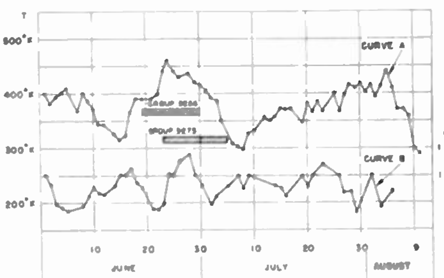
Circularly Polarized Solar Radiation on 10.7 Centimeters*

The solar radio emission on a wavelength of 10.7 cm shows a small circularly polarized component which changes slowly from day to day. One definite change from left-handed to right-handed polarization can be associated with the appearance of a sunspot on June 23, 1948. The observations have been conducted with the aid of a phase-shifting plate placed in front of the 4-foot parabolic reflector which is regularly used to observe the sun at Ottawa.[1] This plate is constructed from parallel metallic sheets, and is so proportioned that two incident waves, perpendicularly polarized, have a phase difference of 90° upon emerging from the plate. This property enables a circularly polarized wave to be changed into a linearly polarized wave. The orientation of the dipole with respect to the axis of the plate and the strength of the signal determine the sense of rotation of the rotating electric vector; a right-handed polarization is defined as clockwise rotation in space when looking along the direction of propagation.

Before observing the solar radiation with the plate, the dipole is rotated through 360° in order to find any linearly polarized component. No variation greater than the experimental error of about 2 per cent was detected. This observation indicates that the solar radiation consists of either randomly or circularly polarized radiation, or a mixture of both polarizations. The solar radiation as regularly measured with the dipole is plotted in Fig. 1(a). With the phase-shifting plate in front of the reflector, two measurements of



1948

Fig. 1(a)—Daily variations of solar noise as measured with a dipole, expressed in temperature of antenna radiation resistance. (b)—Ratio of right-handed to left-handed solar radiation.

the solar radiation were made; one with the dipole placed to receive right-handed radiation, and the other placed to receive left-handed radiation. The ratio of the right-handed to the left-handed radiation from June 1 to August 8 is plotted in Fig. 1(b). Near June 15 and July 8, the sun was relatively free from large spots and a polarization ratio of unity was observed. This indicates that the radiation from the quiet sun is randomly polarized. From June 23 to July 2, the radiation changed from predominately left-handed to predominately right-handed, and back again. During this period, two large sunspot groups (Mount Wilson Nos. 9266 and 9275) were simultaneously visible on the sun's disk. The definite change in the polarization ratio and in the radiation as measured with the dipole occurred with the appearance of the large sunspot group No. 9275. These observations indicate that sunspots can produce a circularly polarized 10.7-cm radiation. Besides, for the period under consideration, the average value of the polarization ratio is about 0.975, indicating an excess of left-handed radiation produced by the spots present on the sun. The observations used in calculating a ratio were taken several times during a half-hour period. This procedure was performed at least once a day. Because of experimental difficulties, the error in a single point is about ±0.015.

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* Received by the Institute, December 2, 1948.

[1] A. E. Covington, "Solar noise observations on 10.7 cms," Proc. I.R.E., vol. 36, pp. 454–457; April, 1948.

Network Representation of Input and Output Admittances of Amplifiers*

The analysis of input and output admittances of linear amplifiers with negative grid bias is usually obtained considering separately the real and imaginary components and introducing such simplifying or limiting assumptions as the case may suggest. It may be shown, however, that it is always possible to represent these admittances with linear passive networks, which are derived quite simply from the equivalent circuit of the amplifier with certain additions of parallel or series branches. The added elements can be expressed as combinations of resistances, capacitances, and inductances which are functions of μ.

This representation may be of didactic interest because it provides an immediate and complete physical picture of the circuit performance.

With reference to Table I, let us consider, for instance, the grounded-cathode amplifier. The input and output admittances are[1]

$$Y_{in} = Y_1 + \frac{Y_2(Y_3 + Y_L)}{Y_2 + Y_3 + Y_L} + \frac{\mu Y_3 Y_2}{Y_2 + Y_3 + Y_L} \quad (1)$$

$$Y_{out} = Y_3 + \frac{Y_2(Y_1 + Y_S)}{Y_1 + Y_2 + Y_S} + \frac{\mu Y_3 Y_2}{Y_1 + Y_2 + Y_S} \quad (2)$$

where Y_1, Y_2, Y_3, Y_S, Y_L are specified in the amplifier circuit.

The expressions (1), (2) may be represented by networks (see Table I) derived from the original passive quadripole with the addition of a parallel branch at the (input, output) terminals. The admittances of these branches are, respectively,

$$Y' = \frac{\mu Y_3 Y_2}{Y_2 + Y_3 + Y_L}, \quad Y'' = \frac{\mu Y_3 Y_2}{Y_1 + Y_2 + Y_S}$$

and may be represented as simple combinations of elements.

For instance, when the load is resistive, we have $Y_1 = 1/R_g + j\omega C_{gk}$, $Y_2 = j\omega C_{gp}$, $Y_3 = 1/R_p$, $Y_L = 1/R_L$ and, consequently,

$$\frac{1}{Y'} = \frac{R_p}{\mu} + \frac{1}{j\omega\mu C_{gp}}\left(1 + \frac{R_p}{R_L}\right).$$

In this case, the network representation of Y_{in} assumes the form of Fig. 1(a) From it one sees how the input capacitance approaches a maximum value $C_{gk} + (\mu+1) \cdot C_{gp}$ when the load impedance is very high.[2]

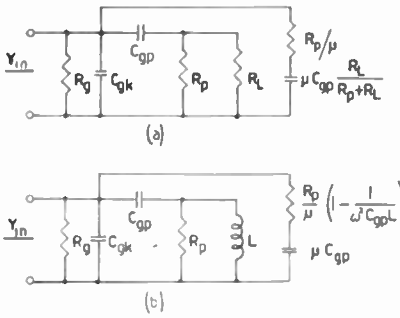


Fig. 1—Network representation of the input admittance of a linear grounded-cathode amplifier; (a) with resistive load, (b) with inductive load.

When the load is inductive, $Y_L = 1/j\omega L$, we have

$$\frac{1}{Y'} = \frac{R_p}{\mu}\left(1 - \frac{1}{\omega^2 C_{gp} L}\right) + \frac{1}{j\omega\mu C_{gp}}.$$

In this case, the network representation (Fig. 1(b)) shows that a negative input resistance is obtained, provided $w^2 C_{gp} L < 1$.

An interesting application of (2) is offered by the reactance tube (Fig. 2). Neglecting the capacitance C_{gp}, $Y_1 = j\omega C$, $Y_2 = 1/R$, $Y_3 = 1/R_p$, $Y_S = 0$; and, consequently,

$$\frac{1}{Y''} = \frac{R_p}{\mu} + j\omega C \frac{RR_p}{\mu}.$$

The output admittance of the reactance tube is then represented by the network of Fig. 2(c), which gives complete information on the actual operation of the tube, usually approximated by reducing the whole admittance to $j\omega CRR_p/\mu$.

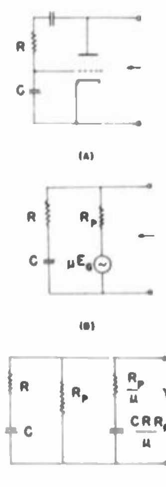


Fig. 2—Network representation of the reactance tube and of its output admittance.

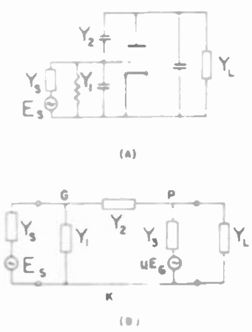
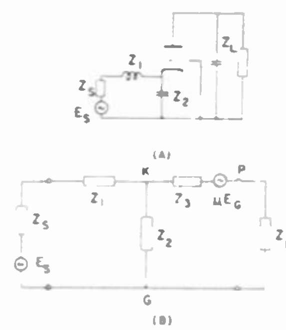
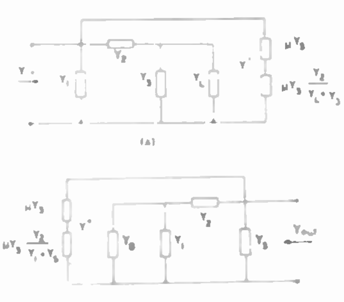
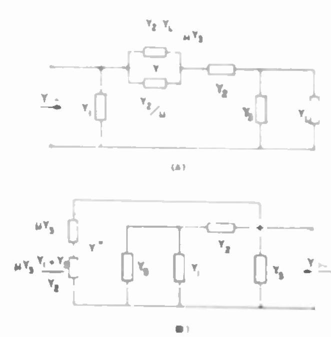
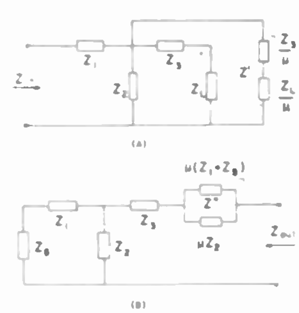
* Received by the Institute, November 15, 1948.

[1] E. L. Chaffee, "Equivalent circuits of an electron triode and the equivalent input and output admittances," Proc. I.R.E., vol. 17, pp. 1633–1648; September, 1929.

[2] F. E. Terman, "Radio Engineering," McGraw-Hill Book Co., Inc., New York, N. Y., 1947; p. 365.

TABLE I

NETWORK REPRESENTATION OF Y_{in} (OR Z_{in}), Y_{out} (OR Z_{out}) FOR LINEAR AMPLIFIERS

Type of amplifier	*Grounded cathode amplifier*	*Grounded-plate amplifier*	*Grounded-grid amplifier*
Actual and equivalent circuits			
Input admittance Y_{in} or impedance Z_{in}	$Y_{in} = Y_1 + \frac{Y_2(Y_3 + Y_L)}{Y_2 + Y_3 + Y_L} + \mu Y_3 \frac{Y_2}{Y_2 + Y_3 + Y_L}$	$Y_{in} = Y_1 + \frac{1}{\frac{1}{Y_2} + \frac{1}{Y_3 + Y_L} + \frac{\mu Y_2}{Y_2(Y_3 + Y_L)}}$	$Z_{in} = Z_1 + \frac{1}{\frac{1}{Z_3 + Z_L} + \frac{1}{Z_2} + \frac{\mu}{Z_3 + Z_L}}$
Output admittance Y_{out} or impedance Z_{out}	$Y_{out} = Y_3 + \frac{Y_2(Y_1 + Y_S)}{Y_1 + Y_2 + Y_S} + \mu Y_3 \frac{Y_2}{Y_1 + Y_2 + Y_S}$	$Y_{out} = Y_3 + \frac{1}{\frac{1}{Y_2} + \frac{1}{Y_1 + Y_S}} + \frac{\mu}{Y_2} \frac{Y_3}{\frac{1}{Y_2} + \frac{1}{Y_1 + Y_S}}$	$Z_{out} = Z_3 + \frac{1}{\frac{1}{Z_2} + \frac{1}{Z_1 + Z_S}} + \frac{1}{\frac{1}{\mu Z_2} + \frac{1}{\mu(Z_1 + Z_S)}}$
Network representation of Y_{in} (or Z_{in}), Y_{out} (or Z_{out})			
Admittances (or impedances) of added branches	$Y' = \mu Y_3 \frac{Y_2}{Y_2 + Y_3 + Y_L}$ $Y'' = \mu Y_3 \frac{Y_2}{Y_1 + Y_2 + Y_S}$	$Y' = \frac{Y_2}{\mu}\left(1 + \frac{Y_L}{Y_3}\right)$ $Y'' = \frac{\mu}{Y_2} \frac{Y_3}{\frac{1}{Y_2} + \frac{1}{Y_1 + Y_S}}$	$Z' = \frac{Z_3 + Z_L}{\mu}$ $Z'' = \mu Z_2 \frac{Z_1 + Z_S}{Z_1 + Z_2 + Z_S}$

In Table I are similarly indicated the network representations for the grounded-plate and the grounded-grid amplifier; for the latter, impedances have been used instead of admittances because of formal convenience. Specific examples could be worked out easily; for instance, it may be shown[3,4] that the input conductance of a cathode follower with capacitive load is negative.

LUCIO M. VALLESE
Duquesne University
Pittsburgh, Pa.

[3] K. Schlesinger, "Cathode follower," PROC. I.R.E., vol. 33, pp. 843–855; December, 1945.

[4] H. Reich, "Input admittance of cathode-follower amplifiers," PROC. I.R.E., vol. 35, pp. 573–576; June, 1947.

Contributors to Proceedings of the I.R.E.

John T. Bangert (S'42–A'44) was born in Chicago, Ill., in 1919. He received the B.S. degree in electrical engineering from the University of Michigan and the M.S. degree from Stevens Institute of Technology. During the first half of 1942, he was a teaching fellow and graduate student at the University of Michigan, after which he became a member of the technical staff of Bell Telephone Laboratories. At the Laboratories, during the war he was engaged in the design and development of electronic equipment for the armed forces, and now is occupied with research in the field of high-speed signaling.

John T. Bangert

Mr. Bangert is an associate of Sigma Xi and a member of Tau Beta Pi, Eta Kappa Nu, and Phi Kappa Phi.

❖

Edward J. Barlow (SM'48) was born in East Orange, N. J., on September 5, 1920. He received the bachelor of electrical engineering degree from Cooper Union in 1942, after which he joined the Sperry Gyroscope Company, where he served as a product engineer on gyro-compasses. In 1943 he entered the radio engineering group of Sperry's research laboratories, engaged in theoretical studies of noise, doppler radar, and klystrons. In 1945 he carried out a technical mission on microwave radar and tube development in England and France.

Edward J. Barlow

Mr. Barlow was awarded the Sperry Graduate Scholarship for 1945–1946, and attended Columbia University. As a senior project engineer, he has directed basic radar studies, and at present he serves as a consultant in advanced development of klystrons and radar systems. He has presented various papers before the IRE radar symposium, the IRE tube conferences at New Haven, Conn., and Ithaca, N. Y., the Instrument Society of America, and the Electrical Engineering Colloquium at Princeton, among others.

❖

R. R. Bush was born in Albion, Mich., on July 20, 1920. He received the B.S. degree in electrical engineering from Michigan State College in 1942. From 1942 until 1946 Mr. Bush was a research engineer at RCA Laboratories, Princeton, N. J., where his work was devoted principally to the development of beam-deflection tubes, mixers and pulsed magnetrons, and to the modulation of magnetrons.

R. R. Bush

Mr. Bush is now an instructor in physics at Princeton University, and is working toward the doctorate. He is a member of Sigma Xi, Tau Beta Pi, Phi Kappa Phi, and the American Physical Society.

❖

A. G. Clavier (M'30–F'39) was born in Cambrai, France, in 1894. He received a degree in electrical engineering from Ecole Supérieure d'Electricité in 1919, and then joined the staff of engineers organized by General Ferrié at the Etablissement Central de la Radiotélégraphie Militaire. He was in charge of research on high frequencies from 1920 to 1925.

A. G. Clavier

In 1929, Mr. Clavier joined Les Laboratoires Standards in Paris, which later became Laboratoire Central de Telecommunications, and has been continuously engaged in research on centimeter and millimeter waves. He was in charge of experiments which, in 1930, resulted in 17-cm wave transmission across the English Channel, and of the developments for the Lympne-St. Inglevert microwave radiotelephone link, which was inaugurated commercially in 1934. He was assistant director of research in 1945, when he was transferred to Federal Telecommunication Laboratories in New York, N. Y., where he now holds the same position. Mr. Clavier has published extensively on high-frequency oscillators, waveguides, and general electromagnetic theory, and has taught field theory and applications of ultra-high frequencies at the Ecole Supérieure d'Electricité.

Mr. Clavier is a past president of the section of the Société des Radioélectriciens dealing with hyperfrequencies, and is a member of the Institution of Electrical Engineers.

William Dite was born in New York, N. Y., on March 2, 1919. He received the B.E.E. degree from the College of the City of New York in 1940. From 1940 to 1943 he was associated with the Signal Corps Laboratories, Fort Monmouth, N. J., working on sound-ranging and radar equipment. Since 1943, Mr. Dite has been employed by the Federal Telecommunication Laboratories, in Nutley, N. J., where his work is largely concerned with communication systems employing pulses.

William Dite

Mr. Dite is a member of the American Institute of Electrical Engineers.

❖

J. S. Donal, Jr. (M'40–SM'43) was born on June 10, 1905, at Philadelphia, Pa. He received the A.B. degree, with High Honors in electrical engineering, from Swarthmore College in 1926, and the Ph.D. degree in physics from the University of Michigan in 1930. From 1930 to 1936, Dr. Donal was associated with the Johnson Foundation for Research in Medical Physics and with the department of pharmacology of the University of Pennsylvania, developing methods and electrical instrumentation for blood-gas analysis and for the measurement of cardiac output in man. In 1936, he joined the research laboratories of the RCA Manufacturing Company in Harrison, N. J., engaging in research on light valves for television reproduction, on the earliest mica-to-glass or metal seals and, later, in research on magnetrons.

J. S. Donal, Jr.

Since 1942, Dr. Donal has been at the RCA Laboratories in Princeton, N. J., working on pulsed and cw magnetrons and on methods for the modulation and control of cw magnetrons. He is a member of Sigma Xi, Sigma Tau, Delta Sigma Rho, and of the American Physical Society.

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For a biography and photograph of Eugene F. Grant, see page 877 of the July, 1948, issue of the Proceedings of the I.R.E.

Frederick R. Gracely was born on September 22, 1911, at Des Moines, Iowa. He received the A.B. degree in 1933 from Drake University, and the B.S. in electrical engineering in 1934 and the E.E. degree in 1944, from Iowa State College.

F. R. GRACELY

Mr. Gracely was engaged in investigation of characteristics of the ionosphere at the National Bureau of Standards from 1938 to 1944. He specialized in the analysis of radio field-intensity measurements and the solar and geophysical effects on radio wave propagation, particularly in connection with the establishment of a world-wide system for short-term predictions of radio transmission conditions. He has made radio, magnetic, and upper atmosphere observations in Greenland and arctic Canada on an expedition in 1941 for the National Bureau of Standards and the Department of Terrestrial Magnetism, Carnegie Institutiion of Washington. Since 1946, he has been engaged in general analysis of radio wave-propagation conditions for the technical information division of the Federal Communications Commission.

Mr. Gracely is a member of Phi Beta Kappa, the American Geophysical Union, and the American Meteorological Society.

❖

Donald W. R. McKinley (VA'39–SM'46–F'48) was born in Shanghai, China, on September, 22,1912. He received the degrees of B.A. in 1934, M.A. in 1935, and Ph.D. in 1938, from the University of Toronto in experimental physics. He joined the staff of the National Research Council, Ottawa, in 1938, where he worked on the development of airborne cathode-ray direction finders, and the establishment of a primary frequency standard. When war broke out, he was given charge of research and development of high-power ground radar, including the model shop production of Canadian MEW radars and allied equipments. He spent six months of 1940 in England as scientific liaison officer for the National Research Council, and from 1943 to 1944 he visited Australia, New Guinea, and India in connection with tropicalization and operational problems. During the war he served as chairman of two interservice committees dealing with tropicalization and wave propagation.

D. W. R. McKINLEY

From 1944 to 1947 Dr. McKinley was connected with the development of radio aids to navigation, and represented the National Research Council at the Commonwealth and Empire Conferences on Radio for Civil Aviation and the Provisional International Civil Aviation Organization. He was chairman of the IRE Ottawa Section from 1946 to 1947.

In August, 1947, he initiated a research program on the study of meteors and the physics of the upper atmosphere by radar techniques, in close co-operation with P. M. Millman, of the Dominion Observatory, who has supervised the visual and photographic phases of the combined project.

❖

Norman T. Lavoo (A'41–M'45) was born in St. Louis, Mo., on October 12, 1918. He received the B.S. degree in electrical engineering from Washington University in 1940. Since that time he has been associated with the General Electric Company, working in several of their development laboratories. He completed the three-year General Electric Advanced Engineering Program in 1943. At present he is in the electronics group of the Research Laboratory.

NORMAN T. LAVOO

❖

Peter Mackenzie Millman was born on August 10, 1906, in Toronto, Ont. Canada. He received the B.A. degree from the University of Toronto in 1929, majoring in astrophysics. The degree of A.M. was obtained at Harvard in 1931, and he received the Ph.D. in astronomy at Harvard in 1932.

P. M. MILLMAN

From 1933 to 1945, Dr. Millman was on the staff of the Department of Astronomy at the University of Toronto, where is located the large 74-inch telescope of the David Dunlap Observatory. He was on leave of absence in the Royal Canadian Air Force from January, 1941, serving first as a navigation officer, and was later in charge of operational research for the RCAF, being retired with the rank of Squadron Leader in June, 1946. Since then he has been on the staff of the Dominion Observatory, Ottawa.

Dr. Millman's chief field of research since 1929 has been meteoric astronomy, and, in particular, the photographic study of meteors and their spectra. He has also carried out work in the field of stellar radial velocity determination, and he made spectrographic observations from the air of the total eclipse of the sun on July 9, 1945.

❖

Philip F. Panter (A'43–SM'48) was born in 1908 in Poland. After early schooling in Tel-Aviv, Palestine, he later received at McGill University, Montreal, Que., Canada, the following degrees: B.Sc., 1933; B.Eng. in electrical engineering, 1935; and Ph.D. in physics, 1936. He continued research in spectroscopy at McGill for an additional year.

P. F. PANTER

After teaching mathematics and physics in Palestine for a year, Dr. Panter returned to Canada as assistant professor of mathematics and physics in the evening division of Sir George Williams College in Montreal. He served also on the staff of the physics department of McGill University until late 1945.

Early in 1941, Dr. Panter joined the transmitter department of the Canadian Marconi Company in Montreal. In October, 1945, he was appointed senior engineer, responsible for the development of FM broadcast equipment, at Federal Telephone and Radio Corporation. He later transferred to Federal Telecommunication Laboratories, and is now in charge of the theoretical group of the communications division. Dr. Panter is a member of the Radio Club of America.

❖

Sidney Rosen was born on December 18, 1907. He received the B.S. degree in electrical engineering from the University of Southern California, and joined the technical staff of the Bell Telephone Laboratories in 1930. From 1930 to 1941, he was engaged in the development of test equipment for the testing and maintenance of voice-frequency and carrier-frequency telephone systems. From 1941 to 1946, he was a project engineer for the development of microwave test equipment used to test and maintain radar equipment supplied to the armed forces.

SIDNEY ROSEN

Since January, 1947, Mr. Rosen has been technical consultant of the firm of Rosen and Rosen, Los Angeles, Calif.

Institute News and Radio Notes

West Coast Members Cited by Government

The Institute of Radio Engineers was given a citation by the U. S. Army, Navy and Air Force at the IRE West Coast Convention, held in the fall of 1948. The following message was received from Major General Spencer B. Akin, Chief Signal Officer of the Army; Rear Admiral Earl E. Stone, Chief of Naval Communications; and Major General F. L. Ankenbrandt, Director of Communications of the Air Force:

The U. S. Army, Navy, and Air Force salute you, the western members of the Institute of Radio Engineers at your 1948 Western Convention, and congratulate you on your outstanding professional achievements during the past year. Your work has resulted not only in a manifold expansion of the communications and electronics industry, but also may become an important cornerstone in the structure of a permanent peace.

AFCA Meet in Washington

The third annual meeting of the Armed Forces Communications Association, an organization made up of civilians and military members to ensure that the Navy, Army, and Air Force will have the best in communications, radar, and photography, was held in Washington on March 28 and 29.

The first day's activities were climaxed by the annual banquet, at which many leaders and other distinguished government officials spoke, including Admiral Louis E. Denfeld, Chief of Naval Operations, Senator Millard F. Tydings, chairman of the Senate Armed Services Committee, and Representative Carl Vinson, chairman of the House Armed Services Committee. David Sarnoff, president of the AFCA, introduced the speakers.

IRE-RMA Will Hold Spring Meeting in Philadelphia

The Fourth Annual Spring Meeting sponsored jointly by The Institute of Radio Engineers and the Radio Manufacturers Association, will be held from April 25 through April 27 at the Benjamin Franklin Hotel in Philadelphia, Pa. At the first technical session, under the leadership of A. N. Curtiss, papers on a 3-kw medium-frequency transmitter utilizing iron-core interstage and output circuits, the use of the cavity resonator in the mobile communications field, the symmetron 50-kw FM broadcast amplifier, and an instantaneous deviation control for phase modulation transmitters will be read.

The second session, with M. R. Briggs as chairman, will cover television recording technique, the utiloscope—a new television visual modulator—and the reality of invisible forces. O. W. Pike will head the final session, which will include papers on high-efficiency coolers for forced-air-cooled power tubes, the audio power amplifier with positive and negative feedback, longitudinal interference in audio circuits, and a commercial PTM telephone microwave link.

Stuart L. Bailey, President of the IRE, will speak at the annual dinner on April 26.

NEW STANDARDS AVAILABLE

Standards on Radio Receivers; Methods of Testing AM Broadcast Receivers (1948, vi+24 pages, 8½×11 inches) is now available for $1.00 per copy. Please address orders to The Institute of Radio Engineers, Inc., 1 E. 79 Street, New York 21, N. Y., and enclose remittance.

Calendar of COMING EVENTS

Annual Symposium, Engineers Council of Houston, Houston, Tex., April 2, 1949

Semiannual Convention, Society of Motion Picture Engineers, New York City, April 4–8

NAB Engineering Conference, Chicago, Ill., April 6–9

AIEE Conference on Electron Tubes, Buffalo, N. Y., April 11–12

AIEE Southwest District Meeting, Dallas, Tex., April 19–21

Third Annual Spring Conference, Cincinnati Section, IRE, Cincinnati, Ohio, April 23

IRE-RMA Spring Meeting, Philadelphia, Pa., April 25–27

IRE-URSI Spring Meeting, Washington, D. C., May 2–4

Twentieth Anniversary Meeting, Acoustical Society of America, New York City, May 5–7

AIEE Summer General Meeting, Swampscott, Mass., June 20–24

AIEE Pacific General Meeting, San Francisco, Calif., August 23–26

1949 IRE West Coast Convention, San Francisco, Calif., August 29–September 1

AIEE Midwest General Meeting, Cincinnati, Ohio, October 17–21

1949 National Electronics Conference, Chicago, Ill., September 26–28

1950 IRE National Convention, New York, March 6–9

Technical Committee Notes

Material for the 1948 Annual Review was submitted by the **Annual Review Committee** to the Editor, who commended the excellent work. It was published in the March issue of the Proceedings and reprints may be made available. . . . The **Electroacoustics Committee**, in co-operation with the ASA and the Acoustical Society of America, has prepared an acoustical glossary, which has been published as a proposed ASA standard. Copies have been circulated to the members of all IRE technical committees with a request that their comments be sent to chairman Eginhard Dietze. The glossary will be submitted to the IRE Standards Committee for approval shortly. . . . At the **Electron Tubes and Solid State Devices Committee** meeting held on January 20, J. A. Morton was elected Chairman of the 1949 Electron Tube Conference, to be held at Princeton University two days during the week of June 20. The 1950 Electron Tube Conference will be held at the University of Michigan. . . . The **Industrial Electronics Committee's** definitions on induction and dielectric heating are ready for submission to the Definitions Co-ordinating Subcommittee. The newly formed Subcommittee on Measurement of Dielectric Materials will work out the measurement technique on frequencies from 1 to 100 Mc and also above 100 Mc in co-operation with the American Society for Testing Materials. . . . The **Modulation Systems Committee**, which met on January 21 and February 15, is continuing to work on the formulation of definitions in the field of pulse-code modulation and modulation theory. . . . The **Nuclear Studies Committee** met on January 26; H. H. Goldsmith and R. E. Lapp will represent the Committee on the planning group for the next nucleonics conference to be held in co-operation with the AIEE. The conference will probably be held in the fall. . . . The first meeting of the new **Sound Recording and Reproducing Committee** was held on February 4. S. J. Begun, the Chairman, requested the attendance of H. A. Chinn, J. E. Keister, and Eginhard Dietze, chairmen of the Audio Techniques, Video Techniques, and Electroacoustics Committees, respectively. The Committee will include within its scope the work which had already been started in the former Audio and Video Techniques Committee in the field of audio recording, and the meeting's purpose was to ascertain the names of people qualified and willing to participate in its work. . . . The **Antennas and Wave Guides Committee** and the **Wave Propagation Committee** have initiated a petition for the formation of a professional group on Antennas and Wave Propagation, and the formation of this group under the IRE Professional Group System has been authorized.

JTAC PROCEEDINGS NOW AVAILABLE

Volumes I and II of the Proceedings of the Joint Technical Advisory Council are now available at the following rates:

Volume I	Utilization of Ultra-High Frequencies for Television (Docket 8976, September 20, 1948)	$7.50
Volume II	Allocation Standards for VHF Television and FM Broadcasting (Docket 9175, December, 1948)	$3.00

IRE Awards, 1949

Medal of Honor

Ralph Bown

"For his extensive contributions to the field of radio and for his leadership in Institute affairs."

Morris Liebmann Memorial Prize

Claude E. Shannon

"For his original and important contributions to the theory of the transmission of information in the presence of noise."

B. J. Thompson Memorial Award

R. V. Pound

"For his paper in the December, 1947 Proceedings, entitled 'Frequency Stabilization of Microwave Oscillators'."

Fellow Awards

Herman A. Affel

(Left) "For his contributions to the communications art, and his guidance of important developments in carrier systems for multiplex telephone and television transmission."

K. C. Black

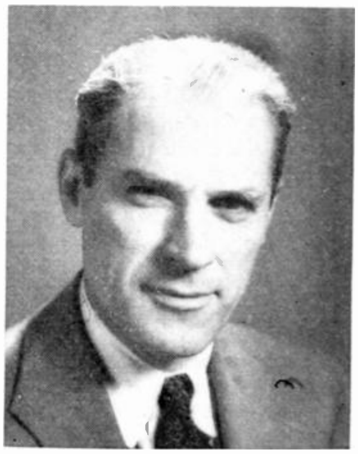
(Right) "For his outstanding wartime work on radio countermeasures and his many contributions to the design of coaxial-cable transmission systems."

J. E. Brown

"For his contributions in the field of broadcast receiver design."

Cledo Brunetti

"In recognition of his pioneering work on printed circuits."



Wendell L. Carlson

"In recognition of his contributions over many years to the development of radio receivers and their components."

Philip S. Carter

"For his many contributions in the fields of radio transmission and communication systems."

F. E. d'Humy

"In recognition of his long service in the communications field and for pioneering in the application of radio relays to telegraph message service."

John N. Dyer

"For administrative and technical contributions to radio, including polar-expedition communications and important wartime radio countermeasures."

L. A. Gebhard

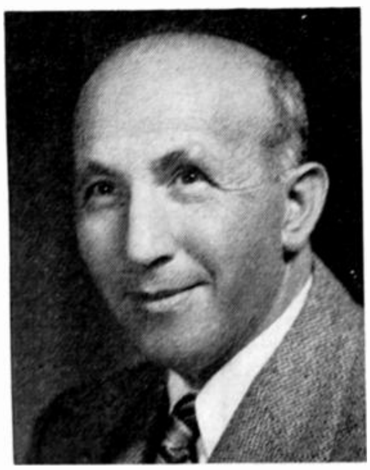
"For his pioneering work in the military application of radio."

Thomas T. Goldsmith, Jr.

"For his contributions in the development of cathode-ray instrumentation and in the field of television."

Frederick W. Grover

"For his long activities and contributions in the field of electrical units and measurements, and for his publications."



Ernest A. Guillemin

"For outstanding work in the field of electric circuit analysis and synthesis, and for his inspired leadership as a teacher."

Ross Gunn

"For his long service and many technical contributions in the radio and electronics fields."

Andrew V. Haeff

"For his contributions to ultra high-frequency radio tubes and electronics."

L. C. Holmes

"In recognition of his leadership in the design and manufacture of radio broadcast receivers."

J. Kelley Johnson

"For his contributions to theory and practice in the field of magnetic recording."

S. R. Kantebet

"For his services as an educator, engineer, and administrator in the fields of radio and cable communication in India."

William B. Lodge

"For his many contributions to broadcast engineering and in particular for his work in the field of frequency allocations."

Keith A. MacKinnon

"For his technical contributions in Canada to the theory and design of transmitting antennas and the development of a coverage plan for a national network."

Harry F. Olson

"For his outstanding developments and publications in the fields of acoustics and underwater sound."

George D. O'Neill

"For his work in electron-tube theory and design."

Leonard S. Payne

"For his contributions in Canada to the field of international communications."

Lloyd M. Price

"For his contributions to the development, production, and application of electron tubes in Canada."

HERBERT J. REICH

"For his contributions as a teacher and author in the radio and electronics field."

JOHN D. REID

"For his developments in radio-frequency circuits."

KARL SPANGENBERG

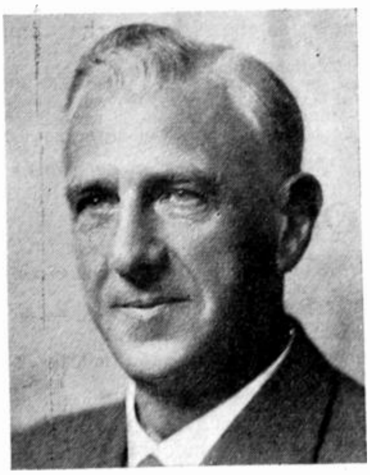
"For his many technical contributions, particularly his analytical work on vacuum tubes."

GEORGE E. STERLING

(Left) "In recognition of his long public service in the radio communication field and, in particular, for the organization and operation of radio wartime intelligence activities, which were of significant importance."

CHARLES E. STRONG

(Right) "For his pioneering work in the radio equipment design and development field, particularly broadcasting transmitters, both medium and high-frequency, and his many wartime contributions in England."



FRANZ TANK

"For his contributions to the field of radio education in Switzerland, and his accomplishments in ultra-short-wave communications."

W. N. TUTTLE

"For his application of sound theoretical principles to the design of commercial measuring equipment."

IRVIN R. WEIR

"For his pioneering work in the development and application of transmitting equipment for higher frequencies and higher power."

Industrial Engineering Notes[1]

German TV Patents Released by Government

Six U. S. Letters Patent relating generally to television image projection devices and tubes are now available for licensing by the Office of Alien Property on a royalty-free, nonexclusive basis for an administrative fee of $15.00 per patent. Titles to these patents were formerly held by Manfred von Ardenne, a German national. A list of the six patents, together with licensing information, may be obtained without cost from the Office of Alien Property, Department of Justice, Washington 25, D. C. Copies of the patents are available from the Commissioner of Patents, Washington 25, D. C., for 25 cents each.

Bureau of Standards Developing Small Computer

The National Bureau of Standards is developing a small-scale electronic computing machine to be used until the several large-scale machines now being built become available. The new high-speed machine, to be known as the NBS Interim Computer, will have an electrostatic type of memory based on the standard cathode-ray-tube memory device developed at Manchester University in England. Expected to be ready for testing by this summer, the machine will perform a substantial portion of the computation work of the Bureau's Laboratories.

[1] The data on which these NOTES are based were selected by permission from "Industry Reports," issues of March 5, 12, and 17, 1949, published by the Radio Manufacturers' Association, whose helpful attitude in this matter is hereby gladly acknowledged.

Navy to Build Underwater Laboratory

The U. S. Navy plans to build a permanent Underwater Sound Reference Laboratory at Orlando, Fla., to be used for testing and evaluating submarine and antisubmarine underwater sound equipment. Operated under the direction of the Office of Naval Research, the laboratory is intended to provide an independent authority to which sonar equipment may be referred for evaluation.

Portable Teletypewriter Developed by Army

Portable teletypewriter equipment so light that a parachutist can carry it on a jump from an airplane has been developed and adopted by the U. S. Army Signal Corps. Weighing only 45 pounds, compared with the current field equipment which weighs 225 pounds, the new portable machine is but one-fourth the size of the old, has 300 fewer parts, is considerably stronger, and consequently requires far less maintenance. It is capable of transmitting and receiving messages 66 per cent faster than existing types, and will operate on both wire and radio circuits.

Authorized Radio Stations Increase 11 Per Cent in 1948

Authorized radio stations in the various services rose to almost 140,000 in 1948, representing a gain of 15,000 for the year. Broadcast authorizations passed the 4,000 mark, and nonbroadcast station grants exceeded 135,000, not counting associated mobile units.

Although FM authorizations decreased by 44 in 1948, FM and television stations increased by 165 and 51, respectively. There were 10 additional noncommercial educational grants, and the new facsimile service started off with two authorizations.

FM Stations on the Air

A total of 738 FM Stations were on the air as of February 11, including 30 noncommercial educational outlets. New FM stations began program operations in the following states: *Calif.*, Berkeley (KRE-FM), and *N. J.*, Atlantic City (WFPO-FM) and Elizabeth (WPOE).

Television News

Television stations operating number 57. There were 66 construction permits outstanding and 314 applications pending before the FCC as of February 11.

Books

Nuclear Radiation Physics, by R. E. Lapp and H. L. Andrews

Published (1948) by Prentice-Hall, Inc., 70 Fifth Ave., New York 11, N. Y. 480 pages, 7-page index, xiv pages. 183 figures. 5¼×8¼. $6.00.

The age of the atom or atomic nucleus, born in the 1890's, undoubtedly came to full flower at some time between December 2, 1942, and August 6, 1945. With that flowering, information that had previously been the concern of a few specialists became of pressing importance to many people in their everyday life. This book constitutes one excellent step in the dissemination of such material. The authors have an excellent vantage point from the very center of activity in the atomic sciences, and have utilized it to give a well-balanced, completely up-to-date picture of the field of nuclear physics.

Those already specialists in the field do not make up the intended audience, but rather the large number of engineers, doctors, and others who need to know the broad outlines of nuclear physics without pursuing the mathematical intricacies of the detailed theory. With the needs of this group in mind, the authors have included background material from other branches of physics in the first few chapters. Perhaps the most outstanding feature of the book is the treatment of the new field of health physics that has been the inevitable adjunct of the large-scale production and use of radioactive material.

According to the preface, the book is the outgrowth of a wartime training manual in radiology. The deft manner in which concepts are presented is evidence of much trial and revision in the light of student reaction. Errors and misprints have been held to an apparently irreducible minimum. All in all, the book may be highly recommended as an authoritative and readable text in this newest of fields.

S. N. Van Voorhis
University of Rochester
Rochester, N. Y.

Radio Aids to Navigation, by R. A. Smith

Published (1948) by the Macmillan Co., 60 Fifth Ave., New York, N. Y. 110 pages, 2-page glossary, 2-page index. xii pages. 37 figures. 6×8¾. $2.50.

This book is a simple but complete nomenclature of the various radio aids to navigation which have been developed during the war, and will be quite helpful as a reminder to those who will need to find quickly the most important characteristics of "Gee," Loran, Eureka, "Rebecca," etc., with more emphasis on those systems originated in or supported by Great Britain. In most cases, a little history of the origin of the system is given, as well as its applications and usefulness.

The main text is preceded by an introduction on the situation of navigational aids before the war. This preface might be considered somewhat too brief, as it is from the research, development, and talent available at the start of the war that the new developments and achievements sprang.

"Gee" and loran are more completely described than the other systems. The information is always clearly presented, and in such a manner that it is possible for the nonspecialized reader to become familiar with the elementary characteristics of each system. Although no schematics or detailed descriptions are given, a few interesting photographs are included.

The chief drawback of the book is that it does not deal with new developments which started at the end of the war for the practical application to commercial flying of the new techniques developed during the war. Moreover, there is no indication of which systems will be most useful in the postwar period.

On the whole, the book will prove a useful volume to the specialist looking for the various characteristics of systems listed, and for the radio engineer who needs information on the techniques involved in the radio aids to navigation developed during the war.

H. Busignies
Federal Telecommunication Laboratories
Nutley, N. J.

Television Production Problems, by John F. Royal

Published (1948) by the McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 18, N. Y. 155 pages, 14-page glossary, 9-page index, xi pages. 16 figures. 5¼×8. $2.50.

This book is a compilation of the lectures delivered at the Columbia University "Television Production Problems" course by eight employees of the National Broadcasting Co. All aspects of television are covered, but, since the book is written primarily for individuals interested in production, the scientific material is necessarily on the most elementary level, and, therefore, of little value for radio engineers.

Quality Control in Industry; Methods and Systems, by John G. Rutherford

Published (1948) by the Pitman Publishing Corp., 2 W. 45 St., New York 19, N. Y. 197 pages, 3-page index, xvii pages. 70 illustrations. 34 tables. 6×9¼. $3.50.

This book, written from the practical quality-control viewpoint, is somewhat unusual in that the fundamentals of practical quality control are given in Part I, followed by a discussion in Part II of the more important statistical quality-control techniques which may be applied to achieve the desired over-all results. It is excellent for students who need an introduction to quality-control organizational methods and practices in industry, as well as for those in industrial quality-control work who need a simple and comprehensive guide to theoretical methods and procedures.

The first part covers general principles and administrative practices for a quality-control department, and includes such general subject headings as: functions and responsibilities, organization economics of quality control, control of quality department costs, records and reports, personnel training, relations to other departments, and relations to customers' representatives. These eight chapters contain much excellent material, even though some of the forms shown and procedures recommended are applicable primarily to a particular industry.

Presented in easy-to-read language, the material is accompanied by clear-cut statements of the problems involved, and is supported by illustrations, organization charts, and graphs. Those chapters relating to organization, control of quality-control department costs, and personnel training are really noteworthy; the chapter on relations to other departments is condensed too much to be of as much value as it should.

One feature of the book jars throughout, and that is the indiscriminate and interchangeable use of the terms "inspection" and "quality control." Inspection is a means of determining whether something has been done in accordance with specification requirements. Quality control, on the other hand, is an important factor in determining if these same specification requirements were established properly in the first place.

The second part of the book illustrates and discusses several statistical quality-control tools, which, if applied properly and pursued carefully, should result in improved product control and attendant lowered product cost. The chapters in this section cover principles of statistical methods, sampling inspection, process control, specialized applications, and the practical applications of statistical methods. Clearly written and easily understood, the material should well serve the needs of the average quality-control engineer.

The author is successful in demonstrating that statistical quality-control methods are a necessity in industrial production. The information presented, while not all-inclusive, is sufficient to serve as a guide for the quality-control engineer, and as a reference manual for the industrial engineer, the executive in charge of quality control, and for supervisors in charge of inspection groups.

On the whole, the book is well worth the price. The author is a man who has had considerable experience in the quality control field, from the standpoint of both practical and theoretical consideration. The results of many of his experiences are indicated, and should assist the reader to avoid many of the pitfalls which may open before him.

JEROME R. STEEN
Sylvania Electric Products Inc.
Flushing, L. I., N. Y.

Television Receiver Construction

Published (1948) by Iliffe and Sons, Ltd., London. 47 pages, 57 figures. 7¼×9¾. 2/6.

This book is a reprinting of ten articles which originally appeared in the English publication *Wireless World* during the year 1947. Primarily of a constructional nature, the articles describe the building of a television receiver for the home constructor.

Unfortunately, all the information presented is based upon British television standards, which differ materially in many respects from the corresponding U. S. standards. Furthermore, the television receiver as described covers only the single television channel currently in use in the British Isles. These are serious limitations for the American constructor, since considerable adaptation will be needed to make the design suitable for local use. A similar, though perhaps less fundamental, disadvantage is that British components are described, and the equivalents of these (particularly tubes) are not readily obtainable here.

The book is well written, and the discussion of the circuit details and operation is clearly handled. For those well versed in the television receiver art who are interested in finding out how foreign designs compare with our own products, this book should be particularly worth while. In addition, certain of the constructional methods suggested are ingenious and well described, and would, in themselves, be useful to a constructor. The book is apparently up to date and, within the limitations mentioned, represents interesting and profitable reading.

F. J. BINGLEY
Radio Station WOR
New York, N. Y.

Rider Public Address Equipment Manual, Volume I

Published (1948) by John F. Rider, Inc., 480 Canal St., New York 13, N. Y. Over 2,000 pages, illustrated, plus 100-page illustrated pamphlet containing explanation and index. 8½×11. $18.

This enormous volume, published in convenient loose-leaf format, offers schematic diagrams and accompanying service notes on audio amplifier models for all uses, manufactured from 1938 to date and representing material from 147 different manufacturers. Different circuit innovations have been analyzed, and such topics as response curves, preamplifiers, equalizers, mixer circuits, tone compensation, and volume expansion and compression circuits have been included in the book. The accompanying pamphlet is virtually a book in itself, containing, besides a chapter on basic considerations, material on input systems, mixer circuits, tone compensation and coupling, volume expansion and compression, push-pull circuit, inverse feedback, and the output transformer and speaker system.

Molybdenum: Steels, Irons, Alloys, by R. S. Archer, J. Z. Briggs, and C. M. Loeb, Jr.

Published (1948) by the Climax Molybdenum Co., 500 Fifth Ave., New York 17, N. Y. 363 pages, 26-page index. 188 figures. 6×9¼. Free of charge.

This book consists of ten sections, treating of the technical effects of molybdenum on steel, cast steel, and cast iron; the fundamental effects of heat treatment of microstructure, including the transformation of austenite, the effect of molybdenum on microstructure and in ferrite and low carbon iron alloys, acicular cast irons, the partition of molybdenum in steel, the diffusion of molybdenum, and the iron-carbon, molybdenum system; the addition of molybdenum to iron and steel; wrought alloy engineering steels; wrought corrosion resistant steels; wrought steels for elevated temperature service; tool steels; cast iron; and special purpose and nonferrous alloys. Bibliographies are given at the end of each section, and seven appendixes, covering the chemical composition ranges for alloy engineering steels in the U.S.A., the chemical composition ranges for alloy engineering steels, the chemical composition ranges for CETAC alloy engineering steels, the determination of "equivalent rounds," the maximum allowable working stresses at temperature, conversions, and the physical properties of metallic molybdenum.

American Electricians' Handbook, by Terrell Croft

Published (1948) by the McGraw-Hill Book Co., 330 W. 42 St., New York 18, N. Y. 1,735 pages, 34-page index, xv pages. 1,372 illustrations. 5×7¼. $6.00.

The object of this book, now in its sixth edition, has been to collect such information as will enable workers in the practical fields of electricity—wiremen, contractors, linemen, superintendents of small plants, operators, and construction engineers—to select and install commercial electrical apparatus and materials intelligently for the performance of given services, and to qualify them for operating the equipment after it has been installed.

Post War Audio Amplifiers and Associated Equipment
Post War Communications Receiver Manual

Published (1948) by Howard W. Sams and Co., Inc., 2924 E. Washington St., Indianapolis 6, Ind. 8½×11.

These two Photofact Publications give servicing data in pictorial and diagrammatic form for postwar radio equipment.

"Post War Audio Amplifiers and Associated Equipment," which sells for $3.95, describes a representative group of postwar FM tuners, amplifiers, recorders, and reproducers, and points out the advantages to the radio service technician of offering his services in the selection and installation of custom built sound equipment.

"Post War Communications Receiver Manual," retailing at $3.00, gives service data on almost all the communications receivers produced from the end of the war until the middle of 1948, showing how they are made and how to align and adjust them for maximum efficiency.

Most-Often-Needed 1949 Radio Diagrams and Servicing Information, compiled by M. N. Beitman

Published (1949) by Supreme Publications, Chicago, Ill. 160 pages, including 3-page index. 8½×11. $2.50.

This manual, volume 9 in a series which describes radio sets dating from 1926 up to the present, gives extensive schematic diagrams and repair data for most 1949 radio sets. The models of thirty-nine different manufacturers are covered.

Underwater Explosions, by Robert H. Cole

Published (1948) by the Princeton University Press, Princeton, N. J. 426 pages, 6-page bibliography, 5-page index. 121 figures. 6×9. $7.50.

This book is an attempt to supply a reasonably comprehensive account of the research on underwater exlposions carried out in the years 1941 to 1946, a time at which the results were not readily available to scientists. Valuable both to workers in the field of underwater explosions and to others interested in the basic physical processes involved the book develops necessary hydrodynamical relations from first principles before covering the detonation process in explosives, the theory of the shock wave and its measurements, measurement of underwater explosion pressures, photography of underwater explosions, motion of the gas sphere, secondary pressure waves, and surface and other effects.

New Publications

"Scientific and Industrial Glass Blowing and Laboratory Techniques," by W. E. Barr and Victor J. Anhorn, will be published this year by the Instruments Publishing Co., Inc., 1117 Wolfendale St., Pittsburgh, Pa. The book, containing 388 pages and 212 illustrations, is priced at $6.00 postpaid in the United States, $6.50 outside. . . . **"Analysis of Temperature, Pressure, and Density of the Atmosphere Extending to Extreme Altitudes,"** by George Grimminger, has been published at $2.85 by the Rand Corp., 1500 Fourth St., Santa Monica, Calif. This monograph was prepared to satisfy the constantly arising need in various investigations for information concerning the properties of the upper atmosphere at extremely high altitudes. It represents a first attempt to determine the vertical distribution of atmospheric properties over a range of altitude extending from sea level out to interplanetary space, and brings together in one place the various types of information which bear on the problem. Values of the properties of the atmosphere from sea level up to extreme heights of the order of 5,000 to 10,000 miles or more have been derived both at the equator and at middle latitudes. From sea level up to the height of the F_2 layer of the ionosphere the calculations are based on the best information available in the literature concerning the vertical temperature distribution, including the ionosphere temperatures which are derived by radio wave soundings. However, in view of the rather complete lack of knowledge concerning atmospheric conditions in regions above the F_2 layer, calculations for these regions have been carried out on the basis of three different atmospheric models or theoretical concepts.

IRE People

R. A. Hackbusch (A'26–M'30–F'37) of Toronto, Ont., Canada, was elected president of the Canadian Radio Technical Planning Board at its annual convention in Ottawa, Ont., in 1948, and Gordon W. Olive (A'29–VA'39) was elected vice-president. **R. C. Poulter** (A'30–M'37–SM '43) was returned as director of public relations. During the meeting, **G. C. W. Browne** (SM'45), Controller of Radio in the Transport Department, lauded the Planning Board on the work it has done to keep broadcast channels in Canada from becoming jammed with too many stations near the same frequencies. The Institute of Radio Engineers is one of the contributing sponsors to the Canadian Radio Technical Planning Board.

Mr. Hackbusch was born in Hamilton, Ont., Canada, on September 18, 1900. Educated at the Hamilton Collegiate Institute, he later specialized in electrical engineering. After being associated with the Canadian Westinghouse Co. and then with the Canadian Brandes Co., in 1930 he became chief engineer and factory manager of the Stromberg-Carlson Telephone Manufacturing Co. at Toronto, and in 1940 was appointed vice-president and general manager. In that year he was drafted by the Canadian Government to head the radio division of the government-controlled Research Enterprises, Ltd., at Toronto, but he left in 1943 to return to Stromberg-Carlson as managing director. Mr. Hackbusch served as Vice-President of the IRE during 1944.

Mr. Olive was born in 1898. After graduating from McGill University, he served overseas during World War I with the McGill Seige Battery. In 1923 he built and operated radio broadcasting station CFCO for Semmelack and Dickson, while managing that company's radio department. Joining the staff of the Canadian National Railways' radio department, in 1924, as a radio engineer, he was appointed technical assistant to the director of radio of CNR three years later.

Mr. Poulter (A'30–M'37–SM'43), although born and educated in England, began his career in the electrical and radio field in London, Ont., where he was engaged in electrical construction and wattmeter and instrument repair. While heading the radio department of the Benson and Wilcox Electric Co. from 1922 to 1925, he did early research on electropolygraphs and heat-sound amplifiers at the University of Western Ontario in 1923 and 1924. In 1926 he joined Fada Radio, Ltd., in Toronto as a test engineer. Two years later he was appointed to the staff of Hugh McLean Publications, Ltd., as an editor, resigning in 1936 to become director of education of the Radio College of America and president and managing director of Poulter Publications, Ltd.

Frank Henry Fay (A'38–VA'39), prominent member of the Portland, Ore., IRE Section, was killed recently in the crash of a privately-owned airplane which he was piloting.

Mr. Fay was born in Seattle, Wash., on February 29, 1912. In 1932 he entered the employment of the Portland, Ore., Radio Specialty Manufacturing Co., and rose to become foreman of the crystal grinding department. He organized his own firm, the Oregon Electronic Manufacturing Co., in 1939, and managed it, as well as operating the Sentry Crystal Co. These two companies produced large quantities of specialized radio and electronic equipment, including variable voltage regulated power supplies, and fixed and mobile units for the communication systems of the Oregon and Washington state forestry departments and for the Bonneville Power Administration. Mr. Fay also spent some time as staff engineer with the Massachusetts Institute of Technology's Research Construction Corp.

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Elmer D. McArthur (M'42–SM'43–F'45), holder of thirty-nine patents in the field of electronics, has been appointed head of the General Electric research laboratory's high-frequency electronics division. Mr. McArthur was head of the laboratory's ultra-high-frequency vacuum-tube section, which has been expanded into a division to meet increasing needs for pure research in high-frequency electronics.

Born in Salamanca, N. Y., on May 3, 1903, Mr. McArthur holds two degrees from Union College, the B.S. degree in electrical engineering and the M.S. degree in physics. Upon receiving the latter degree in 1925, he joined the General Electric Co. as a student engineer, and then entered the research laboratory to work in the field of electronics.

In 1930 he was transferred to the vacuum-tube engineering department. Twelve years later he and another engineer set up the company's electronics laboratory, which operated during the war years. In 1945 he rejoined the research laboratory a head of the ultra-high-frequency vacuum-tube section, and in that year received the Charles A. Coffin Award, the company's highest recognition, for his work in vacuum-tube design. The following year he received a citation from the U. S. Navy for his wartime work on electronics for military uses.

Earl E. Eldredge (A'31-VA'39) has been named chief engineer of the Press Wireless Manufacturing Co., Inc., of Hicksville, Long Island, N. Y., and West Newton, Mass.

After graduating from Brown University with the electrical engineering degree, Mr. Eldredge served as chief engineer at the Erco Radio Laboratory, chief engineer of Press Wireless's communications division, field engineer at the Mackay Radio and Telegraph Co., and test engineer for the General Electric Co. at Schenectady.

During World War II, Mr. Eldredge designed and supervised the building of high-power transmitters used by the Signal Corps, Navy, OSS, and other government agencies as part of a radio communications network that reached battlefronts in all parts of the world.

❖

Webster F. Soules (A'46), sales manager of the Buchanan, Mich., plant of Electro-Voice, Inc., was recalled by the U. S. Army in January for one month of active duty at the Signal Corps Laboratories in Fort Monmouth, N. J., where Army communications equipment is developed.

Mr. Soules was stationed at Fort Monmouth for two years after he joined the Army in 1940. At the Signal Corps Laboratories there and as Signal Corps member of the Armored Force Board in Fort Knox, Ky., he engaged in developmental work on radio apparatus and installations in Armored Force vehicles. Appointed engineering and executive officer of the Armed Force's Signal Section, he was graduated the following year from the Command and General Staff School at Fort Leavenworth, Kan., with the rank of lieutenant-colonel. He commanded a signal service battalion in Ceylon and subsequently served as signal officer in China at the tactical headquarters. At the termination of war in the Pacific, he became executive officer of the China staging area. In 1946 he joined Electro-Voice.

Previous to joining the Army, Mr. Soules had graduated from the University of Minnesota's School of Electrical Engineering and had spent 17 years with the Northern States Power Co. He was president of the Minneapolis Radio Club and was active in the ARRL and the Army Amateur Radio System.

❖

Frank P. Barnes (A'43-M'44) has been appointed sales manager of broadcast equipment for the transmitter division at General Electric's Electronics Park in Syracuse, N. Y.

A graduate of Stanford University, Mr. Barnes joined General Electric in 1937, taking the engineering test course in Schenectady, N. Y. For over three years he was Western district representative for the electronics department, covering northern California, the Northwest, and Rocky Mountain areas. He also spent a number of years in Seattle, where he specialized in industrial electronics and radio communications for General Electric. Well known in radio and electronics on the West Coast, he taught courses in industrial electronic engineering and radio engineering at the University of Washington. In August, 1948, he was named assistant to the manager of sales for GE's transmitter division.

❖

Henry P. Kalmus (A'39–SM'45), formerly a member of the research laboratory of the Zenith Radio Corp., has been appointed to the staff of the National Bureau of Standards, where he will conduct investigations in advanced electronic techniques in the Ordnance Research Laboratory.

Born in Vienna, Austria, in 1906, Mr. Kalmus attended the Technical University of Vienna from 1924 to 1930, and was awarded an engineering diploma after several years of graduate work in electrical engineering. From 1930 to 1938 he headed the research laboratory of the Orion Radio Corp., a subsidiary of Tungsram, in Budapest, Hungary, designing receivers for Sweden, Belgium, Switzerland, and other European countries. In 1939 he came to the United States as development engineer with the Emerson Radio Corp. in New York City, and left that company in 1941 to conduct research for Zenith. His investigations include the fields of ultra-high frequency, frequency modulation, and television.

❖

Nicholas D. Glyptis (A'42), an expert on electron-beam tubes of all types, and, at the age of 27, perhaps the youngest consulting physicist in the country, has been named director of the Multi-Tron Laboratories, recently established in Chicago.

Formerly assistant chief director of research at the Rauland Corp., Chicago, and later project engineer in charge of electron optics at the Sperry Gyroscope Co., Mr. Glyptis is a specialist in charged-particle trajectories, and in space-charge or space-charge-free configurations or combinations thereof; and he has contributed a number of outstanding inventions and designs to this field. His newest contribution is an electron-beam tube of radical design which uses a special effect of secondary emission and achieves a highly sensitive, inertialess, powerful electron beam.

Mr. Glyptis is a member of the American Association for the Advancement of Science, the AIEE, the Society of Motion Picture Engineers, the American Astronomical Society, and the American Physical Society.

❖

Leon Riebman (S'43–A'44), formerly senior engineer on the research and development staff of the Philco Corp., has become an instructor at the Moore School of Electrical Engineering at the University of Pennsylvania.

Born on April 22, 1920, in Coatesville, Pa., Mr. Riebman received the B.S. and M.S. degrees in electrical engineering from the Moore School in 1943 and 1947, respectively. He was assigned to duty in radar development at the Naval Research Laboratory in 1944, after he had attended the U. S. Navy Midshipman School, and, upon separation from the service, joined Philco in 1946.

❖

F. H. Rockett, Jr. (S'43–A'44), formerly associate editor of *Electronics*, has joined the staff of the Airborne Instruments Laboratory in Mineola, L. I., N. Y. There he will assist in editing and providing technical review for research reports published by the laboratory.

After receiving the B.S.E.E. degree from Lehigh University in 1942, Mr. Rockett worked on the Navy's proximity-fuze project at the Applied Physics Laboratory of The Johns Hopkins University for the latter part of that year. In 1943 he returned to Lehigh University to serve as a laboratory assistant in the department of electrical engineering, but transferred in June of that year to the Columbia University department of electrical engineering, where he was employed as a laboratory instructor.

Appointed to the staff of *Electronics* in June, 1944, Mr. Rockett edited the "Electron Art" department in that publication, and performed other editorial duties centered on the basic sciences, electronic tubes, and circuit theory.

❖

Milton E. Mohr (M'45), a member of the Bell Telephone Laboratories technical staff, has been awarded an honorable mention in Eta Kappa Nu's selection of the outstanding young electrical engineer of 1948.

Mr. Mohr was born in Milwaukee, Wis., on April 9, 1915. He received the bachelor's degree in electrical engineering with honors from the University of Nebraska in 1938, and subsequently joined the Bell Telephone Laboratories, where he first worked with automatic switching, to which he contributed various cold-cathode gas tube circuits, and supervised the trial of electronic devices on the crossbar machine switching system. His research in electron-beam studies during World War II is still on the secret list. Afterward, he worked on frequency modulation, where he was outstanding for his ability to obtain highly stable linear operation over a large percentage swing of carrier from a form of multivibrator, a type of circuit ordinarily regarded as of low precision and stability. Most recently he has been doing research in the field of use of negative resistances and the newly developed transistor.

The holder of thirteen U.S. patents and ten foreign ones, Mr. Mohr is a member of the AIEE, the Summit (N.J.) Association of Scientists, Sigma Xi, Sigma Tau, and Pi Mu Epsilon. At present he is engaged in advanced studies at the Stevens Institute of Technology in Hoboken, N. J.

❖

Raymond F. Foster (SM'41), a development engineer in the General Electric Co.'s receiver division, has been given the company's highest honor, a Charles A. Coffin award, for work of outstanding merit on television receiver development during 1948.

A native of Southampton, Mass., Mr. Foster is a graduate of Northeastern University in Boston. He joined General Electric in 1935 as a student on the test course, and since 1939 has been engaged in television engineering.

Nicholas G. Anton (M'44-SM'44) is no longer president of engineering of the Ampere Electronic Corp., but has formed his own company, the Anton Electronic Laboratories, Inc., in Queens, Long Island, N. Y., to manufacture electronic instruments for the generation and measurement of radiations, vhf equipment, and special vacuum tubes and vacuum devices.

❖

Ralph S. Yeandle (SM'46), General Electric television engineer, has toured South America for six weeks in order to encourage the adoption of U. S. television standards in Latin American countries. Although most of his time was spent in Brazil, where South America's first television station is to be constructed, he also spent some time in other Latin American countries.

Several European countries are also seeking to have their equipment and standards adopted in South America, principally the British and French television companies, whose standards differ radically from those established here by the Federal Communications Commission.

❖

Richard T. Orth (A'31-M'38-SM'43), formerly merchandise manager of the RCA tube department and holder of several patents in the electronic field, has been appointed general manager of the RCA tube department.

Mr. Orth was born on February 20, 1907, in Chicago, Ill. He joined RCA in 1930, after receiving the B.S. degree in electrical engineering from Purdue University. From 1933 to 1938 he headed design groups on cathode-ray and receiving tubes at RCA. Receiving the company's Alfred P. Sloan fellowship in 1938, he took a year's leave of absence to do graduate work at the Massachusetts Institute of Technology.

When the United States entered the war, Mr. Orth was given the assignment of speeding all the company's plant facilities to war production. Later he assumed supervision of war-contract service activity in all RCA Victor plants. In 1944 he was named manager of the Bloomington, Ind., plant, where top-secret proximity fuzes for the armed services were manufactured. The following year he returned to the Harrison, N. J., plant as manager of receiving-tube engineering, becoming merchandise manager for tubes and components in 1947.

Mr. Orth is a member of Sigma Xi and Eta Kappa Nu.

❖

Donald W. Gunn (M'47), formerly a special representative for the Sylvania Electric Products Inc.'s radio tube division, has been appointed assistant to the general sales manager for the same division.

Graduated from Northeastern University with the B.S. in electrical engineering, Mr. Gunn joined Sylvania in 1931 as a tube plant quality engineer at Salem, Mass. Later he was transferred to the Chicago office as midwestern representative for equipment tube sales. Prior to and during World War II, he served as manager of quality control at Sylvania plants in Pennsylvania.

Carrol J. Burnside (A'27-M'38-SM'43), former manager of the Westinghouse Electric Corp's. industrial electronic division, has resigned to organize an independent industrial consultant service with headquarters in Baltimore, Md. He will continue his association with Westinghouse as a consultant.

Born in Des Moines, Iowa, in 1901, Mr. Burnside received his early schooling in western South Dakota, and was graduated with the bachelor of science degree in electrical engineering from the South Dakota School of Mines in 1924. Two years later he received the master's degree, and in 1943 he was awarded the honorary doctor's degree in engineering for his pioneer work in radio.

Mr. Burnside started his radio career in 1924 as a development engineer in the Westinghouse Corp.'s department of radio operations. Assigned to KDKA, the world's pioneer broadcasting station, he designed, built, and operated some of radio's first remote pickup and mobile transmission equipment, and engaged in some of the first frequency-modulation broadcasting tests.

Appointed manager of engineering for the industrial electronics division in 1934, Mr. Burnside had charge of the design and installation of some of the first 50-kw broadcasting stations in the world. He was appointed sales manager for the division in 1939, and became division manager three years later.

During the war, Mr. Burnside supervised the production of more than $400,000,000 worth of radio and radar equipment for the military services, a task which involved expanding output to more than eighty times the prewar production level. For this work he was awarded the U. S. Navy Certificate of Commendation. At the end of the war he directed the reconversion of the industrial electronic division operations to the manufacture and sale of a wide variety of peacetime products.

Active in radio trade and professional societies, Mr. Burnside is a member of the Radio Manufacturers Association's executive committee. He is also a member of the American Society of Naval Engineers and the National Electrical Manufacturers Association.

❖

William E. Shoupp (SM'45), distinguished nuclear physicist, has been named director of research of the Westinghouse Electric Corp.'s new atomic-power division. Part of his duties in guiding the division's activities will be directing the construction and testing of an atomic power plant for the propulsion of naval vessels.

A native of Troy, Ohio, Dr. Shoupp was graduated from Miami University in Oxford, Ohio, in 1931, with the B.A. degree in physics. For the next six years he served as graduate assistant and instructor in physics at the university of Illinois, receiving the M.A. in 1933 and the Ph.D. four years later.

In 1938 Dr. Shoupp joined the Westinghouse Corp. as a research fellow, becoming a research engineer in 1941 and manager of electronics and nuclear research in 1943. During the war years he was in charge of all Westinghouse radar research and development, which included the T-R tube, the resnatron, the standard-frequency cavity, the magnetron, and the crystal rectifier. In the field of nuclear physics, Dr. Shoupp has been associated with the discovery of photofission, the splitting of uranium atoms by high-speed gamma rays, with a commensurate release of large amounts of energy, and the discovery of the threshold of fast neutron fission of uranium and thorium, having published numerous articles on nuclear reactions and nuclear properties.

Dr. Shoupp is a fellow of the American Physical Society and of the Pittsburgh Physical Society, and a member of the AIEE, Sigma Xi, and Phi Beta Kappa.

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Robert E. Moe (S'33-A'35-SM'46) has been appointed division engineer for electronic receiving-tube product lines of the General Electric Co.'s tube divisions.

A native of Appleton, Wis., Mr. Moe is a graduate of the University of Wisconsin, with the B.S. degree in electrical engineering. He joined the test department of General Electric in 1934, and the following year was assigned to work on receiver engineering in radio, television, and radar set design. In 1944 he was transferred to airborne radar research at the transmitter division in Schenectady, and two years later was reassigned to receiver work, this time for the company's government division.

He is a member of Tau Beta Pi and Eta Kappa Nu.

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Edgar H. Felix (J'17-A'19-M'25-SM'43), formerly Washington representative for the Allen B. Du Mont Laboratories' transmitter division, has been promoted to northern district supervisor for the division.

Mr. Felix was born on March 29, 1898, and studied at New York University, Columbia University, and Yale University's Sheffield Scientific School. After the first World War, during which he served as an engineer in the Signal Corps' radio development section, he became associate editor of *Aerial Age Weekly* in 1919. From 1922 to 1924 he directed public relations for radio station WEAF and the American Telephone and Telegraph Co. In 1924 he became technical director of the radio department of N. W. Ayer and Son, and from 1926 to 1941 he acted as radio consultant to broadcast stations, networks, and publishers, at the same time serving as radio director of the National Electrical Manufacturing Co. and as contributing editor of various technical publications. In 1935 he became director of Radio Coverage Reports.

When the United States entered the second World War, Mr. Felix joined the Signal Corps as a captain, rising to the rank of major in 1943. He joined Du Mont after the war.

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Clarence L. Coates, Jr., (S'43-A'43) and **Charles V. Jakowatz** (A'45), former instructors at the University of Kansas and Kansas State College, respectively, have been appointed instructors at the University of Illinois.

Roger M. Wise (A'26–M'30–F'37), a leading authority on electron tubes, and his group of tube engineers are joining the technical staff of the Philco Corp.

Born on April 6, 1898, Mr. Wise has had more than thirty years of experience in the radio industry. After serving as a chief electrician in the Navy in World War I, he completed his education at the University of California, and then worked successively for the Remler, Cunningham, and Grigsby-Grunow Companies. In 1929 he joined what is now Sylvania Electric Products Inc. He was made director of engineering for Sylvania in 1943 and vice-president in charge of engineering a year later, resigning in 1945 to form his own firm of tube consultants and engineers.

While an engineer for Sylvania, Mr. Wise participated in such important new tube developments as 6.3-volt tubes for home and automobile radio use, loktal tubes, and 1.4-volt tubes for battery-operated equipment. During World War II he was active in designing special receiving tubes, cathode-ray tubes, and transmitting tubes, as well as in the development and production of subminiature tubes for the VT proximity fuzes. For this work he received the Naval Ordnance Development Award.

Mr. Wise has been a member of a number of Institute Committees, including Admissions, Papers Procurement, Standardization, and Vacuum Tubes.

❖

Richard M. Somers (A'42-M'42-SM'43), who has been serving since 1936 as assistant chief engineer of the Edison Co.'s Ediphone Division in Orange, N. J., has just been promoted to chief engineer.

Mr. Somers was born on December 22, 1904, in Orange, N. J., and received the E.E. degree from the Rensselaer Polytechnic Institute in 1926. Joining the faculty of Rensselaer at that time, he taught for one year; then became a student engineer with the Radio Corporation of America, where he worked on transoceanic transmission and reception.

In 1926 he joined Edison as a research engineer and later served as chief engineer and factory superintendent of the former lamp division. He is a member of the AIEE, the New Jersey Society of Professional Engineers, and the National Society of Professional Engineers.

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Alfred K. Wright (A'37–SM'47), chief radio engineer at the Tung-Sol Lamp Works, Inc., at Bloomfield, N. J., has been appointed a member of the Joint Electron Tube Engineering Council. Active in the Council's standardization program since its inception, he was formerly chairman of the receiving tube committee.

Dr. Wright was graduated from Northeastern University with the E. E. degree in 1931. A year later he received the Master's degree from Harvard University, and in 1934 the D.Sc. Before joining Tung-Sol in 1937, he was associated with the National Union Radio Corp.

Abe Mordecai Zarem (S'42-A'46), has been selected as the outstanding young electrical engineer of 1948 by the jury of award of Eta Kappa Nu, national honor society for electrical engineers.

Dr. Zarem was born in Chicago, Ill., on March 7, 1917. He entered the Chicago Technical College in 1939, but transferred after one year to the Armour Institute of Technology, now the Illinois Institute of Technology. Valedictorian of his class at his graduation in 1939, Dr. Zarem was given a graduate scholarship in electrical engineering at the California Institute of Technology in Pasadena. While working for the M.S. degree, which he received in 1940, he taught physics, electrical engineering, and mathematics at the Institute. In 1943 he received the doctorate for his research on the physical properties of the electric spark.

Joining the electronic receiver division of the Allis-Chalmers Manufacturing Co. in Milwaukee, Wis., Dr. Zarem developed an electrical control method for monitoring power systems, conducted extensive research on the gaseous conduction of electricity, and developed specialized electrical instrumentation for studying mercury-arc conduction. He also invented an "automatic oscillograph with a memory," which is triggered by random changes in the phenomena being studied.

In 1945 Dr. Zarem returned to the California Institute of Technology to become research engineer and a group leader in work connected with the atomic bomb. Subsequently he joined the staff of the U. S. Naval Ordnance Test Station at Pasadena as head of basic and electrical engineering research for the electronics group.

Named to head the electrical section of the newly formed Physical Research Division in 1947, Dr. Zarem turned again to the study of transient electrical discharges and the development of a method for photographing them. Probably his most outstanding invention is the Zarem camera, with a framing rate up to 100,000,000 per second and effective exposure time down to 0.000,000,001 second.

During the past few years, Dr. Zarem has been acting as consultant to industrial and governmental organizations in varied fields. In 1948 he was appointed chairman of physics research and manager of the new Los Angeles Division of the Stanford Research Institute.

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George F. Devine (M'48) has been appointed assistant to the sales manager of General Electric's electronics department at Syracuse, N. Y.

A native of Philadelphia, Pa., Mr. Devine has been employed by General Electric since 1935. Prior to his new appointment he was commercial engineer for the specialty division, having worked previously on radio receiver design for the electronics department's receiver division.

During the war Mr. Devine was assigned to naval ordnance projects by the receiver division, and in 1945 he received the Naval Ordnance Development Award for his work on antisubmarine electronic devices.

John R. Niles (S'42–A'45), formerly group supervisor at the University of Michigan's Engineering Research Institute, has accepted the position of chief engineer with Radioactive Products, Inc., of Detroit.

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Niles P. Christiansen (A'47), formerly electronics department head of Frazar and Hansen, Ltd., in San Francisco, Calif., has been appointed eastern division manager with headquarters at their newly opened offices in New York, N. Y.

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Harold Goldberg (A'45), recently appointed chief of the National Bureau of Standards' Ordnance Research Section, will be assisted by **Donald P. Burcham** (M'45), who will act as alternate chief of the section, a unit of the newly organized Electronics Division.

Born in Milwaukee, Wis., in 1914, Dr. Goldberg attended the University of Wisconsin, from which he received all four of his degrees: the B.S. in electrical engineering in 1935, the master's degree in 1936, the doctorate in electrical engineering in 1937, and the doctorate in physiology in 1941. From 1935 to 1937 he was a research fellow in engineering and graduate assistant in mathematics, and from 1938 to 1941 a post-doctorate research fellow.

In 1941 he joined the Stromberg-Carlson Co., where he was responsible for the design and development of the SCR-582 radar receiver, the Stromberg-Carlson Mark II airborne vest-pocket modulator, and the SCR-668-T5 radar set. He has also done considerable work in connection with microwave research and communications systems, including television. Leaving Stromberg-Carlson in 1947 to join the Bendix Aviation Corp.'s radio division as principal research engineer, he left the Bendix Corp. to head the NBS Ordnance Research Section.

Dr. Goldberg is a member of the AIEE, the American Physical Society, the American Association for the Advancement of Science, Sigma Xi, and Tau Beta Pi. Four patents have been granted him in the field of electronics, and he has some fifty others pending.

Dr. Burcham was born in Baker, Ore., in 1916. Receiving the B.A. in physics and mathematics from Reed College in Portland, Ore., in 1937, he became a graduate fellow at the University of Wisconsin, and was given the Ph.D. degree in 1942.

From 1941 to 1943 Dr. Burcham was on duty with the U. S. Navy's Ordnance Bureau where he assisted in the instrumentation of the "Operation Crossroads" atomic bomb tests and directed underwater defense research. In 1943 he was appointed to the staff of the Applied Physics Laboratory of the University of Washington in Seattle. He was presented with the Naval Ordnance Development Award in 1945, and the following year joined the staff of the National Bureau of Standards.

Dr. Burcham has conducted extensive research in electronic ordnance devices, especially torpedo exploders and guided missiles. He is a member of the American Physical Society and Pi Mu Epsilon.

Karl Troeglen

Chairman, Kansas City Section

Fred J. Van Zeeland

Chairman, Milwaukee Section

Karl Troeglen (A'30–M'42–SM'43) was born in Kiel, Germany, on August 17, 1908, and came to the United States in 1913. Active in amateur radio work since 1923, he started his commercial career as a shipboard operator on the Great Lakes in 1927.

From 1929 through 1930 he was employed by the Universal Wireless Communications Co., and had risen to be engineer-in-charge of the Plainfield, Ill., plant at the time the company ceased operations. Subsequently he joined the Topeka Broadcasting Co. in Topeka Kan., as chief engineer, and worked there for the next thirteen years.

During the war years, 1943 to 1945, he was on the staff of the Western Electric Co. as a field engineer in the radar division of the Bureau of Ships. In this capacity he headed the engineering group at the Pearl Harbor Navy Yard, and later he held the same position at the Brooklyn Navy Yard in New York. He joined the KCMO Broadcasting Co. in Kansas City, Mo., in the fall of 1945 as technical director in charge of all engineering activity.

During 1941 and 1942 Mr. Troeglen was a member of the National Association of Broadcasters Engineering Committee.

Fred J. Van Zeeland (M'47) was born in Kimberly, Wis., in 1906. After graduating from the Milwaukee School of Engineering in 1928 with the B.S. degree in electrical engineering, he became a member of that school's teaching staff. In 1933 he was named head of the electrical department, and in 1945 became director of the school's college of electrical engineering. He has taken graduate work in education at Northwestern University.

Mr. Van Zeeland has also served as a consultant for several electrical manufacturers. In 1940 he designed a type of self-excited alternator for the Kurz-Root Co. of Appleton, Wis., and in 1941 he developed an electronic motor control system for the Louis Allis Co. of Milwaukee. From 1943 to the present time he has served as a consultant for the Chain Belt Co., of Milwaukee, working on applications of high-frequency heating and the development of electronic sorting equipment.

Appointed Chairman of the Milwaukee Section's Educational Committee in 1947–1948, Mr. Van Zeeland helped sponsor a successful Transmission Symposium. The following year, as Section Chairman, he organized an equally well-attended Audio Discussion Group.

The Development of Physical Facilities for Research*

R. B. DITTMAR†

Summary—The problems of designing a modern research laboratory approach the design complexities of a large industrial plant, and the satisfactory solution requires architectural and engineering talent and, in most cases, actual scientific participation. Over-all problems must be considered, programs of research must be analyzed, and facilities must be developed to meet future requirements. This paper describes how these problems were met by the Bureau of Ordnance in establishing the Naval Ordnance Laboratory at Silver Spring, Md.

Background

FOR MANY YEARS the research laboratory in both industry and educational institutions was a stepchild. Many hours of design time and money went into the construction of the pilot and production plants and the institutions of higher education. The research laboratory usually occupied a dilapidated warehouse or stable in the plant, or a dark, damp basement in the university.

Early in the 1930's, with industries recognizing the value of professional personnel and the application of scientific research to improvement of product, a new era in laboratory construction came into being.

The Mellon interests in the period 1930 to 1937 built the new Mellon Institute, which is devoted to basic researches in the fields of chemistry and chemical engineering. The Bell Telephone Laboratories, outgrowing their West Street facilities in New York early in the 1940's, constructed a new and modern plant at Murray Hill, N. J. The Radio Corporation of America followed the same course at Princeton. Firestone, Goodyear, Standard Oil, and many more were close behind.

The fever spread to the colleges and universities; Northwestern Technological Institute was finished during the war, and many others followed. The Army, the Navy, and other government departments contributed their share: Aberdeen, Wright Field, Naval Research Laboratory, David Taylor Model Basin, Naval Ordnance Laboratory, the Regional Research Laboratories of the Department of Agriculture—all were rebuilt and expanded.

Decentralization of Laboratory from Plant

The thinking behind all the construction pointed in one direction: flexibility of interior design, involving the use of mobile metal partitions, availability of services in all areas, unit construction and interchangeability of furniture, etc. But perhaps the one outstanding departure from the pattern of the past was the trend toward decentralization of laboratory from plant. Thus, instead of occupying a small corner of the plant yard or the college campus, the research laboratory, of necessity, becomes a full-grown independent activity with all the associated problems that face a small community. Streets and roads must be planned. Fire protection, fire stations, domestic water systems, warehouses, restaurants and cafeterias, technical shops, public works shops, and even street lights and traffic signals must be considered and plans developed. The problem of designing a modern research laboratory now approaches the design complexities of a large industrial plant, with the additional problem of designing into the buildings a flexibility that will provide adequate services for a search in the future in unknown fields.

Establishment of Construction Schedule and Program

From June, 1944, until January, 1948, the writer was concerned with the planning and construction of the Naval Ordnance Laboratory at White Oak, Silver Spring, Maryland, a suburb of Washington, D. C. (Fig. 1). A brief summary of the problems encountered by the laboratory during this period will probably best illustrate how physical facilities for research are developed.

The Naval Ordnance Laboratory is one of the principal research agencies of the Bureau of Ordnance and exists to develop new weapons for the Navy. Fuzes, mines, depth charges, torpedo exploders, guided missiles, high explosives, the degaussing of ships—all are part of the laboratory program.

The Navy and the Bureau of Ordnance realized early in the war that production and research, while essential to each other, could not occupy the same structure and operate harmoniously. The regimentation of industry and the restrictions of necessity imposed on the workers of a large plant such as the Naval Gun Factory in Washington, where the Naval Ordnance Laboratory had grown up, did not and would not allow the scientists and engineers the freedom required to produce the devices so urgently needed by the Navy. The exigent demands from production for more space, and the urgent needs of the laboratory for expansion, caused continual pressure on management which could not be satisfied without the establishment of new facilities. Thus, the Laboratory was instructed to look around and find a suitable location where facilities similar to those it occupied in the Gun Factory could be constructed.

Management's first look at the problem indicated that about twenty-five buildings, two hundred and fifty acres of ground, and about $5,000,000 would be required to start the job. The Officer in Charge, Captain W. G. Schindler, then appointed a Building Committee of three from the staff of the Laboratory to assemble the requirements of the Laboratory personnel, and to establish the construction schedules and program.

The requirements were assembled and were checked against existing facilities, and the water in the program was wrung out. Then, in co-operation with the Bureau of Yards and Docks, and Eggers and Higgins, New York architects and engineers, the requirements were turned into so much floor space and so many buildings.

The Laboratory, operating as part of the Naval Gun Factory, rode along with the larger operation of the Gun Factory and obtained many services and utilities with no apparent expense or worry to the management of the Laboratory. Warehouse space was furnished, light, heat, water, sewer, gas, roads, telephones, fire protection, etc.—all were part of the services provided by the large Gun Factory. The planning for the new location had to include these items, which naturally expanded the size and cost of the first estimate.

Housing

An important factor in planning is the determination of the future population of the facility. It was estimated that in 1946–1947 the population of the Laboratory would be 2,000, and from this estimate the size of the utility systems, boiler plant, roads, warehouses, sewage plant, fire department, and shops, both technical and maintenance, were determined.

When the study was complete and a tentative plot plan had been developed, the original estimate of the Laboratory was found to be considerably in error. One thousand acres of ground and a total of ninety permanent and temporary structures

Fig. 1—Naval Ordnance Laboratory, White Oak, Silver Spring, Md., showing the Administration and Laboratory Building, looking northeast from the main gate.

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† University of California, Los Alamos Scientific Laboratory, Los Alamos, N. M.

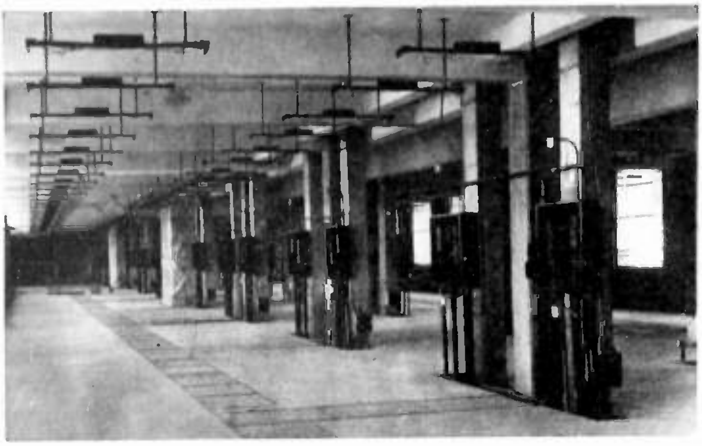

Fig. 2—Typical interior view of the Administration and Laboratory Building, prior to the installation of metal partition units.

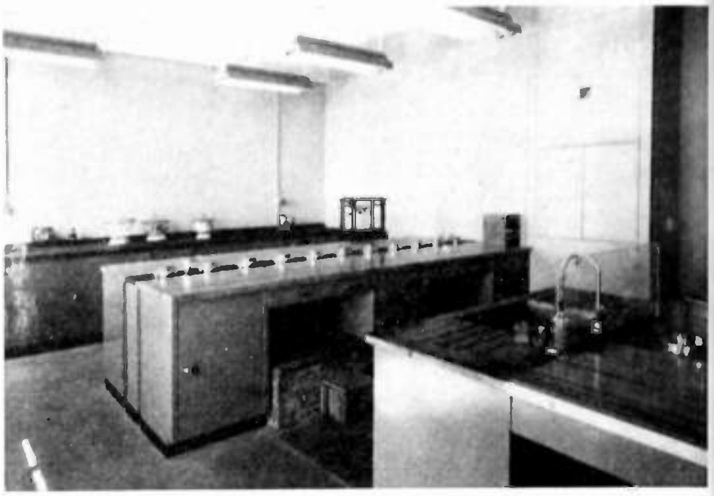

Fig. 4—Typical laboratory room in the Explosives Laboratory.

were required, as compared with the initial estimate of two hundred and fifty acres and twenty-five buildings. It was further determined that the construction cost would be nearer $15,000,000 than the $5,000,000 originally estimated.

These figures were submitted to the Navy and the "go-ahead" was obtained, but before the major items were built the program had almost doubled to a new total figure of about $26,000,000. Fortunately, the original space and utility planning was sound, and no additional land or basic utilities were needed. After approval of the plan by the Navy, the problem of locating the best site for the laboratory became a reality.

It must be remembered that a laboratory of approximately 2,000 people does not mean 2,000 scientists. Actually, the ratio of scientists to service personnel is about one scientist for every two service employees. By definition, a service employee ranges from secretary and clerk to shop machinist, instrument maker, carpenter, janitor, etc. As is true in all government operations, with the exception of top supervision, salaries are limited to $10,000 per year. It can, therefore, be understood that the average income of the group is not nearly the income of the top personnel; consequently, the site must be so selected that housing available to all income groups must be within reasonable distance, and that public transportation must be provided.

Location

The desire to decentralize the laboratory from an industrial area indicated a residential area as the most desirable location, but the fact that 1,000 acres of ground was required complicated that desire. After many tours through the suburban area of Washington, and with the help of a United States Geodetic Service topographic map, an area just on the fringe of the residential zone of Silver Spring, Md., located on a well-developed highway—New Hampshire Avenue extended—was selected. While public transportation was not then available, it was within a short distance and, with the coming of the Laboratory, could be extended to the grounds with very little difficulty.

The housing problem was considered as important as the construction of the new laboratory, and a housing committee was organized and an office with paid employees was set up and given the problem of stimulating interest in new housing among the local realtors and builders.

Development of Design and Construction

With all the preliminary requirements thought out and under way, the actual construction job was tackled. First, the ground was graded and drainage was installed. One by one the roads were developed, all following the pattern of the master plan. Then came the utility systems, including water lines, sewers, gas lines, underground electrical service, telephone ducts and cables and fire-alarm circuits. Simultaneously with this construction, the boiler plant with the underground steam tunnels was built. The warehouses for the storage of materials of construction and the technical and maintenance shops followed soon after.

During this period of construction, plans for the laboratory and administration buildings were shaping up. With scientific developments advancing as they were in 1944–1945, it was becoming difficult to keep the construction of facilities on schedule, and it was soon decided that no fixed facility designed in 1945 would satisfy the needs of the scientist in 1946. This conclusion led to the adoption of modular-type construction for the laboratory and office buildings, and the use of mobile metal partitions throughout these areas.

Study indicated that the minimum amount of office space could be established and that it could be treated separately from laboratory requirements; however, the layout of the main building was so arranged that, should expansion become necessary

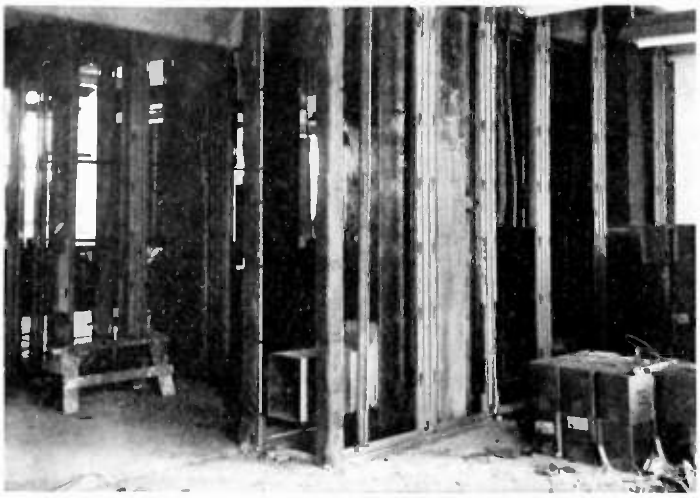

Fig. 3—Typical interior view on the first floor of the Explosives Laboratory.

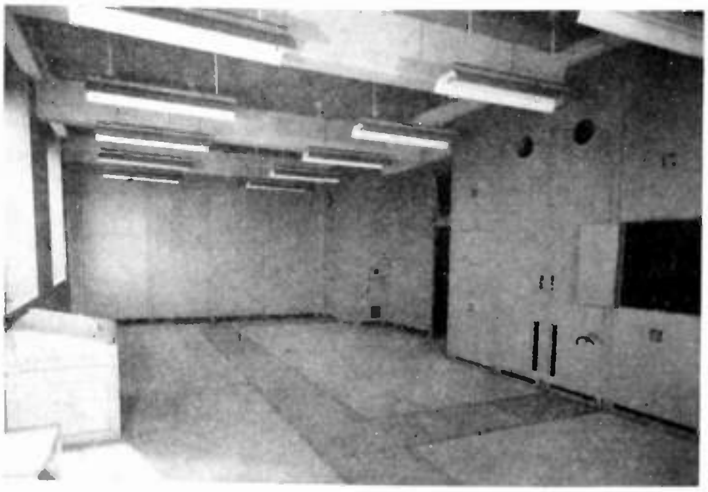

Fig. 5—Interior view showing a typical first-floor laboratory room in the Explosives Laboratory.

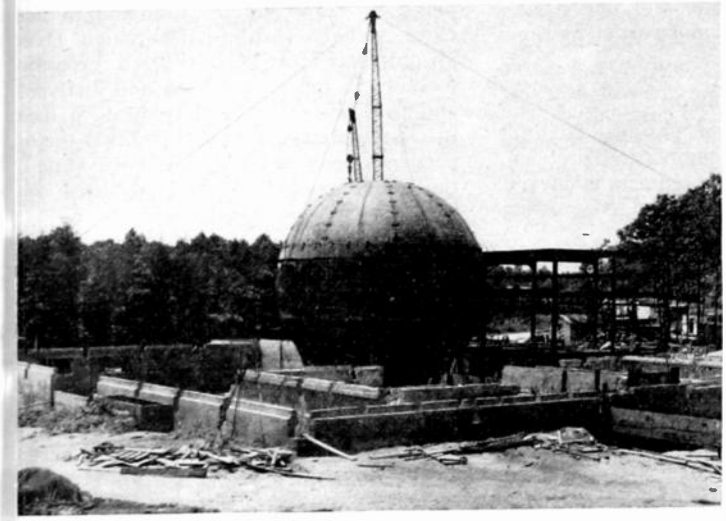

Fig. 6—The Supersonic Wind Tunnels Building, looking southwest.

Fig. 8—The Explosives Research Laboratory, one of several located in an isolated area of the Naval Research Laboratory grounds at White Oak, Md., near Silver Spring.

additional laboratory space could be obtained through new wing construction, and the office area could be expanded into laboratory area without the hindrance of physical separation. As no utilities were needed in the basic office area, considerable saving in over-all construction cost resulted.

The main laboratory structure, as designed, consisted of a building of two and three stories, structural steel, reinforced concrete and brick construction, and had a gross floor area of about 400,000 square feet and a net usable area of 275,000 square feet. A cafeteria seating 600 people, and directly over the cafeteria an auditorium seating 550 people, was connected to the structure.

As mentioned before, the laboratory structure is built on a modular basis and the separating partitions as well as the corridor and exterior wall partitions are of metal (Fig. 2). The module selected was 11 feet, and the laboratories were so designed that all services—i.e. 120/208 volt four-wire single- and three-phase power, compressed air, gas, hot and cold water, steam, hood exhaust, and laboratory acid waste—are brought through service shafts located on the corridor walls at 22-foot centers. (Fig. 3) All services, with the exception of the hood exhaust, are run on the basement ceiling directly under the service shafts. The hood exhaust rises to an attic, and each stack has an exhaust fan discharging the fumes to the atmosphere.

The entire laboratory structures, cafeteria, auditorium and two adjacent buildings, the mechanical test laboratory, and the explosives laboratory are provided with summer-winter air conditioning (Figs. 4 and 5). The air-conditioning control is laid out so that each room has its own individual thermostat, and the occupant can control the temperature in the room through a range of 68° to 80°, both summer and winter. Humidity control of 50 per cent relative humidity in summer and 30 per cent relative humidity in winter is provided at every central fan system. Air is introduced under the window at each 11-foot module through a Carrier Weathermaster unit.

Furniture Requirements

The adoption of this type of construction allowed the rapid completion of plans and specifications and an early starting date. As construction progressed, individual space requirements were determined and standard steel laboratory furniture units were developed. In order to speed up the furniture procurement, the steel units were designed as interchangeable units, and a variety of types was selected prior to actually determining individual needs. A catalog showing the available furniture, plus a blank space diagram, was given each division of the Laboratory, and a selection of furniture by catalog number was made by each individual occupant.

As construction of the plant neared completion, a review of space and furniture requirements was made, as well as space changes due to revision in organization or function of the divisions, and the occupancy drawings and ultimately the partition layouts were revised. The use of a fixed wall construction would have made this type of planning an impossibility.

Additional Structures

In addition to the main laboratory structures constructed at White Oak, several other groups of laboratories were constructed concurrently. A group of nine magnetic laboratory buildings used for research and development of magnetic mines were built. All but one of these buildings were constructed of nonmagnetic materials, even to brass rods for reinforcing and concrete block instead of brick, as red brick contains iron oxide (Fig. 7) The ninth building, located in the center of the group, was designed as a service building containing a boiler plant, dc sources for the magnetic coil systems in the other buildings, an experimental shop, and office facilities for the area.

A group of several buildings known as the Explosives Area, considerably isolated from the main group of buildings, was constructed. These contained special facilities for the study and test of high explosives. (Fig. 8)

Perhaps the most interesting group of buildings, other than the main laboratory structures, was the Supersonic Wind Tunnel Group (Figs. 6 and 9). The high-speed supersonic tunnels used by the Germans in

Fig. 7—View looking east, showing the Spherical Field Laboratory, Building 203, and Long Field Laboratory, Building 204.

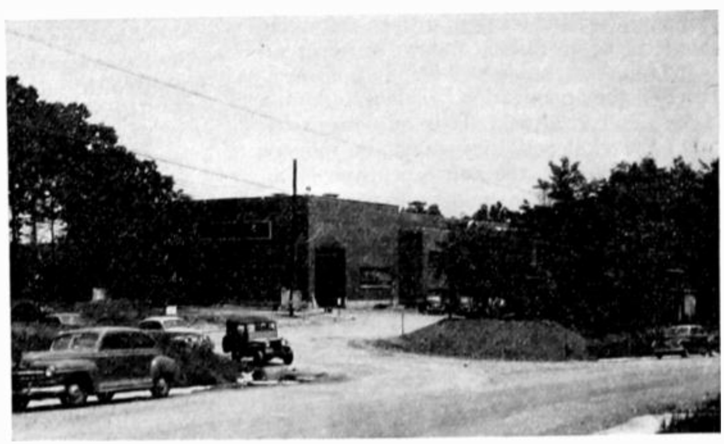

Fig. 9—Supersonic Wind Tunnels Building, view looking northeast.

the development of the V-1 and V-2 missiles at Kochel, Bavaria, were brought to this country by the Navy, completely rebuilt, and installed in new structures at White Oak. In addition to the tunnel structures, a Ballistics Range building was erected. This building contains a free-flight and pressurized range used in the checking of data obtained in the wind tunnels. The supersonic tunnels, when complete, will develop air speeds in the range of seven to eight times the speed of sound.

Conclusion

The development of physical facilities for research becomes a complicated problem. The satisfactory solution requires talent not only in the architectural and engineering fields, but also scientific liaison and, in most cases, actual scientific participation. Overall problems must be considered, programs of research must be known and analyzed, and facilities must be developed to meet future requirements. It is hoped that the experiences described will help solve some of the more pressing facilities problems confronting scientific management.

Personnel Administration in Research and Development Organizations*

C. E. BARTHEL, JR.†

Summary—**After accepting a definition for personnel administration, it is emphasized that the responsibility for a personnel program rests upon all executives and supervisors of an organization, and that an effective program requires the co-operation and understanding of all members of the organization. Typical personnel activities are listed.**

Definitions for basic research, applied research, and development are given, and it is indicated that all three types of effort require persons who are extremely well trained in specific fields of endeavor, persons with creative ability, imagination, and originality of a high order. Conditions conducive to the productivity of such persons are outlined. It is then stressed that general personnel activities are applicable to research and development organizations, and, in addition, specialized functions are required to realize the conditions conducive to the happiness and productivity of scientists and engineers. However, research scientists and development engineers must actively participate in the personnel functions of a research and development group.

The methods of discharging personnel responsibilities employed by two research and development organizations are contrasted. It is emphasized, however, that, with any method, effective personnel administration must be a service, not a control.

I. Introduction

Before proceeding with a discussion of personnel administration in research and development organizations, it should be helpful to agree upon a general definition of personnel administration and to consider typical personnel activities in other types of organizations. Tead and Metcalf[1] define the term "personnel administration" as "the planning, supervision, direction, and co-ordination of those activities of an organization which contribute to realizing the defined purposes of that organization with a minimum of human effort and friction, with an animating spirit of co-operation, and with proper regard for the genuine well-being of all members of the organization." This is, indeed, a broad and vital field. It is quite apparent that the vital activities cannot be the sole responsibility of the top executives of an organization or of a small group which bears the name of "personnel department." Lesser executives and persons in all supervisory positions carry the basic responsibility, since it is generally agreed that the most successful personnel program is one in which personnel problems can be completely handled at the very source of the problems. Then, too, an effective personnel program requires some degree of co-operation and understanding on the part of the individual worker. Thus, an effective personnel program is a co-operative one requiring the understanding and some effort on the part of every member of the organization.

It is widely known that the employer has not always considered "the genuine well-being of all members of the organization." At the present time, however, progressive employers agree that the happy and satisfied worker is the co-operative and productive worker. To maintain a happy and contented group of employees, these employers have, in large organizations, now employed specialists to handle the many personnel functions of their organizations. Days could be spent on a detailed consideration of typical personnel activities, and the time allotted for this paper could be completely consumed by a mere listing of such activities. Listings are given in many books on personnel management, and, to channelize our thinking, the items given by Scott, Clothier, Mathewson, and Spriegel[2] as the first broad breakdown of personnel activities are listed:

1. Employment
2. Promotions, transfers, discharges, and separations
3. Formulation and direction of training program in helping with company objectives
4. Remuneration and incentives
5. Health and sanitation
6. Safety
7. Financial aids to employees
8. Employee service activities
9. Employee-employer and community co-operation.

Further consideration of these activities is beyond the scope of this paper.

II. Personnel Functions of Research and Development Organizations

After briefly listing typical personnel functions for a general type of organization, we should now direct our thoughts specifically to research and development, to research scientists and development engineers. Then we should consider a few conditions which are conducive to the productivity of scientists and engineers, and, finally, we should discuss factors in a personnel program which are essential for a realization of those conditions conducive to productivity.

It is perhaps apropos to quote the definitions of the Scientific Research Board[3] for basic research, applied research, and development:

"1. *Basic Research*

"a. *Fundamental Research*
Fundamental research is theoretical analysis, exploration, or experimentation directed to the extension of knowledge of the general principles governing natural or social phenomena.

"b. *Background Research*
Background research is the systematic observation, collection, organization, and presentation of facts using known principles to reach objectives that are clearly defined before the research is undertaken to provide a foundation for subsequent research or to provide standard reference data.

"2. *Applied Research*
Applied research is the extension of basic research to the determination of generally accepted principles with a view to specific application, generally involving the devising of specified novel product, process, technique, or device.

"3. *Development*
Development is the adaptation of research findings to experimental, demonstration, or clinical purposes, including the experimental production and testing of models, devices, equipment, materials, procedures, and processes. Development research is related to work on an existing model, device, equipment, material,

* Decimal classification: R005×R072. Original manuscript received by the Institute, November 12, 1948. Presented, National Electronics Conference, Chicago, Ill., November 5, 1948.

† Armour Research Foundation, Illinois Institute of Technology, Chicago, Ill.

[1] O. Tead and H. C. Metcalf, "Personal Administration: Its Principles and Practice," McGraw-Hill Book Company, New York, N. Y., 1933; third edition, p. 2.

[2] W. D. Scott, R. C. Clothier, S. B. Mathewson, and W. R. Spriegel, "Personnel Management: Principles, Practices, and Point of View," McGraw-Hill Book Company, New York, N. Y., 1941; third edition, pp. 30–32.

[3] The President's Scientific Research Board, "Administration for Research," Volume III of *Science and Public Policy*, p. 6, 1947.

or product process. Developmental research differs from applied research in that the work is done on products, processes, techniques, or devices previously discovered or invented."[3]

These definitions were obviously devised to differentiate between the three types of investigations. I have repeated them here to emphasize that all three types of work require persons who are extremely well trained in specific fields of endeavor, persons with creative ability, imagination, and originality of a high order, or, as Bichowsky[4] has indicated, persons with a large "number of square inches of frontal lobes . . . the place in a man's brain where new behavior patterns are established." Because of their personal make-up and the general environment in which they work, the research scientist and development engineer tend to be individualists. As individualists, these persons cannot tolerate regimentation. Therefore, a happy, contented, and productive research scientist or development engineer is one who is treated in his organization as an individual, as an integral part of the organization. Such a person *must* feel that in his particular position he is able to develop personally and professionally. He must have freedom of action within the organization, and must be provided maximum opportunity for thinking. In addition, he must be free from worry, and, finally, he must be provided with the necessary tools of his profession and with administrative support adequate for clear-cut, rapid, and business-like decisions on the results of his work. These are a few, but a very important few, of the conditions conducive to the productivity of scientists and engineers.

With these general thoughts on research and development and on research and development personnel, let us now turn our attention to "those activities of an organization which contribute to . . . the genuine well-being of all members of the organization." Certainly those personnel activities which we listed above as typical for a general organization are equally applicable to the research and development institution. However, different types of specialists are required for handling some of these activities. For example, who but an organic chemist could confer intelligently with other organic chemists in the development of a potential supply of organic chemists for a specific organic chemistry program? Who but an electronic engineer could correlate the requirements of an electronic engineering position with the particular qualifications of electronic engineers applying for that position? Who but an optical engineer could evaluate the performance of a person developing new types of lens design? Who but a metallurgist could devise training programs for metallurgists and establish and maintain liaison and effective co-operation with metallurgical departments of colleges and universities for the teaching of these training courses? *It is my firm belief that research scientists and development engineers must actively participate in the normal personnel functions of research and development organizations.*

[4] F. Russell Bichowsky, "Industrial Research," Brooklyn Chemical Publishing Company, Brooklyn, N. Y., 1942; p. 95.

But the normal personnel functions are not adequate for a research and development organization. Management must provide ample opportunity for the professional and personal development of the scientists and engineers: seminars must be held at frequent intervals to permit an extensive exchange of information; funds must be provided for sending personnel to meetings of their respective professional societies so that they might meet their colleagues and keep themselves informed on the progress of their profession; in-service training courses must be established to provide an opportunity for inexperienced personnel to familiarize themselves with the thinking and techniques of the organization's specialists; a college training program must be supported to encourage personnel to receive advanced basic training. The exchange of information and training programs assure an alert, informed, progressive professional group capable of applying modern techniques to problems and of meeting emergencies as they arise. Effective organizational structures must be established and maintained to permit the individual worker freedom of action, to permit immediate procurement of tools and materials if and when he needs them, and to permit clean-cut, rapid decisions on scientific and technical matters. Research and development today assure progress tomorrow, and delays in the procurement of materials and in the rendering of decisions only defer progress. Work conditions contributive to thinking must be provided. A quiet, clean, bright office is a modest investment, but such a facility can double or treble the output of a scientist or engineer who has been working under improper conditions. Very important in these days of confusion, provision must be made for keeping the personal worries of the individual to a very minimum. The output of a scientist or engineer comes from Bichowsky's frontal lobes; if these frontal lobes are occupied with personal worries, it is apparent that they cannot be utilized fully on productive problems. Help could be provided, for example, in the procurement of sound information on housing and on minor legal matters, and of travel reservations and tickets for theater and sporting events, and by the establishment and maintenance of post-office and banking services.

Some persons believe that services of this type represent a glaring discrimination, but during the past few years more and more organizations have been establishing them as necessary for the realization of their purposes. These organizations are actively concerned with the worries of their research personnel. For example, during the Illinois Conference on Industrial Research, held on May 27–28, 1948, Albert L. Elder, Director of Research of the Corn Products Refining Company, said: "This statement has been made to me regarding the research men: 'A research man enjoys a big worry, but a little worry will worry him to death.' So the thing that you want to do is keep him free from the little things that irritate him and then give him one big thing to worry about. Then he is perfectly happy. That's one of the things, at least, that I try to do in my laboratory; that is, keep the little irritations away from them."[5]

Very recently, Leonard B. Loeb, of the University of California, expressed much concern about the decline of scientific proficiency in industrial and governmental research laboratories. He believes that "a broadening element is completely lacking in the present organization of research laboratories."[6] The above-mentioned conditions conducive to the productivity of scientists and engineers include to some extent the broadening element, and an effective personnel program developed around these conditions could assist in the elimination of the stagnation and deterioration of scientific proficiency which so concerns Dr. Loeb.

III. Organizational Procedures for Discharging Personnel Responsibilities

It probably will be noted that, up to this point, we have considered only personnel functions. No positive thought has been given to anchoring the associated responsibilities and, above all, no positive mention has been made of a personnel department. The writer is in complete agreement with Scott, Clothier, Mathewson, and Spriegel[7] in their statements: "The personnel work of an organization cannot be housed within a certain department bearing that exalted name. Personnel management is a leaven permeating all phases of management; the responsibility for it rests upon all executives and persons in supervisory positions. Personnel policies may be defined by the major officers of the organization, but the lesser executives, the foremen, and the supervisors are the ones who carry out those policies—they are the real personnel managers." These "real personnel managers" in a research and development organization are research scientists and development engineers. They *must* participate in the formulation of personnel policies and provide advice and assistance, when necessary, in the administration of a personnel program. But the activity of these persons does not suffice for the development and maintenance of an effective personnel program. It is my belief that the individual scientist and engineer must himself contribute to such a program, must assist in the development of *his* personnel program.

It would be extremely desirable to describe a definite personnel program for a normal research and development organization, and to define an organization for handling details of such a program. Unfortunately, such a description and definition are quite impossible. A formula for expressing the personnel functions necessary for building up and maintaining a happy and contented group of scientists and engineers has many variables, about as many as the number of persons employed. In addition, such constants as the type of work to be accomplished, geographical location, community activities, etc., vary from organization to organization. As indicated above, it is a healthy plan to permit the scientists and engineers themselves to assist in developing the formula and applying it to their problems. The na-

[5] Armour Research Foundation of Illinois Institute of Technology and Illinois Manufacturers' Association, "Proceedings of the Illinois Conference on Industrial Research"; p. 57.

[6] Leonard B. Loeb, "The maintenance of scientific proficiency in nonacademic research laboratories," *Science*, vol. 108, pp. 267–272; September 10, 1948.

[7] See page 26 of footnote reference 2.

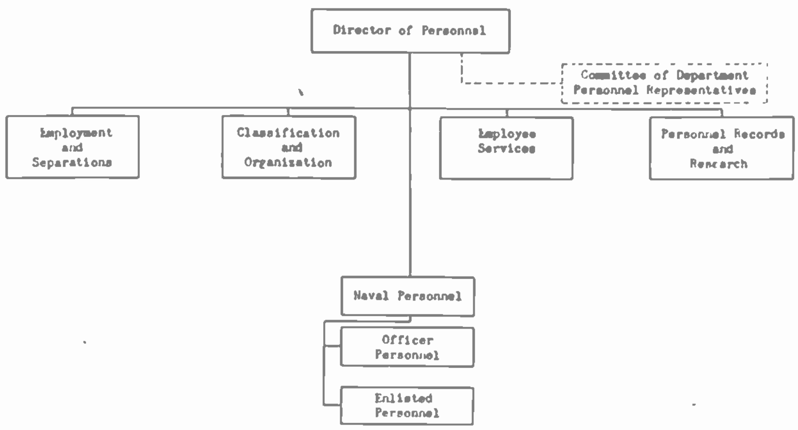

Fig. 1—Personnel Department of the Naval Ordnance Laboratory.

ture of their participation will also vary among organizations. It must be realized that the prime responsibility of research personnel is to conduct research. Their personnel responsibilities should, therefore, be of an advisory type, and in large organizations the details must be administered by a group of administrative and personnel specialists. It is essential, though, that in any organization the scientists and engineers participate in the development of their personnel policies, and assist in the co-ordination of these policies with the policies relating to other groups of the organization.

The size of an organization is no criterion of the need for a personnel department or for the number of persons necessary for handling the routine of a personnel program. Obviously, the number of persons required and the positions of those persons in an organizational structure are completely dependent upon the personnel functions considered necessary for the organization.

It may be appropriate to discuss briefly the methods of discharging personnel responsibilities employed by two research and development organizations with which the writer has been associated. These organizations are the Naval Ordnance Laboratory, and the Armour Research Foundation.

The Naval Ordnance Laboratory had mushroomed from a small group of twenty-odd people in 1939 to an organization of about 2,000 people in early 1944. At the latter time, the Congress appropriated funds for the establishment of the Laboratory as a permanent institution, one of the Navy's largest research and development organizations. Prior to this time, in the early days of its rapid expansion, the Laboratory leaders organized and successfully implemented a vigorous recruiting campaign. By 1942, the employment function had become a routine one and, since the Laboratory was considered a "war agency," very little attention was given to this or to any other personnel activities. In mid-1944, however, it became apparent to the officer-in-charge and his staff that, if the Laboratory was to maintain itself as a vigorous organization in the peacetime years, considerable thought and effort must be given to the development of personnel policies and procedures, to the task of finding good persons for the organization and *keeping* them.

Now the Naval Ordnance Laboratory was operating with four distinct types of personnel: naval officer personnel, naval enlisted personnel, Civil Service employees, and personal-service-contract employees. The procedures for employing, transferring, and promoting the persons in each of these four categories, and for handling the majority of the personnel functions associated with their activities, were different. Since these procedures were established basically by congressional action to assure free competition for positions and to protect public monies, they were quite complex. Ralph D. Bennett, technical director of the Laboratory, has commented on the complexity of this problem.[8] The development of an organization for maintaining a happy and contented group was, therefore, difficult, and of necessity the organization developed was somewhat complex. A chart of this organization is given as Fig. 1. The Naval Personnel Division handled all functions for naval personnel, while the remaining four divisions handled the details of the personnel activities of the Civil Service and contract employees. There is nothing unusual about the divisional breakdown indicated on the chart. However, the Committee of Departmental Personnel Representatives is an interesting and, perhaps, unusual group. The committee consisted of the director of personnel, as chairman, and one representative from each of the other operating departments of the Laboratory. The departments were the Research Department, the Engineering Department, the Technical Evaluation Department, charged with scientific missions, and the Technical Services Department and General Services Department, charged with service or administrative missions. The departmental personnel representatives for the three technical departments were scientists or engineers, as, incidentally, was the director of personnel. Each representative was responsible for maintaining close liaison between his department and the Personnel Department for personnel functions, and the group as a whole was responsible for recommending personnel policies and procedures for the over-all organization, thus co-ordinating the policies for the different groups of the Laboratory. Once these policies and procedures were approved, they were administered by the director of personnel as service functions. About thirty-five persons operated in the Personnel Department.

The success of the personnel program of the Naval Ordnance Laboratory is evidenced by the fact that, at the termination of the war, a high percentage of persons remained with the Laboratory. They remained in spite of the fact that they came to the Laboratory as "temporary" employees on leave of absence from former employers.

The personnel methods of the Armour Research Foundation are considerably different. This organization consists of some 650 persons, over half of whom are scientists and engineers. The writer has been with the organization a relatively short time, and is most impressed by the spirit of the personnel. The group could be nothing but happy and contented, and the enthusiasm for their work is indeed of a high order. In this relatively short time with the Foundation, the writer has been convinced that it has a very sound personnel program. Yet it possesses not a single person whose sole responsibility is for personnel matters. Personnel functions are handled within the operating departments, with the necessary co-ordination being handled by the regular administrative staff.

The methods employed by both institutions are, in this writer's opinion, quite sound, because they were designed to (1) care for the necessary personnel functions of the organizations, (2) permit the scientists and engineers to participate in the development of their personnel programs, (3) permit solutions of personnel problems at their very sources, and (4) provide a wholesome and healthy service for the administration and for the individual staff members.

After observing the personnel methods of many other organizations, the writer is convinced that, to be effective, *personnel administration must be a service, not a control.*

IV. Conclusion

In conclusion, the writer would like to re-emphasize that personnel administration of research and development organizations is exceedingly complex, and, instead of neglecting this vital activity as do some of our prominent institutions, considerably more attention should be applied to the personnel matters of such organizations than for organizations of a production type. Research scientists and development engineers must participate in the formulation of personnel policies, in personnel operation, and in the conduct of personnel research and planning. Only these persons are in a position to understand sufficiently the make-up of the human research and development resources. Finally, it is a prime responsibility of management to develop these resources, since our Nation's progress and future depend so much on them. As Marquis Childs pointed out in his column, "Washington Calling," in the Washington *Post* of September 16, 1948: "After all, only scientists can conduct scientific research."

[8] Ralph D. Bennett, "Engineers and scientists in government service," *Elec. Eng.*, vol. 64, pp. 383–386; November, 1945.

Information Exchange as a Management Tool in a Large Research Organization*

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Summary—The foundation of teamwork in a large research organization is the effective exchange and dissemination of information regarding technical and administrative matters. Present-day researchers and managers are faced with a tremendous task in making effective use of the information-exchange methods that are available. This paper discusses information exchange and dissemination from the viewpoint of management. Examples of successful information exchange as a management tool are given, as well as examples where inefficient information-exchange methods have been a hindrance to the over-all research program.

I. Introduction

THE FRAMEWORK of a large research organization is formed by the welding together of individual researchers into an organized research team. The very nature of scientific work requires a high degree of specialization by individual workers, thereby making it particularly important for management to insure that correct information exchange is accomplished. In a sense, the management of a research activity may be considered as a communication unit whose purpose is to establish and maintain a convenient and readily understandable system whereby individuals may best accomplish the over-all purpose of the organization. There are other aspects to management besides that of communications, but if, as Schade[1] has said, the purpose of the management of a research activity is to "... provide the means by which the efforts and creative talents of a large group of individuals are harmonized and made more effective ...," certainly the conscious and efficient exchange of information is one tool, and a very important tool, whereby this purpose can be achieved. Like most tools, however, it must be used skillfully to be effective.

II. Personal Contacts and Indirect Contacts

There are two basic types of contacts that can be used for information exchange. These are "personal" contacts and "indirect" contacts. Personal contacts include face-to-face meetings of two or more people where an easy flow of information can be maintained in both directions among the participants. Essentially instantaneous contact through a two-way communication device such as a telephone is considered to be a personal contact. Indirect contacts are characterized by an essentially one-way flow of information. All written reports, memoranda, instructions, etc., are indirect contacts. A lecture where information is principally transmitted from the speaker to an audience is considered to be an indirect contact.

Good research management places more emphasis upon direct personal contacts than it does upon indirect contacts. Personal contacts make possible an easy interchange of information beneficial to both parties and, in addition, can have very important morale advantages. It so happens that managers, executives, research directors, or whatever they are called have many things to do and a limited time in which to do them. The amount of time that they can spend in personal contacts is limited. Personal contacts must, therefore, be delegated[2] and supplemented by indirect contacts.

III. Personal Interest by Management as an Incentive

The practice by management of discussing problems with individuals and small groups before final policy decisions are reached is a positive moral factor and production incentive either in the laboratory or in the factory. Certainly, a memorandum affecting the research program of a laboratory is more carefully thought out and is accepted and followed much more willingly if it is the result of co-operative effort on the part of management and the individuals affected.

It is difficult to measure how much the use by management of direct contacts with subordinates increases production in a research organization, because the yardstick for measuring research production is not yet well defined. Instead, I would like to cite the well-known example given by Chase,[3] and taken from Roethlisberger and Dickson[4] showing how sustained interest by management affected the production of a group of six experienced workers assembling telephone relays. These workers were asked to participate as a team in helping the company solve the problem of determining the kinds of incentives that would result in most efficient production. After consenting, they were placed in a room away from other workers doing similar work. During about a two-year interval these workers, together with the management representative conducting the experiments, were first subjected to standard conditions of work successively modified by morning and afternoon rest periods, free hot snacks, a shorter working day, and a shorter working week. All of these successively improved working conditions appeared to increase production, or at least not to have a detrimental effect. Finally, the original standard condition of work was tried, with no rest periods, no snacks, a full working day, and a full working week. Production then skyrocketed to an all-time high.

The management staff conducting the study came to the conclusion that the sustained personal interest that they had shown in working conditions had changed the workers' attitude. They no longer considered themselves as cogs in a machine, but rather felt that they were a congenial team helping management to solve significant problems. The change in attitude caused by the sustained personal contacts and exchange of information with the management representative was a factor certainly as important as good physical working conditions.

Research-laboratory management can be sure that the basic psychological makeup of research workers is, in many ways, similar to that of persons making relays. In fact, nonquantitative experience has shown that the amount of research effort is increased by the simple act of maintaining sustained sincere interest in the work of the small research teams and the individual workers.

IV. Scheduled Personal Contacts in the Large Laboratory

In a large research organization, scheduled personal contacts between the research director and his division superintendents, and among the division superintendents themselves, form an information-exchange tool that is widely used to attack technical and administrative problems. Such a group, called by a suitable name such as the "Laboratory Scientific Program Board," meets at regular scheduled intervals. In a large laboratory, a two-week interval is a good compromise between meeting too often and not meeting often enough. Usually, an agenda, prepared by a secretary, is transmitted to the board members a day or so before the scheduled meeting, to serve as a reminder and to initiate discussion on the various problems at hand. In my experience, I have seen the Naval Research Laboratory operated both with and without a scientific program board. This tool has proved to have considerable value in the planning and execution of a laboratory-wide scientific program, as compared with the natural tendency of the separate divisions to work on separate divisional programs. Of course, the use of a scientific program board does not cure all ills, but it helps.

It has also been found that similar boards within the smaller units of the laboratory help in promoting teamwork. Such a group, which may be called a "Division Technical Planning Board," consists of the division superintendent as chairman and the section heads as members. This board considers the technical and administrative problems in the special technical field of the di-

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[1] H. A. Schade, "Inception and development of a research project," Proceedings of the Conference on administration of Research, Pennsylvania State College, School of Engineering, Technical Bulletin No. 29.

[2] V. A. Graicunas, "Relationships in Organization," Papers on the Science of Administration, Institute of Public Administration, New York, N. Y., 1937; pp. 183–187.

[3] Stuart Chase, "Men at Work," Harcourt, Brace and Company, New York, N. Y., 1945; pp. 9–27.

[4] F. J. Roethlisberger and W. J. Dickson, "Management and the Worker," Harvard University Press, Cambridge, Mass., 1946.

vision in much the same way as the scientific program board handles similar problems for the laboratory. The question arises as to whether a technical planning board should discuss administrative matters. It has been found best to discuss such matters, because, in many cases, the planning and carrying out of the technical program is directly affected by administrative problems and policies. Such discussions also serve to broaden the administrative outlook of the section heads. There is a tendency, however, for the unguided discussion of administrative matters to consume an unduly large proportion of the time. Thus, it is up to the chairman to see that suitable administrative as well as technical matters are continually injected into the agenda. The calling in of subordinate specialists on the technical aspects of particular problems is important in stimulating the exchange of technical information. This helps in the planning of the technical program and, in turn, aids morale. In like manner, administration specialists may be called on to advise on perplexing administrative problems.

In addition to formal personal contact meetings such as those described above, informal meetings are certainly helpful as a tool for information exchange. Many large organizations provide separate lunchroom facilities for the convenience of the director, his division superintendents, and guests. Observation has indicated that a part of the exchange of information that takes place at these luncheon meetings would not have occurred during the usual course of operations. Such an arrangement has been found to be worth while. There is, of course, a good deal that a division superintendent can gain through informal contacts with his section heads. Hence, it is up to the superintendent to provide the occasions for informal contacts with his subordinates.

Luncheon contacts and program board meetings have been discussed as tools for information exchange and laboratory-wide co-ordination. There is another tool that has been used by management to disseminate technical information by personal contacts. This involves the scheduling of visits by the scientific program board members to the locations where the actual research is taking place. Discussion and demonstrations by subordinate scientists give management people a firsthand knowledge of what is being done in all parts of the laboratory. This procedure helps to stimulate realistic program planning on a laboratory-wide basis. Scheduling of such tours to selected parts of the laboratory on a monthly basis provides an efficient tool for accomplishing a type of information exchange that probably will not otherwise take place.

V. Co-ordination Between Laboratories by Personal Contacts

Often, in the prosecution of a large research and development project, it is found necessary to divide the work among several laboratories or other agencies. This is necessary when the job is too large or too involved for any one laboratory to handle. Usually, the division of work is made on the basis of the accomplishments and specialties of the various laboratories and agencies involved. It is necessary for one group to have prime cognizance and to be responsible for the co-ordination problem, in theory, at least. A co-ordinator may be selected, and usually a co-ordination board consisting of members of each of the participating laboratories and agencies should be formed. The success of the project will be affected to a large degree by the ability of the co-ordinator, and by his effective use of good information-exchange techniques. A research and development project too large and too involved for any one laboratory to handle will be very difficult to co-ordinate. This is true because laboratory organizations, like other organizations, do not like to be co-ordinated by any outsider. The co-ordinator may know the over-all picture quite well, but certainly he cannot know and understand of all the technical details in all of the fields involved. And it is the technical details that must be fitted together satisfactorily for the research and development project to be successful. Thus the exchange of the technical details between all of the participants must be accomplished before co-ordination becomes a fact. Usually, written technical reports are used as a means of information exchange, and they are most important. However, such reports are indirect contacts and are some months behind the actual work, due to the time consumed in writing, editing, duplicating, and distributing the reports.

A direct information-exchange tool that has been used with some success in co-ordinating a large involved project will be described. In this case, no co-ordinating board was formed and the co-ordinating agency proved to be a high impedance through which the flow of technical information had to filter. One of the laboratories participating in the project initiated a monthly technical seminar to report progress on the part of the project for which they were responsible. Also included in the agenda were a few specific technical reports by the working-level scientists. Advance copies of the agenda were sent to individuals in the various participating organizations, inviting the attendance of all interested persons. The seminars provided the occasion for technical, information exchange in informal groups before and after the meetings and by active discussion during the meetings. On succeeding meetings, various representatives of the participating organizations were invited to give technical reports on their work. Thus, in effect, the working-level engineers and scientists were able, by personal technical information exchange, to co-ordinate the program without the term co-ordination actually being used. Full reports of the seminars were written and distributed to all interested laboratories and agencies. The official co-ordinating agency, although slightly cool toward the technical seminars at first, became their chief supporter when the benefit to the program became evident.

VI. The Use of Some Indirect-Contact Media

1. General Announcements

Often, in the operation of a large laboratory, it is necessary to issue general administrative memoranda concerning personnel actions, statements of or changes in policy, etc. Quite often these memoranda come out in the form of arbitrary orders which neither explain nor provide the facts used in arriving at the decisions. It is admitted that detailed justification cannot be provided in all cases, but this does not alter the fact that an explanation is desirable. For example, at one time a general announcement was prepared calling all personnel of a division to meet at a certain time and place, to be addressed by the division superintendent. The exact subject of the talk was not given because the superintendent intended to talk about miscellaneous administrative and technical items. Thus, this terse announcement hit the members of the division with no reason given for the action. Speculation immediately started rumors that the division superintendent was going to resign or that the division was going to be merged with another division. Time wasted in speculation had an appreciable effect on the over-all work output until the division superintendent actually gave his talk. A few additional words on the memorandum would have avoided this confusion.

Another simple example involved a memorandum from the central administration urgently requiring a report on the location of all safes of a particular type in the divisions. Some divisions speculated that these safes were going to be removed for some unknown urgent use, and worried about how they would store the material so displaced. The speculation continued until one of the superintendents asked the assistant director of the laboratory at lunch time to explain the order. The simple reason was that the safes had been found defective and that a minor repair was needed to correct this defect. These are trivial examples, but they illustrate the importance of giving enough information on indirect-contact memoranda and announcements to minimize useless speculation on the reason for their issuance.

There is a psychological background which requires the emphasis of this administrative principle in those laboratories associated with the national military departments. The requirements for instant, unthinking obedience in emergencies dictates the training of military officers throughout their careers to issue brief, clear orders without elaborate explanation. The scientist, on the other hand, is trained not to accept conclusions without examining thoroughly the supporting data. Realization of these backgrounds and adjustment to meet the circumstances has been demonstrated by excellent working relationships between scientists and officers in many of the service laboratories.

2. Technical Reporting

Several years ago, it was the established policy of the Naval Research Laboratory to issue final technical reports on the various research and development problems only at the conclusion of the work on each problem. Interim reports were not issued because it was thought that erroneous conclusions would be drawn from incomplete data. Since many of the problems required a year or more to complete, there was no periodic progress information available to the agencies financing the work. As a result, one agency became dissatisfied and considered withdrawal of financial support. This meant that information exchange between the Laboratory and the sponsoring agency was inadequate. The matter was settled by amending the Laboratory policy to permit the issuance of interim technical reports. Such reports proved to be very helpful to the sponsoring agency and to the Laboratory as well. Interim reports speeded the application of the results of the Laboratory's work and were a positive factor in co-ordinating the program of the laboratory. No serious incorrect conclusions were drawn from incomplete data because conclusions under such circumstances were either held for a subsequent interim report or were given with proper caution.

Nowadays, many large laboratories are required by their sponsors to submit regular progress reports. Often these reports are required every month and give a few lines indicating the progress or lack of progress on every problem that the laboratory has. Since no background of the problems is given, the progress report under such circumstances is almost meaningless to anyone who is not familiar with the program of the laboratory. A monthly report of progress on individual research and development problems is usually too frequent except where considerable diversified effort is being expended. A method for handling this problem, as applied at the Naval Research Laboratory, may be of interest. The Technical Information Office of NRL issues a monthly report of progress based on material supplied by the scientific divisions, but not all problems are reported every time. Instead, only those showing significant progress are described. Each problem report sketches the background of the problem and provides a report of its progress. The problems reports are grouped into broad Laboratory programs. Since the problems are reported only when significant progress is attained, there is space for a reasonably complete report on those described.

An additional feature of the NRL Monthly Report of Progress worthy of note is its inclusion of three to five technical articles written by members of the staff on Laboratory work of particular importance. These articles are long enough to permit a reasonably complete technical presentation of the subject. Publication here does not preclude publication of unclassified material in professional journals because the Report of Progress has a restricted distribution.

3. The Library

The laboratory library is one of the most important tools used in the indirect-contact type of technical-information dissemination. It is not considered to be a management tool, but management recognizes that research cannot be conducted without good library services. Such services have been described by Hooker[5] and others. Bush[6] has said: "Publication has been extended far beyond our present ability to make real use of the record." This is all too apparent to the scientist who tries to keep abreast of his chosen field by reading the many scientific journals, abstracts, books, and reports that pour out each month. About all that he can hope to do is to note articles or items in a few fields of narrow specialization that are of particular interest to him. Even with this screening, he cannot retain very much of what he reads for very long. It is particularly hard to relate associated items that are read at different times. Photostating is used by many researchers to help in this regard. When an item is discovered in one of the worker's narrow fields of specialization, a photostat is made. As time goes on the photostats of associated items are accumulated in separate folders for each of the narrow fields. In this way, items are gathered together from widely separated sources, to form what may well become the basis for a new book on the special subject. The process described contains the elements, but not the gadgetry, of Bush's "Memex." For very large quantities of material, microfilm can be used in the same way, but it has the inconvenience of requiring special film-reading equipment. Hence, its use is generally confined to the laboratory library. The above example implies to management that adequate library, photostating, and microfilm facilities should be conveniently available to the laboratory personnel.

4. Training Program

The present shortage of trained scientists makes it particularly important for management to apply the tool of systematized training in order to extend the present supply. Scientists are, by inclination, persons with a desire to learn, and an opportunity for additional academic training serves to attract, improve the quality, and hold good technical people. Many laboratories enlist the co-operation of the local technical schools, which, in turn, conduct credit courses at the laboratory. Usually, the instructors are persons selected from the laboratory staff. Scholastic degrees are thus possible with a minimum of time wasted in traveling to distant points. Sometimes it is possible for thesis work to be a part of regular laboratory research activity.

General informative and educational meetings of interest to the technical personnel of the entire laboratory, or major portions of it, are of considerable importance in efficient laboratory management. Seminars on subjects included in the basic sciences have proved to be particularly informative and stimulating. The success of such seminars rests upon the committee charged with obtaining experts both within and outside the laboratory to lecture on their specialties.

Research "Internships," as described by Hobson,[7] have been used by a few large laboratories to train and acquaint prospective employees with the activities of such organizations. These internships, which sometimes only cover the scholastic summer vacation periods for graduate or undergraduate students, also permit laboratory management to study the capabilities of the individuals before offering regular positions.

VII. Conclusions

This paper has not attempted to formalize what is good and bad practice in all the methods of information exchange in a large research organization. In fact, it is most difficult to do this because there is always the compromise between too little and too much time being spent on such activity. Instead, a few examples have been given showing what have been observed to be fairly effective information-exchange tools under particular circumstances. Many of the examples have been taken from experience at the Naval Research Laboratory by an engineer who finds himself involved in managerial activities as well as technical work, and who acutely realizes the importance of both. It is believed that NRL experience is probably similar to that of other large research organizations.

The general conclusions that can be drawn from the series of examples that have been given turn out to be a list of obvious "truisms." Nevertheless, they are sometimes overlooked by busy research supervisors. They are given below to remind laboratory managers that their main purpose is to serve the working scientist, and not vice versa.

1. Personal contacts between management and subordinate workers, where there is a two-way exchange of information, is preferable to indirect contacts.
2. Personal interest by management serves as a production incentive.
3. Scheduled personal contacts in various forms serve to conserve time, promote teamwork, and co-ordinate the scientific program.
4. Co-ordination of a large research program is aided by direct personal contacts between subordinate members of the different organizational groups that are involved in the program.
5. Indirect-contact media should be used to supplement and not supplant direct contacts between management and subordinates.

In conclusion, it is evident that information exchange as a management tool has become increasingly important in large research organizations where problems of co-ordination are complex. It is hoped that further discussion of this tool will increase its usefulness to laboratory managers and to their subordinates, thereby forming more enthusiastic and effective teams to attack the fascinating problems of science.

[5] Ruth H. Hooker, "Naval Research Laboratory Library," *Special Libraries*, vol. 35, pp. 442–444; November, 1944.

[6] Vannevar Bush, "As we may think," *Atlantic Monthly*, vol. 176, pp. 101–108; July, 1945.

[7] J. E. Hobson, "Men in research," Proc. I.R.E., vol. 36, pp. 650–651; May, 1948.

Electrometer Tubes for the Measurement of Small Currents*

JOHN A. VICTOREEN†

Summary—**The measurement of small currents of the order of 10^{-12} ampere by thermionic electron tubes requires special considerations not ordinarily encountered with larger currents. Electrometer tubes are especially designed for this service. The necessary differences between ordinary electron tubes and electrometer tubes is discussed from a practical viewpoint. Plate current and plate voltage are much lower in electrometer tubes, and this presents problems both in tube and circuit requirements. Low grid current and high leakage-resistance techniques are presented. Characteristics of such tubes are given in terms which are generally applicable.**

Introduction

THE MEASUREMENT OF radiant energy from radioactive elements is distinguished by the fundamental nature of the determination. This is most easily explained by considering beta rays or high-speed electrons. Although these may be ejected in nuclear transformations, they are, nevertheless, identical with all other electrons. Each electron carries the basic electrical unit of charge and is, in every way, the same as an electron ejected from a hot cathode by thermal agitation. Electrons, when collected without secondary effects and passed through a wire, comprise an electric current, and 6.24×10^{18} of these per second flowing through a wire produce 1 ampere, so that the passage of each electron represents $1/6.24\times10^{-18}$ ampere. A single ampere, being composed of so many single units, appears as a continuously flowing or direct current under ordinary circumstances; but when very small currents are being measured, fluctuations in current are observed due to the statistical properties of the number of small units.

It is often convenient to collect electrons and measure the quantity collected, rather than the number which flow through a wire per second. In this manner, measurement can be deferred until enough of them have been collected to be conveniently measured. Each electron carries 4.8024×10^{-10} esu of charge, and, if deposited without secondary effects upon a perfectly insulated plate, would change the potential of the plate with respect to ground. The electrostatic capacitance of radiation-measuring instruments can be so used for storing electrons over a period of time sufficient to produce a potential difference of a number of volts. When the capacitance C and the potential difference E are known, the number of stored electrons may be calculated easily from the formula $Q=CE$ where Q is the number of electrons times the quantity of charge carried by each electron.

Collecting only the ejected electrons from a given source gives information as to the number of electrons collected, but there are times when it is desired to know the kinetic energy content of the moving electrons, as well as their number. In this case, the high-speed electrons may be absorbed in a gas until their motion is stopped by ion-pair production. Each ion produced by collision liberates one electron as part of an ion pair and reduces the kinetic energy of the moving electron until, after repeated collision, it finally comes to rest. A single high-speed electron is capable of liberating many identical electrons by ionization, and the number liberated is proportional to the energy content of the original electron.

In general, corpuscular-radiation measurements are made by utilizing the primary and secondary effects of the radiation to make available electrons which may be collected or transported and measured in accordance with the simple principles of fundamental electricity or electrostatics.

Under these circumstances, it is not surprising that electrostatic instruments were the first to be used in investigating the properties of corpuscular radiation, for, in some cases, they are uniquely adaptable to measuring very small charges or potentials. As the name implies, electrostatic instruments provide a deflection caused by the electrostatic fields acting upon the moving element resulting from the potential difference of the deflecting part. These forces are truly static in character, and, being due to the charges of individual electrons, will remain constant until some of the electrons are transported from one part of the system to another part. Thus, no current flow is required to maintain the deflection, and the only current taken from a supply source consists of the momentary flow of enough electrons to charge the electrostatic capacitance to the potential of the source.

If an ion chamber is irradiated by a constant radiation source, the chamber becomes a source of a constant number of electrons per second. This number of electrons per second (or current) is independent of the potential applied to the collecting electrodes of the chamber, provided this potential exceeds a certain minimum value, depending upon the design of the particular chamber.

It may, at first, appear inconsistent that electrostatic methods are required to measure a constant current, but the currents obtained are usually so small that they can be most conveniently measured in terms of the potential differences caused by the current flowing through a resistor of the order of 1 million megohms. Any potential-measuring device capable of use across 10^{12} ohms

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must have an internal resistance considerably in excess of 10^{12} ohms, and this condition normally requires an electrostatic instrument.

No attempt will be made here to give details on "electrometer circuits." Emphasis is placed upon the characteristics of electrometer tubes from which circuits may be evolved as required.

Electron tubes, in principle, are electrostatic devices, for the operation of the tube depends upon grid potential only. In principle, the grid of an electron tube should be capable of being charged to a given potential and the connection to it removed, whereupon the grid should remain at this potential, and plate current should be an index to the grid potential.

An electrometer using an electron tube is identical in utility with any other electrometer, such as a quadrant electrometer, or a string electrometer, or even a gold-leaf electroscope with an indicating scale. They are all constant-potential measuring methods which are not supposed to take current from the source of potential under measurement.

An electron-tube electrometer is equivalent to a gold-leaf electroscope or string electrometer when a suitable capacitor and charging source are added in the grid circuit. Once charged, the capacitor will retain its potential, which is indicated by the plate current. An electron-tube electrometer should do everything that might be expected of its static counterpart. At the same time, it is more accurate and rugged.

In practice, ordinary tubes fail even to approach the requirements of electrometer service. Usually, the tube base itself provides too great a leakage path for electrons. Most tubes exhibit spurious grid currents from internal ionization and photoelectrons which are prohibitive. Tubes can be obtained, however, which are essentially electrostatic in character, and these are available from several manufacturers. Originally designed for laboratory purposes, some are not suited for portable instrumentation.

The VX-series tubes with 10-ma filaments were specifically designed for electrometer service in portable radiation-measuring instruments, and will be used for illustration.

By the simplest definition, a three-element electron tube is a device in which the current flowing from cathode to anode is controlled by the potential of a grid to which no current flows. Under this definition, an operating triode would show no change of plate current if the grid lead were disconnected; for, if no current flows to or from the grid, its potential will not change. Tubes which can do this are called electrometer tubes. It is a paradox that, of the many millions of electron tubes which have been produced, few have been required to fulfill this definition in its strictest sense. It is also fortunate that, in the vast majority of uses, it is not necessary; for many practical difficulties arise, both in the manufacture and in the use of tubes under conditions where almost perfect operation is to be attained.

The rapid growth of electronic instrumentation, particularly in nuclear physics, has initiated the demand for tubes which will fulfill the definition as nearly as possible. All electron tubes have combinations of cathodes, grids, and plates. Tubes for instrumentation purposes, and electrometer tubes in particular, must, therefore, differ most from other tubes in perfection of small details. It is the present purpose to describe some of these significant details and, where possible, to offer practical suggestions to those working in the field of instrumentation.

One of the most difficult applications of electron tubes is for instrumentation where dc potentials are to be indicated, as in so-called electrometer circuits. In many cases, it is impossible to define a given tube as an electrometer tube or a circuit as an "electrometer circuit" because, under suitable conditions, any of the tubes described may be used as an electrometer. Such tubes may well be called instrument tubes. All of them are intended to operate with a dc "signal" on the control element (which may or may not be the grid) which changes the output current, causing a desirable indication (which may or may not be a meter in the plate circuit).

The word "electrometer" is properly applied only where it refers to a tube or circuit in which the control element may have an input leakage resistance of the order of 10^{16} ohms or grid current of 10^{-15} amperes.

We shall consider here that an "instrument" tube is one having exceptional static stability. Such a tube may be used in connection with an electrometer tube in many ways, or which, under suitable conditions, may be used as an electrometer tube. The choice of circuit varies radically, depending upon many factors which are not usually encountered or are of negligible importance in ac amplifiers.

In dc instrument circuits, it is best to eliminate the use of the word "amplifier" whenever possible, for it often leads to confusion. This becomes evident when a meter is used directly in the plate circuit of a single electrometer tube, in which case the deflection in milliamperes obtained for a change of one volt on the grid is, naturally, milliamperes per volt. If the meter has a reasonably low internal resistance, this defines the transconductance of the tube; and the higher the transconductance, the greater the deflection. The amplification factor is not involved.

We may consider a dc instrument circuit as being composed of three distinct tube requirements. Let us consider a most common application where a 1/10-volt change in potential across a 10^{12}-ohm resistor is to be indicated finally on a 250-μa meter. First, we must have a tube and circuit capable of operating with 1 million megohms in the control-grid circuit. That is the sole requirement of the first part. The last or third tube serves only to operate the indicating meter. Its sole requirement is to provide a current of 250 μa through the meter. Now, the voltage output obtainable from the first part may not be enough to operate the last part. The second, or middle, part must then be a suitable tube and circuit which serves as a coupling device between parts one and three. We have specified an input of 1/10 volt for an output of 250 μa, or 2,500 μa per volt. The entire circuit behaves, then, like a single tube having a static transconductance of 2,500 μa per volt. We are not concerned with the amplification factor of the tube, so long as an ordinary meter is used as the indicating element in the output.

When the input and output conditions permit, the coupling tube may be omitted, and part one and part three may be directly connected. In many applications where requirements are not rigorous, the meter may be inserted directly in the plate circuit of the first tube.

If a single tube is to be used with an ionization chamber or other source of current which permits the use of a high resistance in series with it, we may increase the resistance of the input circuit and obtain a proportional increase in deflection in the output. A practical limit of grid resistance, however, is determined by the RC time constant of the complete input circuit, which eventually becomes minutes instead of seconds; or else a value is approached at which potentials due to grid currents or leakage currents constitute a prohibitive proportion of the potential causing the deflection. There is usually no good excuse for tolerating spurious currents greater than 1 per cent in the grid resistor, and RC time constants in excess of 1 second are, at the least, not desirable.

It would appear that an electrometer tube should have the highest possible transconductance, thus making possible the use of a single tube and low grid resistance. Unfortunately, however, grid currents are, in general, proportional to transconductance, and the potential across the grid resistor developed by the grid current is proportional to the value of the grid resistor; thus, the proportion of the output-meter deflection due to grid current is relatively independent of over-all circuit transconductance. Although the magnitude of transconductance is not of great importance, constancy of transconductance does directly determine stability or drift, which ultimately limits the sensitivity.

Transconductance in an electron tube is largely determined by grid-to-cathode spacing and by the power input to the cathode. We may summarize, and conclude that, so far as stability is concerned, the cathode is the important determinant. It may have low power input, but the limit of sensitivity obtained depends upon the stability of the cathode.

The dependence of stability and, therefore, sensitivity on cathode emission puts a restriction on the design of dc instrument circuits. Conversely, cathodes for instrument tubes must have properties which are of relatively little importance in other types of service. For example, in the tube about to be described the filament is freely suspended with spring tension at one end, and there is no contact with any other part of the assembly. If this were not done, variations in thermal loss through contact, as in ordinary tubes, would cause the plate current to assume slightly different values each time the tube was mechanically disturbed.

Change in emission with cathode-heating current is ordinarily not considered in ac amplifier design. In dc instrument circuits it is of major importance, as may be seen in Fig. 1. Curves A and B are for a hearing-aid tube, presumably with a 30-ma nickel filament.

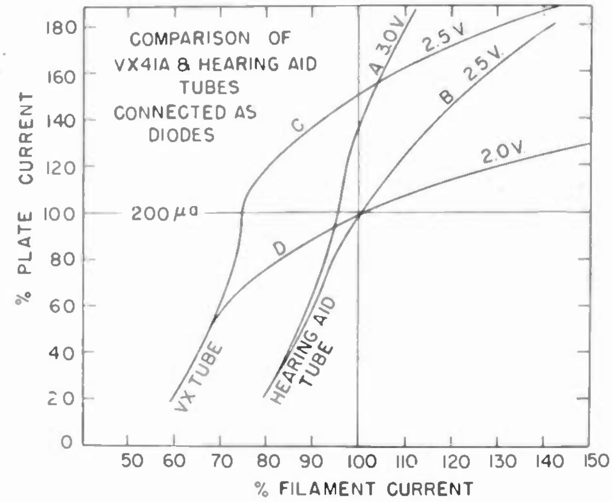

Fig. 1—Comparison of the diode emission characteristics of a VX-41A and a hearing-aid tube as a function of filament-heating current. Graphs are given in per cent so that a comparison could be made at rated values. Each tube was representative of a small group of tubes and should not be taken as representing all hearing-aid tubes. 100 per cent plate current has been taken as 200 μa for both tubes. 100 per cent filament current for the VX tube has been taken as 10 ma, and for the hearing aid tube as 30 ma.

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The difference in microamperes between the two curves for each cathode is the incremental change in plate current for a ½-volt change in plate potential, and represents the conductance as a diode. The VX-tube cathode produces a relatively constant difference over a wide range at either side of the rated operating cathode current. The hearing-aid tube does not have a well-defined saturation point, and the conductance continuously varies with cathode current. Constant conductance and, therefore, transconductance are important in instrument service when dry cells are used for heating the cathode, as the voltage changes from 1.5 to 1.0 during the life of the dry cell, depending upon type. When the first grid is made positive as an accelerator grid, which is usual in electrometer tubes, the curves shown in Fig. 1 represent accelerator-grid current. Tetrode electrometer tubes may be considered as a combination diode and triode, the cathode and first grid being the diode. The triode section then consists of a virtual cathode composed of the openings between the wires of the first or accelerator grid), the control grid, and the plate. Constancy of conductance of the diode section is of the importance under these conditions.

In evaluating cathodes on the basis of diode emission curves, it is difficult to distinguish between a poor cathode and poor exhaust technique, for a potentially good cathode will not be good in a poor environment such as is produced by the presence of gas molecules.

The effects of residual gas on grid currents may be minimized by reducing all positive potentials in the tube to about 4.5 volts, but the presence of a minute quantity of gas may produce poor cathode behavior, even when it is not sufficient to produce an appreciable effect on grid currents.

Too little is known about the theory and behavior of oxide-coated cathodes. The presence of gas in the tube during and after final processing is probably the greatest single contributing factor in the vagaries encountered during use. It is safe to say that it will never be possible to remove all gas molecules from within the envelope. How heroic the attempt can be is determined by the permissible expense.

A cathode is brought into being as an active emitter by heating the cathode to a high temperature during exhaust and to a temperature in excess of operating temperatures after seal-off. The processes are known as activation and aging. Instrument tubes are processed for about 100 hours in such a manner as is found to produce the greatest stability when used as intended. All oxide-coated filaments have a tendency to reach equilibrium under a given set of processing conditions and to stabilize emission at a new point under new constant conditions, the change being slight unless operating potentials are radically changed. If the plate potential is changed from 4.5 to 200 volts under normal operating conditions, the emission current at 3 volts when the tube is diode connected may, thereafter, be changed by 25 per cent.

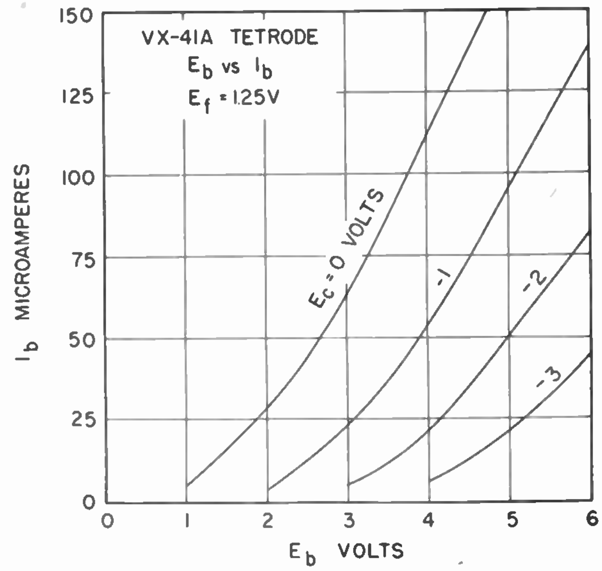
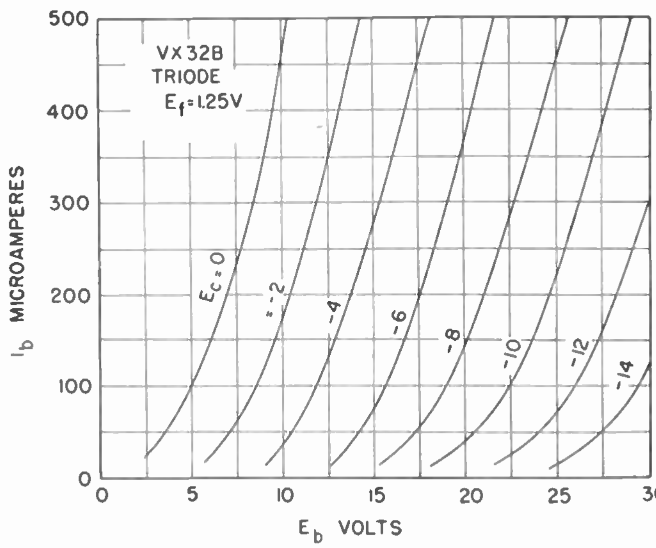


Fig. 2—Plate-current characteristics of a VX-41A tetrode at rated accelerator-grid potential.

Each condition should be stable and give satisfactory performance after time has been allowed for the tube to stabilize. Expensive tubes such as electrometer tubes should be so treated that momentary high voltages will not be inadvertently applied either to cathode or plate. The tube may not be burned out by a momentary overload on the cathode or plate, but much expensive processing may be vitiated in an instant.

It should be recalled that the tentative characteristic curves for ordinary tubes usually are not given below 10 volts plate potential, because of the nonuniformity at low potentials; whereas, in electrometer tubes, the entire operating range may be below 4.5 volts. Figs. 2, 3, and 4 show how well tubes can be made to operate in this region. At low plate voltages the differences between individual tubes tend to become more pronounced, for thermally agitated electrons from the cathode add more



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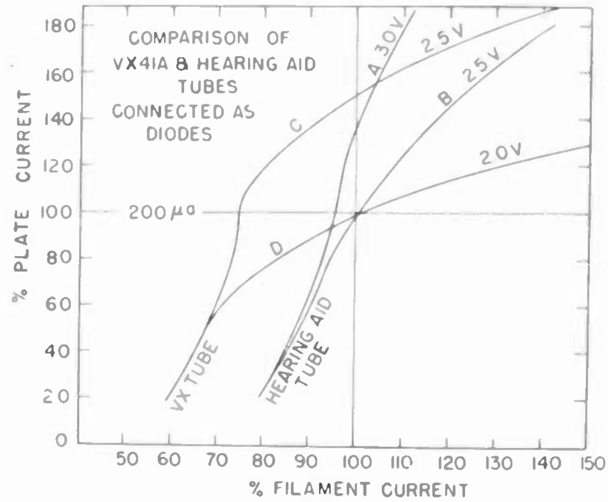

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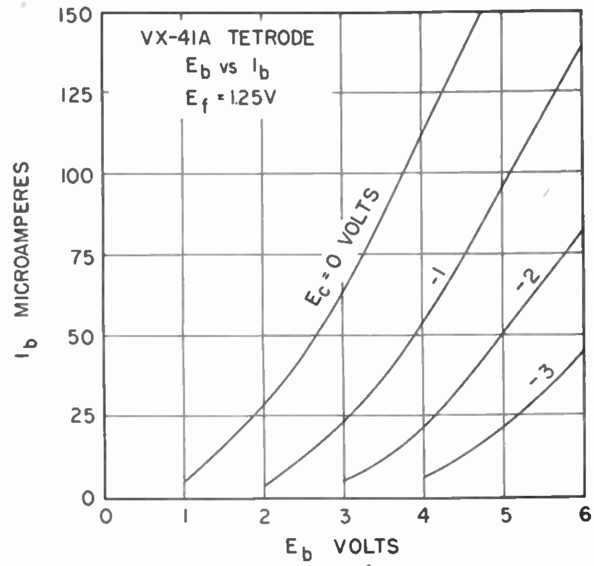

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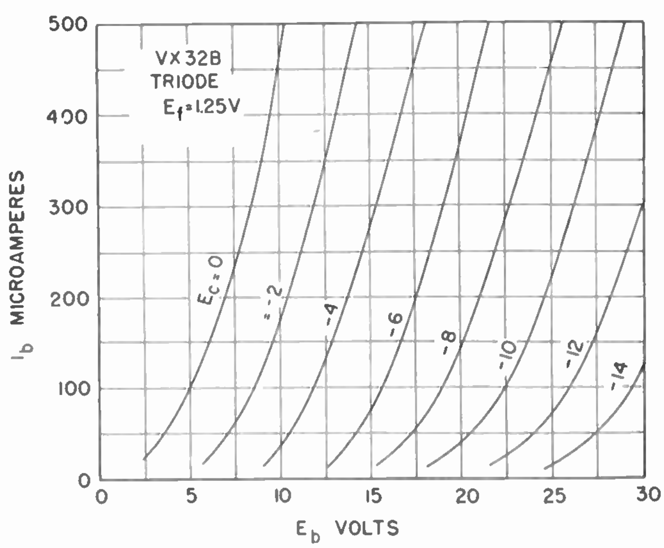

Fig. 3—Plate-current characteristics of a VX-32B triode.

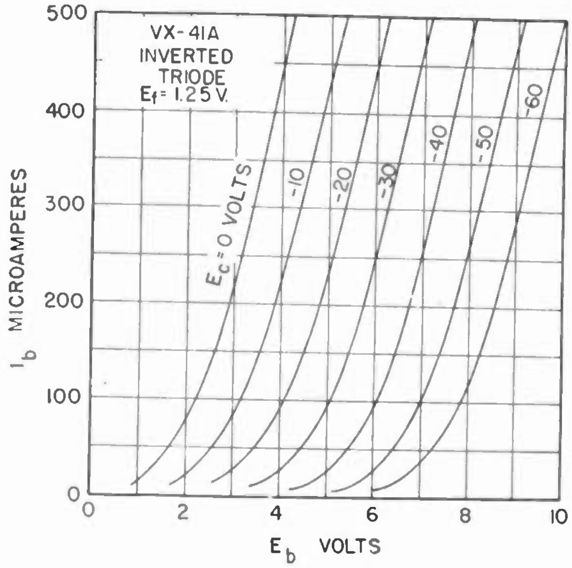

Fig. 4—Plate-current characteristics of a VX-41A tube when connected as an inverted triode.

than ½ volt to the plate potential. In a 1.25-volt filamentary cathode, there also exists a potential gradient of about 1.25 volts from one end to the other; and, as the negative plate potential is returned to the negative end, each single tube is equivalent to an infinite number of smaller tubes in parallel operating at plate potentials of from 3.25 to 4.5 volts. At 1.25 volts plate potential, the potential difference between the positive end of the cathode and plate becomes zero except for the potential of thermally ejected electrons.

In instrument service, the temperature coefficient of resistance of the filament must be considered. In the usual radio tube with a nickel filament, the change in resistance is of the order of 400 per cent from hot to cold. In instrument tubes, it should be of the order of 5 per cent or less. In dc amplifiers, the filament becomes a part of the resistance network; and if it has a high temperature coefficient of resistance, it becomes a nonlinear component which makes it impossible properly to maintain the balance of the circuit with changes in supply potential. The lower temperature coefficient tends to reduce effects due to variations in ambient temperature, and this is particularly true when a constant-current cathode source can be used.

Slight changes in work function of the cathode, contact, and thermal potentials, all ordinarily of no significance, are the limiting design factors of instrumentation, and become more important than the variations in geometry of the tube. Variables of this type largely determine the static plate current obtained under a given set of conditions, and the effects produced are further exaggerated by the necessity of low-resistance plate loads which are required to keep the incremental plate potential within desirable limits. Ordinarily, the manufacturing tolerance of radio tubes is set at about plus or minus 30 per cent for static plate current; this same tolerance can be maintained in the low-voltage region, but the effort is prodigious. All dc circuits must provide for this tolerance by proper design.

VX-type instrument tubes, when used in properly designed circuits, should have a more nearly constant plate current with age and longer life than other tubes. Individual tubes have been operated for 7,600 hours without measurable change in static plate current. The filament should never burn out, but it should be noted that, with a 10-ma cathode requiring only 12.5 milliwatts, it may be instantly destroyed by accidental contact from a small charged capacitor; and it is not likely that the flash will be visible. An open filament is a sign of electrical or mechanical abuse. Because of light cathode mass, relatively stable plate current is attained in less than one second after cathode current is applied; and because of high evacuation, the same plate current is attained after indefinite storage.

The stability of a dc circuit can be no better than the stability of its power source. When cathode, grid, and plate are supplied by dry cells, economy requires that a potential decrease of at least 20 per cent be permitted during the life of the batteries. It is possible to compensate a circuit for changes in any one potential source, but it may be impossible to compensate for the possible changes in a combination of sources. In practical designs, these sources may present more difficulties than the tube itself, so that the number of sources should always be reduced to a minimum. With 10-ma cathodes, it is usually possible to use a single source by inserting a resistor in each leg of the cathode and using the potential drop developed to supply grid and plate potentials, as shown in Fig. 7. This source may be 7.5 or 9.0 volts of dry cells. For ac operation, it may conveniently be a gaseous voltage-regulator tube such as a VR-150 with a large series resistor. In this way, several tubes may be connected in series, with conductive coupling; and with the single source, compensation is possible.. The total load across the VR-150 should not greatly exceed the 10 ma required by the filaments, or the regulation will be impaired. When the series resistor is large, the electrometer series circuit is virtually fed from a constant-current source.

Stability of the cathode emission is, of course, a relative term, but when a single tube is used with a meter directly in series with the plate, no change in plate current should be observable over many hours. If the static current is balanced out and the meter sensitivity is increased, a point is finally reached at which changes in plate current will be observed; but it is only when grid potentials of the order of a few millivolts are to be measured accurately that "drift" becomes serious and, for the most part, unpredictable. Drift in ordinary tubes is often continuous and in the direction of decreasing emission. In instrument tubes, it may be in either direction. In any event, it depends upon the previous history of the tube. For example, if the cathode is permitted to cool beyond the temperature of emission saturation for an instant while any positive potentials remain on any electrode within the tube, rapid drift, which may be called destabilization, will occur, and minutes or hours may be required before the tube is stable again. The effect is present even with a 1-volt positive plate potential.

In very accurate work, all positive potentials should be removed before the cathode-heating current is removed, and cathode-heating current should always be restored before any positive anode potentials are applied. This can be done by a suitable switching arrangement.

The grid circuit must be given careful consideration, for, when a grid resistor is used, any currents flowing through the grid resistor change the grid potential, and any change in these currents therefore appears as instability of plate current. There are a number of sources of current through the grid resistor, and they are all about of equal importance in instrument circuit design. In any electrometer circuit where transconductance is the major factor, the pitch or spacing of control-grid wires does not determine the sensitivity obtained; it only changes the bias required to obtain a given plate current. Hence, the geometry of electrometer tubes is dictated almost entirely by the desirability of getting as much plate current as possible with a negative bias sufficient to prevent negative grid current. Gas current or ionization current within the tube produces positive grid current, which can often be reduced to a satisfactory level by reducing all positive element potentials in the tube to about 4.5 volts or less.

Above this, the effects of grid current rapidly become appreciable when higher grid resistances are used, for 1 micromicroampere produces 1 volt across a 10^{12}-ohm resistor. Negative grid potential has little effect on gas current, and, providing only that the control element is sufficiently negative to repel electrons, it may be raised to several hundred volts without effect except from leakage currents.

Fig. 5 shows the change in total grid currents with grid potential which is obtained from a VX-41A electrometer tube when operated in the dark under recommended conditions. Photoelectric currents may be prohibitive when even a small amount of light strikes the tube.

Positive grid current due to the collection of positive ions by the grid varies with the gas pressure. Even in a very-highly evacuated tube, there are sufficient gas molecules to produce a measurable ionization current. In the presence of a hot cathode, gas pressure changes considerably with time, temperature, and tube potentials. When the magnitude of grid current is sufficiently small so that it does not produce an appreciable potential drop in the grid resistor being used, constancy of grid current is unimportant.

A highly evacuated tube which has not been in use for some time should reproduce previous plate-current values under similar static potentials, but this may not be true with regard to positive grid current. A balance is attained between gas evolution and electrical and chemical clean-up during use of the tube which changes when the cathode is not in use. In a properly evacuated tube, grid current should decrease with age if the tube is in use. It is not unusual for positive grid current to decrease by a factor of ten after one hour's operation of the cathode with or without plate potential. Any reasonable, momentary overload will usually cause some increase in the positive grid current. However, the grid current will return to the original value with continued operation of the tube.

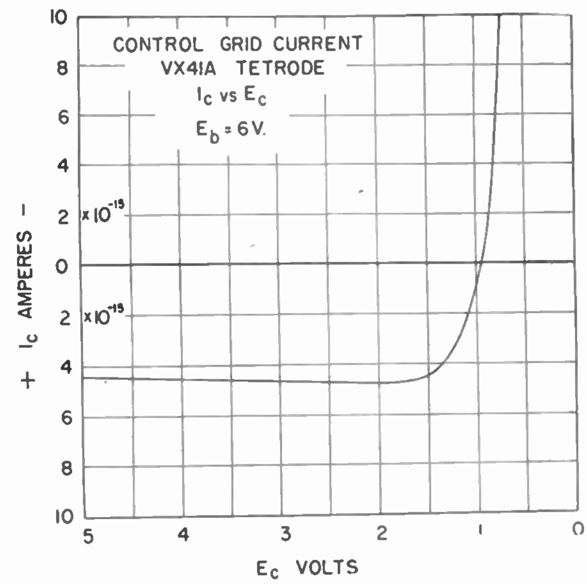

Fig. 5—Control-grid current of a VX-41A tetrode when operated at maximum plate voltage of 6 volts and rated accelerator-grid and filament potential.

In the VX-series tubes, grid current varies with plate potential very roughly with the empirical expression:

$$I_G = mE_p^5$$

where m is a constant determined by the tube type and the particular tube, having a value of about 4×10^{-19}, and I is in amperes.

Actually, the grid current cannot vary exponentially with plate potential, as shown, but must change as various critical ionization potentials are reached. However, a useful approximation is achieved.

Fig. 6 shows the maximum initial values of grid current to be expected at various plate potentials for VX-41 tubes. The average VX-series triode or inverted triode has about the same grid current as the average tetrode at a given plate potential, the main advantage of the tetrode being better static characteristics for low-plate-potential operation in the usual type of circuit.

Positive grid current, in general, varies directly with plate current and is, therefore, reduced by operation of the tube at low plate currents; but variation in grid current does not indicate a corresponding or significant change in plate current. Grid current varies roughly in direct proportion to plate current, and as the fifth power of the plate potential, so that every effort should be made to reduce operating plate potential to the lowest possible value.

Leakage currents within this type of tube are minimized by the use of ceramic or quartz insulation and by supercleansing. External leakage is minimized by special preparation of the glass which, in VX-series tubes, includes a silicone covering which is invisible. This treatment permits the tube to be used under high-humidity condition with minimum leakage.

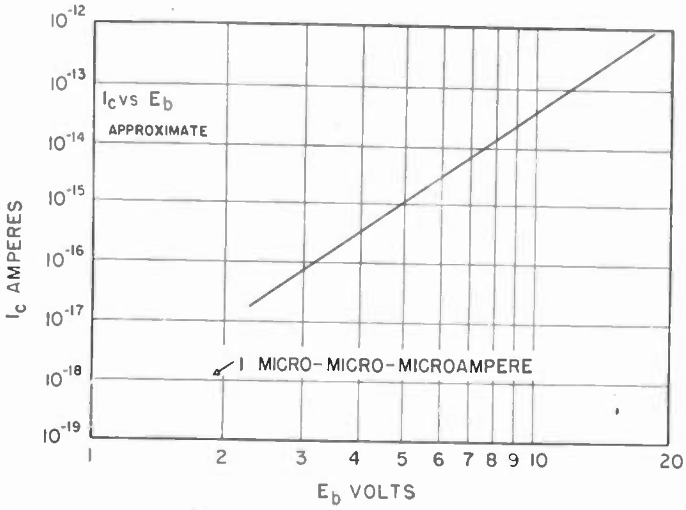

Fig. 6—A very crude approximation of grid current versus plate voltage for VX-type tubes.

When leakage currents of the proper order are not obtained with good electrometer tubes, it is almost certain that external causes are responsible. A fingerprint on the treated surface of the glass may cause endless trouble until removed with absolute ethyl or methyl alcohol and the surface carefully dried. Small pieces of lint give considerable difficulty. The surface of the grid resistor used always should be suspect. Resistors having values up to 10^{14} ohms are available for this type of service; and, because their surfaces have been similarly prepared, they should be handled with great respect. All other insulation which may be used in connection with the grid circuit is equally important. It is recommended with emphasis that only pure polystyrene be used and that its surface be optically polished and cleaned only with absolute methyl or ethyl alcohol. Some grades of polystyrene contain other materials, and these are unsuitable for this service. Capacitors suitable for use in the grid circuit may be made from thin, pure polystyrene; and, when properly made, a 100-$\mu\mu$f capacitor should have a leakage resistance which is in excess of 10^{17} ohms.

When the problems involved and the precautions required are fully understood and anticipated, electrometer circuits are more practical than might be expected. Never will an ounce of prevention pay greater dividends. Instruments using this type of circuit, consequently, must be so designed that dirt cannot get on the input insulation, and so that electrical overloads cannot occur.

Thorough electrostatic shielding and light shielding are required, and all unnecessary insulation should be coated with conducting material to prevent induced or creeping charges from affecting grid potential. A small ac component in the supply voltages also can be insidious. Switches of any kind in the grid circuit should be avoided whenever possible. Aside from leakage, switches add considerable capacitance to the grid circuit, which is undesirable because of increase in time constant. The increased capacitance and inductance loop formed tends to pick up extraneous disturbances. Contact potential is another source of trouble. Wire-wound resistors of low temperature coefficient should be used wherever possible.

A definite advantage of the small VX-tube size is adaptability to mounting at a desired point of measurement. The grid lead may then be kept very short. In ionization chambers, the tube may be sealed within the chamber, which protects it from dirt, humidity, light, and general mishandling. These tubes are not supplied with bases, and for highest possible insulation the tube should be used without a base. The use of base and socket add capacitance and leakage to the circuit, and contact resistance may also be objectional in high-sensitivity circuits.

In instruments where tubes must be replaceable, the electrometer tube may be mounted in a small, moisture-proof capsule or plug-in unit which may also contain the grid resistor and any other resistors which may advantageously be included.

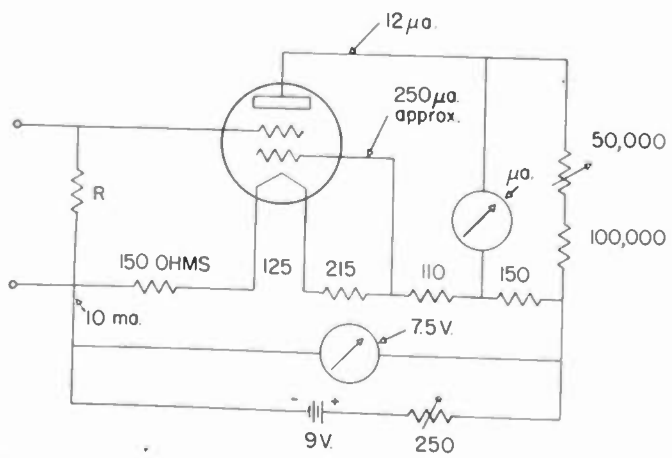

Fig. 7—A typical single-tube electrometer circuit using a VX-41A. Input potential is applied across R, which represents the leakage resistance of all components connected to the grid lead.

As in other simple static electrometers, simple electron-tube electrometers do not produce a deflection which is directly proportional to voltage; i.e., they do not have linear response unless some special provision is made to attain it. Where an electrometer is used only to indicate zero potential, as in indicating the balance of a bridge circuit, the normal plate current may be compensated out, a center-zero meter may be used, and linearity of response over a wide range is then unimportant. A simple circuit using a tetrode is shown in Fig. 7, where the resistance R represents the leakage resistance of the tube and circuit.

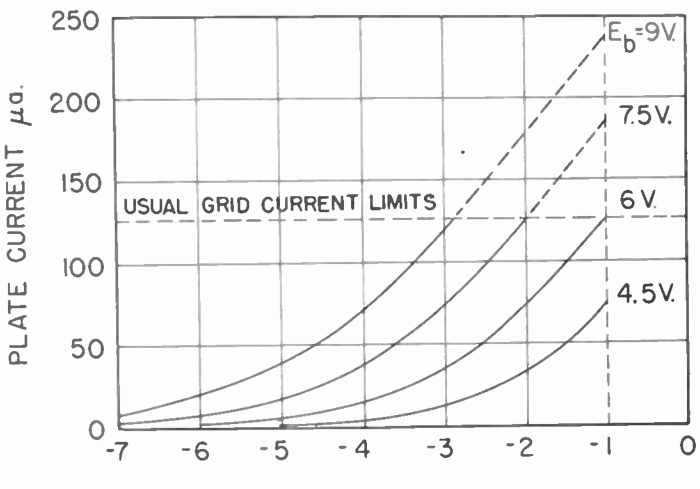

Fig. 8—Portion of the plate characteristic region to which electrometer operation is normally confined.

Transconductance, measured in microamperes plate current per volt of grid potential on the tube used in this circuit, is given in Fig. 8 for various plate potentials. Conditions conducive to low grid currents limit operation to negative grid potentials of at least 1 volt and plate potentials of less than 6 volts, as shown in a region (shown by solid lines) where transconductance is low. Higher transconductance may be obtained by the use of higher plate potential when higher grid currents are permissible, but stability of operation is generally better and grid currents will be lower with lowest plate potential.

By operating the circuit from 5.5 to 1.5 volts negative grid potential and balancing out the static plate current, the curve shown in Fig. 9 is obtained for applied potential E. It may be observed that the change in meter deflection per volt of applied potential is not constant. Occasionally, this nonlinearity may be useful, for large voltage changes can then be tolerated without damage to tube or meter.

The application of positive input potential to the grid is somewhat disadvantageous, because it requires a higher meter suppression current. It is advantageous, however, because the suppression current moves the pointer below scale when the circuit is first energized, instead of up scale, and is, therefore, less liable to damage the meter.

The variable resistor serves to balance the meter to zero, and must be varied as the battery ages. Except for the change due to varying battery potentials, there should be very little observable drift of the zero position when a 50-μa meter is used.

Obviously, in simple circuits it is necessary to maintain a fixed supply potential if the meter deflection is to be constant. This may be done either by occasional adjustment, using a voltmeter as shown, or by the employment of some method of voltage stabilization or compensation.

Zero current through the meter having once been established, the meter may be removed and a more sensitive one substituted, in which case less grid-potential change will be required for full-scale deflection. By using a variable resistor across the meter or in series with the meter, the sensitivity may be adjusted to a desired value.

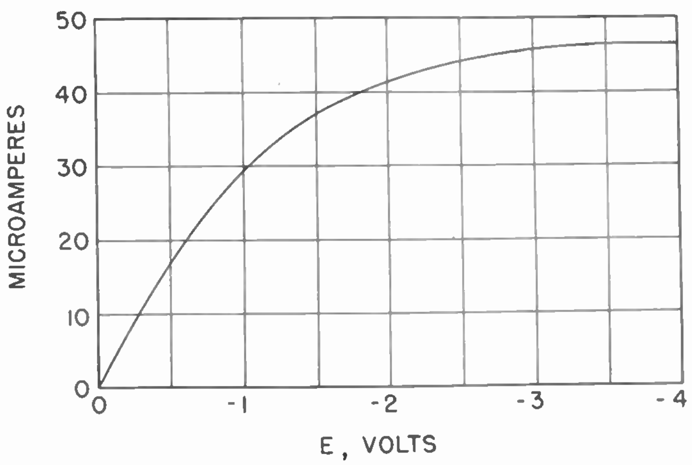

Fig. 9—Transconductance or change in plate current with control-grid potential for the simple circuit shown in Fig. 7.

As the sensitivity of the meter itself is increased, or if a galvanometer is used, a point is reached at which the zero cannot be maintained but constantly drifts due to changes in emission and battery potential. The attainable sensitivity is finally determined by the amount of drift which can be tolerated.

When the output of the electrometer is desired in volts instead of microamperes, a load resistor may be substituted for the meter in Fig. 7 and the supply voltage increased by the amount of voltage drop across the resistor, as shown in Fig. 10; but when high-resistance plate loads are used, with the resulting high plate-supply potential, care must be used that the plate potential at the plate never exceeds 6 volts, and preferably 3 to 4.5 volts.

In balanced circuits employing two tubes, which may be considered as bridge circuits, no current flows through the meter when the circuit is in balance, and it is not necessary to compensate out the static plate current as in the simple circuit of Fig. 7. The balanced two-tube circuit has the advantage that zero drift from potential change is very much less than with the simple one-tube

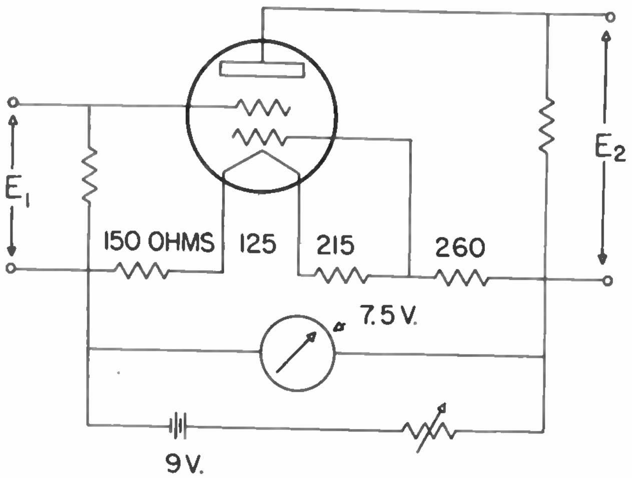

Fig. 10—A simple electrometer-tube circuit as a voltage amplifier. The potential E_2 is generally less than E_1 due to the low amplification factor of most electrometer tubes.

circuit. Both circuits may be satisfactory as indicators, but both have the common disadvantage that sensitivity changes somewhat with supply potential; thus, where absolute sensitivity is important, the supply potential must be held as nearly constant as possible. A simple balanced-tetrode circuit is shown in Fig. 11.

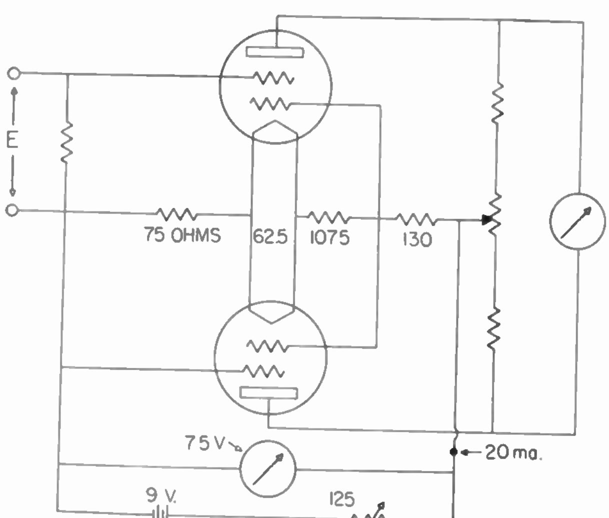

Fig. 11—A general type of the balanced two-tube electrometer circuit. The output meter may be either a center-zero microammeter or a galvanometer, since in the balanced condition no current flows through the meter.

Balanced circuits also have the disadvantage of requiring twice as much supply current, since both filaments are in parallel.

Over small grid-voltage swings, with sensitive plate meters, the linearity of these simple circuits may be satisfactory for many purposes. When large voltage swings are required, the instrument may be made to read directly in volts by calibrating the scale of the plate meter to conform to the nonlinearity.

Higher transconductance may be obtained when higher grid currents are allowable by using a triode instead of a tetrode. Plate characteristic curves of an average VX-32 triode are given in Fig. 3.

Measurement of much larger voltages is made possible by connecting a tube as an inverted triode, as in Fig. 12 for a VX-41A tube. The inverted-triode connection may also be used in a balanced circuit of the type as shown in Fig. 11. Transconductance of a VX-41A inverted triode is given in Fig. 2.

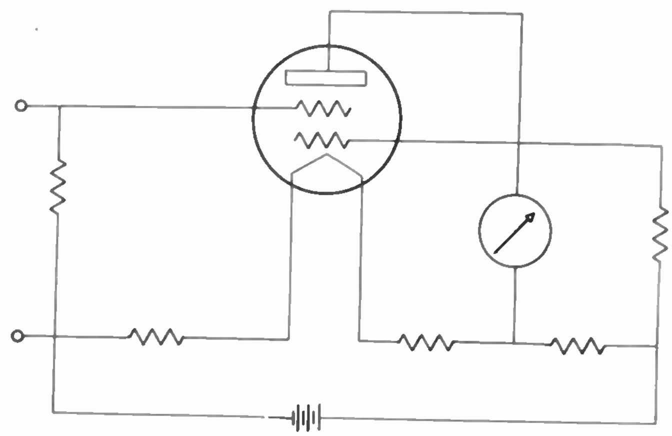

Fig. 12—A method of connecting a VX-41A tetrode as an inverted triode. The connection of the plate to the first grid makes very little difference, and it may be returned to the cathode if desired. It serves as an electrostatic shield for the control grid.

The low current required by VX-tube filaments and the relatively uniform emission characteristics attained make it possible to take full advantage of single-source operation. Under these conditions, a change in supply potential of 1 volt produces a definite change in plate current for a given tube type and associated circuit. This may be considered as supply-plate transconductance and can be expressed in micro-ohms or microamperes per volt, and is a characteristic which will be found useful in determining the behavior of conductively coupled multitube circuits.

Under proper conditions, grid current may be made sufficiently small so that, with the grid lead disconnected and under suitable conditions, the potential of the grid and its associated capacitance of the 2-inch lead will change only a few per cent per hour. A 10^{14}-ohm grid resistor may be used to maintain normal grid bias; but, due to the several $\mu\mu f$ of grid and grid lead capacitance, the time constant of the input circuit may be several minutes. In some work this cannot be tolerated, and it is then necessary to use a lower value of resistance and to obtain current sensitivity in some other way. So long as the potential of the grid with respect to cathode never becomes less negative than about 1 volt (for the VX-41A), the negative potential may be increased to several hundred volts beyond cutoff, and occasionally this may be useful. In some cases, the transconductance may be raised when higher grid currents are permissible by increasing the 4.5 volts on the plate to 6 or 8 volts. But, in general, this is not to be recommended, for stability is also impaired.

The two objections usually encountered to single-tube or balanced-pair circuits are nonlinearity, and insufficient output to operate a rugged meter. Both may be overcome by the addition of one or more tubes to the circuit, in which case a resistor is substituted for the meter (as shown in Fig. 10) in the plate circuit of the electrometer tube, and the potential drop across this resistor fed into a suitable vacuum-tube-voltmeter measuring circuit.

Best linearity is obtained by choosing a plate load (or coupling resistor) of high value, and efficient voltage gain requires a value at least four times the plate resistance of the electrometer tube. The potential drop across this resistor necessitates an increase in plate-supply voltage to maintain 4.5 volts at the plate, and with large values of resistance and, consequently, high supply voltages, it is possible to obtain excessive plate voltage changes with changes in grid potential. Because of this, it is usually necessary to effect a compromise between linearity and voltage gain and excessive plate potential (which leads to high grid currents) by using as low a value of plate load as possible, consistent with requirements. Wire-wound resistors should always be used whenever possible, and this, too, makes advisable the use of low values of resistance.

Tetrode electrometer tubes, when used in conjunction with other tubes, must operate as voltage amplifiers, and the amplification factor which can be attained under conditions conducive to low grid current is very low. The voltage gain obtainable, particularly with low resistance plate loads, is usually less than one.

Under these circumstances, the electrometer tube may be considered only as a coupling tube or resistance-matching device. It is clear that any additional tubes must constitute a vacuum-tube voltmeter capable of operating the desired indicator.

E_2 (Fig. 10) is usually of the order of E_1, so that stability and drift in the additional circuit are at least as important as in the electrometer section. Large negative feedback may be used to stabilize the additional circuit; but, because of the narrow limits of potentials to which the electrometer tube is confined, it is difficult to feed back satisfactorily into the electrometer tube circuit.

In a single-tube or balanced-tube circuit, a tube with the highest possible transconductance consistent with grid-current requirements is desirable when a meter is used for indication. In all other types of circuits using electrometer tubes as coupling tubes, a high amplification factor is preferred.

A two-stage dc circuit has an advantage over a single tube or balanced pair in that, by proper choice of circuit constants, it may be made to have good linearity of transconductance. It is also possible to choose constants which will reduce changes in zero drift due to even a relativity large change in supply potential.

When a single tube is used with a single supply source, the plate current decreases with decrease in supply potential. When two tubes are used in cascade, the plate current in the last tube usually increases with decrease in supply potential, each additional tube reversing the plate-current change. In general, the change in plate current per volt of supply potential becomes greater as the number of tubes is increased; but, because each additional tube reverses the change, it is possible to add a tube in the series in such a manner that this change is reduced or eliminated over a limited range.

Constancy of the zero indication of the meter with change in supply potential does not necessarily mean that the magnitude of deflection will be constant with changes in supply potential. For precision work, an adjustment is advisable for the supply potential to compensate for changes due to battery life, and the like.

The magnitude and direction of ultimate drift in a multitube circuit is difficult to predict, for, as previously stated, the direction of drift reverses with each successive tube, and its magnitude depends upon the individual voltage gain of each stage. With a single supply source, it will at least be a minimum. Although the number of tubes required to produce a given over-all transconductance will be greater when small cathodes are used, the drift is less. Large cathode tubes usually have high drift rates when used with low plate current and low plate potentials.

In multitube circuits, it is very important that the resistors be wire-wound and used well under their rated current capacity, and, also, that they be all of one identical type; i.e., of equal temperature coefficient, which should be as low as possible. Values given for fixed resistors must be closely followed, for, in conductively coupled circuits, any change in the first-tube circuit may have profound effects. In multistage dc circuits, any change in resistor values must be done progressively, beginning at the first tube, so as to assure that finally each tube is operating under optimum conditions.

In nearly all multitube circuits, the linearity of transconductance depends largely upon the last stage; and, where relatively large grid swings are required on the grid of this tube, linearity may be improved by the use of a relatively low plate load resistor in the preceding stage. The resulting nonlinearity in this tube, being of opposite polarity, tends to increase total linearity at the expense of efficiency.

Sensitivity may be increased by using three tubes in such a manner that compensation can be retained. This can be done in a number of ways, but a three-stage circuit of high sensitivity, conductively coupled, is very critical to absolute values of plate current.

A Stereophonic Magnetic Recorder*

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Summary—A high-quality, three-channel stereophonic recorder and playback unit was built for experimental work, to determine the requirements of a unit suitable for home entertainment. The experimental unit was purposely designed to give higher quality than is required for the home, so that the various characteristics could be degraded until performance was unsatisfactory. Results of tests indicate that two channels are adequate for a small room, and that a control track for volume expansion is unnecessary. Experiments were made with loudspeaker and microphone locations, best results being obtained with a "dihedral" mounting of two loudspeakers. This arrangement allows the design of a stereophonic reproducer in a single cabinet, and appears practical for home use.

Introduction

When one listens directly to an artist or orchestra, one of the most important factors which contributes to naturalness of the sound is binaural hearing. The left ear of the listener picks up a sound which is different in amplitude and phase from the sound picked up by the right ear. The two sounds are combined by the human hearing mechanism in such a way that the listener can judge the direction from which the sound comes, and, in addition, the psychological effect is to create a feeling that the sound source is "present."

One of the shortcomings of present-day sound-reproduction systems is that they are single-channel, and give a "flat picture" of the sound, rather than a three-dimensional presentation. To obtain binaural sound reproduction, a two-channel system as illustrated in Fig. 1 may be used. Sound is picked up by microphones placed in the "ears" of a dummy head. Each microphone has its own amplifier, which is fed into corresponding earpieces of a set of headphones. A person listening to sound through a binaural system has the illusion that the sound originates in the room, rather than in the phones. The effect is very striking to one who is used to hearing monaural sound from a headset.

Since wearing earphones is inconvenient, it is advantageous to set up a system with loudspeakers. A loudspeaker arrangement of this type is called a stereophonic system. Although it cannot give the results of a true binaural system, for a number of reasons, the improvement over monaural listening is quite marked. Experiments with binaural and stereophonic sound systems in the past have been made with certain objects in mind:

(a) Laboratory experiments to investigate the mechanism of binaural sound.

(b) Demonstrations to illustrate the advantages of binaural over monaural sound.

(c) Grand-scale demonstrations to show the utmost in high-fidelity reproduction.[1,2]

(d) Systems to give the illusion of side-to-side movement in motion pictures, corresponding to the source pictured on the screen.[3]

(e) Systems to create novelty effects in connection with motion pictures (the sounds not necessarily being intended to give the illusion of originating from a source pictured on the screen).[4]

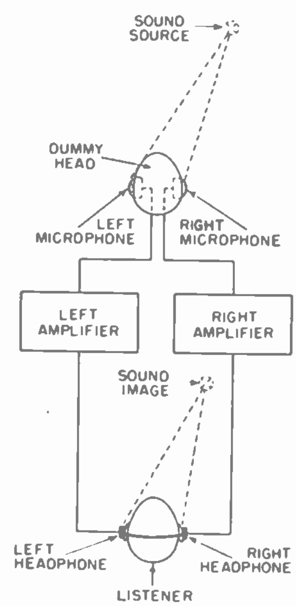

Fig. 1—Binaural sound-reproduction system.

Efforts to bring stereophonic sound into practical use have been directed toward entertainment in the theater and concert hall, since it became apparent that the complexity and cost of a stereophonic system was greater than any other single project that had yet been attempted in the field of sound reproduction.

For example, the "portable" equipment for "Fantasia's" road show had eleven 62-inch racks of amplifiers, plus power supplies and other equipment. It was

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† Armour Research Foundation, Chicago, Ill.

[1] "Symposium on auditory perspective," *Elec. Engineering*, vol. 53, p. 9; January, 1934.

[2] H. Fletcher, et al., "The stereophonic sound film system," *Jour. Soc. Mot. Pic. Eng.*, vol. 37, p. 331; October, 1941.

[3] J. P. Maxfield, "Demonstration of stereophonic recording with motion pictures," *Jour. Soc. Mot. Pic. Eng.*, vol. 30, p. 131; February, 1938.

[4] W. E. Garity and W. Jones, "Experiences in road-showing Walt Disney's 'Fantasia'," *Jour. Soc. Mot. Pic. Eng.*, vol. 39, p. 6; July, 1942.

packed in forty-five cases weighing an average of 330 pounds per case, and occupied half of a standard freight car.

It appears that, even for theater use, which could stand the cost, the complexity of optical sound-on-film stereophonic systems has been great enough to prevent them from being adopted generally. Magnetic recording has been suggested, but it is still in the experimental stage.[5,6]

Stereophonic sound is, therefore, a subject that is always discussed with great enthusiasm, but nobody does anything about it. Everyone agrees that it is an excellent system, but nobody can find a practical use for it.

One field that seems to have been neglected is that of home entertainment. In fact, home entertainment is often cited as an example of a field in which stereophonic sound is entirely impractical. Stereophonic reproduction for the home brings up a number of problems which are entirely different from problems of reproduction in the concert hall or theater. For instance, the listening room in a home is very much smaller, and the listener is closer to the loudspeakers. With loudspeakers on each side of the room, he cannot back away a distance comparable to the distance between loudspeakers; in fact, the seating arrangement is often such that the listener faces the broader wall. Since the listener is free to move about the room to a considerable extent, the stereophonic illusion should be present throughout the room. Acoustics of the room are generally fixed and little can be done about them, so the home stereophonic reproducer should be adaptable to various shapes and layouts of living rooms.

Experimental Stereophonic System

Over-All Requirements

A magnetic recording system seemed to be the only one capable of approaching the economic requirements of a home unit. To investigate the possibilities of home stereophonic entertainment, a tape recorder and playback unit was built to meet the following specifications:

Number of channels	—three
Frequency response	—flat from 50 to 10,000 cycles within 5 db
Distortion	—less than 4 per cent intermodulation distortion, or 1 per cent harmonic distortion at normal recording levels
Dynamic range	—60-db spread between maximum modulation level and noise level
Wow and flutter	—less than 0.1 per cent

[5] M. Camras, "Magnetic sound for motion pictures," *Jour. Soc. Mot. Pic. Eng.*, vol. 48, p. 14; January, 1947.
[6] M. Camras and R. E. Zenner, "Binaural magnetic recorder," presented, Acoustical Society of America, 33rd Meeting, Hotel Pennsylvania, New York, N. Y., May 9, 1947.

The experimental model was purposely designed for better performance than was thought necessary for home application, because it would be used for recording as well as for playback. Also, with the higher-quality system, experiments could be made in which each of the characteristics would be degraded until the performance was unsatisfactory. Thus the requirements for a home system could be established.

Stereophonic Tape

The drive unit for the stereophonic system is shown as the left-hand unit in Fig. 2. It accommodates a 7-inch reel of either $\frac{1}{4}$- or $\frac{1}{2}$-inch-wide tape. At the normal running speed of 1 foot per second, a full reel of $\frac{1}{4}$-inch tape plays for 20 minutes. A full reel of $\frac{1}{2}$-inch tape plays for 20 minutes on one edge, after which it is turned over and played for 20 more minutes on the other edge, giving a total of 40 minutes. Fig. 3(a) shows the arrangement and dimensions of the magnetic tracks on the $\frac{1}{4}$- and $\frac{1}{2}$-inch tapes.

Stereophonic Heads

The stereophonic heads are arranged as in Fig. 3(b). An erase head extends across the entire width of the tape and clears off all three channels. The record heads are staggered along the length of the tape, to permit mechanical and electrical isolation. At the section where head No. 1 is recording on track No. 1, the other two tracks are covered by a keeper made of high permeability alloy. The same applies to the other heads.

For simplicity, the same heads are used also for pickup. In this case, when the No. 1 head is picking up, the adjacent channels are "short-circuited" out by keepers to prevent crosstalk. The other channels are protected in the same way. Without keepers, it has been found that heads are sensitive to recordings on channels as far as $\frac{1}{8}$ inch or more from the head, the effect being especially pronounced at low frequencies.

Amplifiers

Fig. 2 shows the three units which comprise the complete stereophonic system except for the microphones and loudspeakers. At the left is the drive unit, already

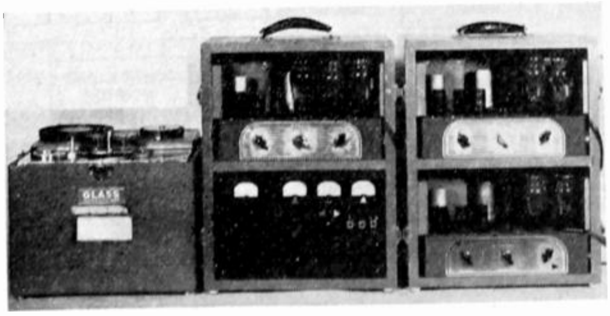


Fig. 2 —Stereophonic system

described. The lower section of the center unit is the master-control panel and oscillator. The upper half of the center unit is the amplifier for channel No. 1, and the sections of the right-hand unit are amplifiers for channels Nos. 2 and 3.

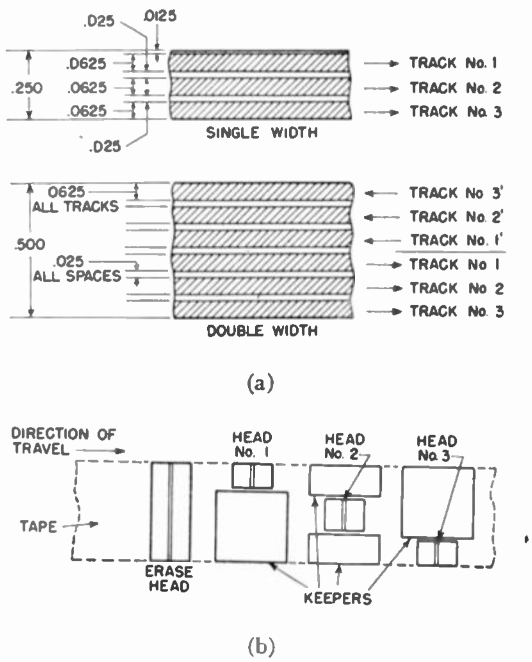


Fig. 3— Stereophonic magnetic tape and heads. (a) Tape dimensions. (b) Head arrangement.

Typical Three-Channel System

A block diagram of a typical recording setup for the three-channel stereophonic system is shown in Fig. 4. Western Electric 639B cardioid microphones were used for pickup of the original sound. Altec Lansing loudspeakers on each side, as well as in the center of the room, were used for reproduction.

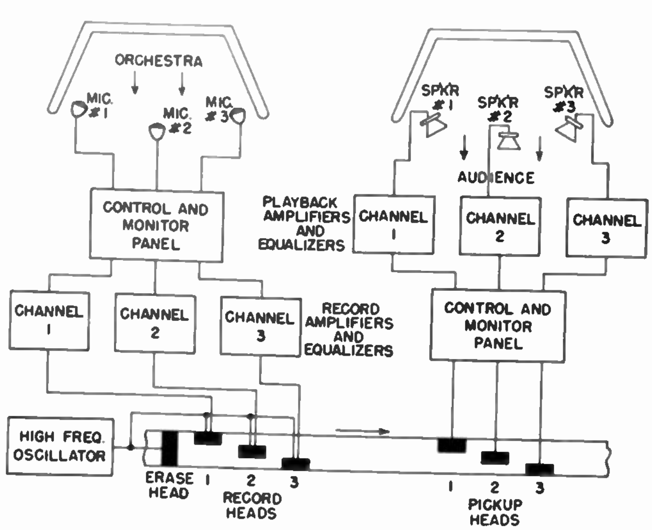


Fig. 4—Block diagram of the stereophonic magnetic recording and playback system

Results of Experiments

Methods of Adjusting Gain

To control the relative gain of the separate channels, the most obvious method is to adjust them separately so that a standard sound intensity in any microphone will set up a standard sound intensity near the corresponding reproducing loudspeaker on playback. Ordinarily, the maximum sound intensity picked up by the different microphones during a rendition will be different. This allows the use of a method for setting as shown

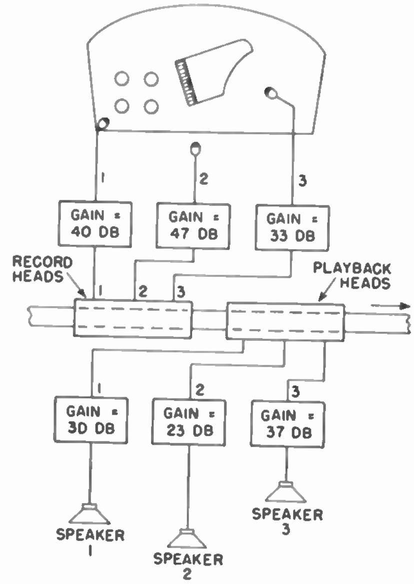


Fig. 5—Method for setting gain of stereophonic channels.

in Fig. 5, where, during recording, the gain of each channel is set at the maximum point that will not produce overload of the recording at any time, On playback, the gains of the amplifiers are set in inverse ratio to the recording amplification. Thus, the playback amplifiers may be set to compensate for the different gains of the record amplifiers, as well as unequal efficiencies of microphones, heads, loudspeakers, acoustic phenomena, etc., and a maximum signal-to-noise ratio is obtained.

Use of a Control Track

With three channels available, one of the possibilities to be explored was the use of two channels for audio, and the third channel as a control track, to vary the volume of the audio channels and thus increase the dynamic range. It was soon found, however, that the dynamic range offered by the audio channels was entirely adequate for home use without resort to compression and expansion. With music reproduction at a comfortable volume level, the ambient room noise of both the recording and the listening rooms was enough to mask the ground noise. It was decided, therefore, that a control channel was an unnecessary complication.

Comparison of the Two-Channel and Three-Channel Systems

Even at the start, it was felt that two channels should be enough for a small room, rather than three. Experiments were made with two and with three channels to determine their relative merits. For solo work and for announcements, the center channel adds realism by giving the effect of clarity and nearness. This is understandable, for, if only the outside channels are used, the sound has to travel a considerable distance before it is picked up by the side microphones, and the ratio of reverberant to direct sound is high. A similar effect occurs on playback through the side loudspeakers. The situation can be improved without adding a third channel by using a separate microphone for the soloist or announcer, and feeding its output into both of the side channels as shown in Fig. 6.[1] An additional loudspeaker could be used on playback, but this also has some disadvantages.

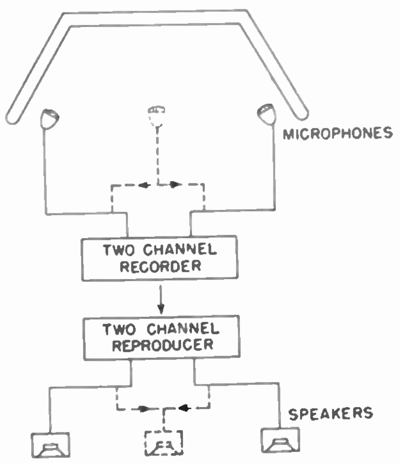
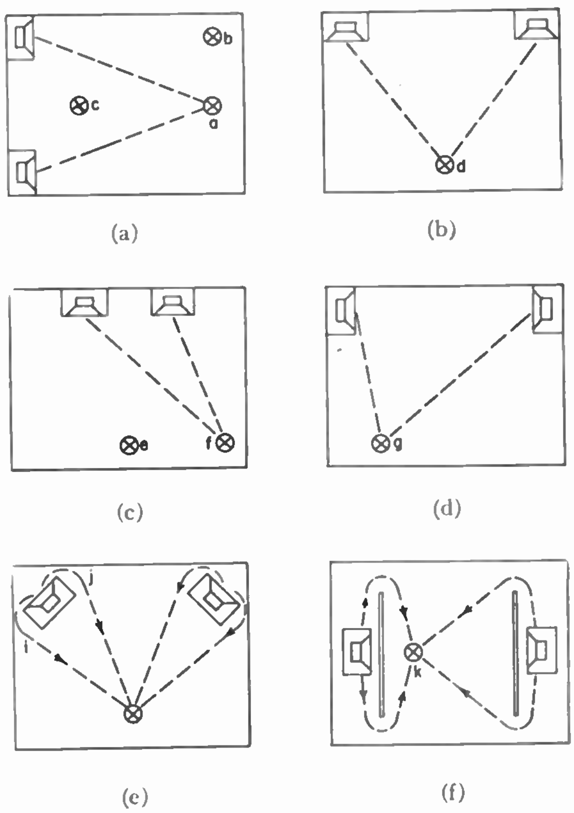


Fig. 6—Experimental two-channel system.

A number of experiments made with orchestral music and other sound indicated that they could be handled adequately by a two-channel system. It would seem that, even for announcements and solos, the lack of a center channel would only give the illusion that the performer is further away from the listener, and for home entertainment this illusion might even be desirable. All things considered, it was decided that a two-channel system gave results that were sufficiently good for a home unit.

Loudspeaker System

The first experiments were made with loudspeakers set up as in Fig. 7(a), since it seemed logical that the best place for the loudspeakers was against the sides of the room and spaced apart. Fair results were obtained with the listener in position *a*, but when the listener stood to one side of the room, as at *b*, the closer loudspeaker predominated. Positions such as *c* were unfavorable because the listener was conscious of two sources. The arrangement of Fig. 7(b) was even worse in this respect.



(a) (b) (c) (d) (e) (f)

Fig. 7—Experiments with loudspeaker placement.

The loudspeakers were now moved closer together in several steps, until they were next to each other. They are shown in an intermediate position in Fig. 7(c). When the loudspeakers were fairly close, the sound no longer seemed to come from two separate point-sources, but at the same time the apparent "spread" of the orchestra was reduced.

To reduce the distorted spatial effects for a listener in a position such as *f* in Fig. 7(c), a sidewise loudspeaker position as in Fig. 7(d) was tried. With such an arrangement, the left-hand loudspeaker is closer to a listener in position *g*, but the right-hand loudspeaker faces him more directly and compensates for the increased distance. By control of the variables, it is possible to work out good compensation over much of the room area.

With all of the above arrangements, a careful listener can usually pick out the location of each loudspeaker, even when blindfolded. It seems that higher-frequency sound can be localized very closely at the loudspeaker cone, especially if the listener is allowed to move his head. When a listener is trying to locate the sound, we find that he turns his head back and forth several times until he has it "fixed." Then he faces the source and points to it.

A listener in the home is not obliged to keep facing in a particular direction, as someone at a concert usually does. Seats in a living room seldom face in one direction. Also, the listener is free to get up and walk about the room. To decrease localization effects, a number of schemes were tried for diffusing the sound. In Fig. 7(e) the loudspeakers were turned toward the corners of the room. The sound has to come from i or j or both, and gives the effect of a larger source. Systems as shown in Fig. 7(f) were also tried. Here some rather large baffle boards were put in front of the loudspeakers to prevent direct radiation. This arrangement gave the interesting effect of having the entire room filled with music, and was effective over a considerable portion of the room area.

A great many other schemes were tried until one was evolved which gave exceptionally good results. This system is shown in Fig. 8. The loudspeakers are placed fairly close together and at an angle to each other facing the wall. Upon first inspection it would not seem that this system could give results as good as with widely spaced loudspeakers. Yet actual listening tests showed that the sound seemed to come from all parts of the front wall (such as p, q, r, s,), as well as from the sides (areas n, o, t, u). The illusion was present over almost the entire room. Troublesome echo effects that were noticeable with the other loudspeaker arrangements were absent with the new system.

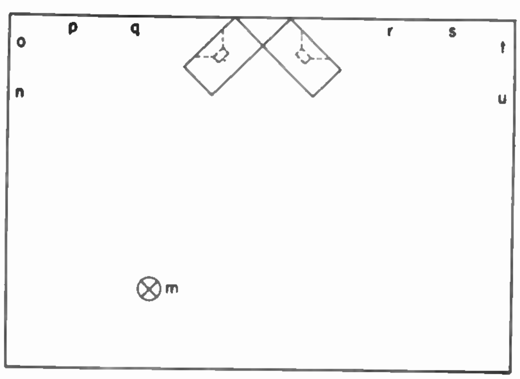


Fig. 8—Dihedral loudspeaker system.

A possible explanation of this effect is given by Fig. 9(a). The reflected sound from loudspeaker (1) acts as if it came from the virtual source (1′), which is located over to the left, beyond the room boundaries. Similarly, virtual source (2′) is located beyond the right-hand wall. Only the short-wavelength higher-frequency sounds will behave in this manner, but it is this class of sounds that the ear uses for localizing a source. As has been noted before, the high-frequency components beaming from a loudspeaker are responsible for the feeling of a point source when an observer faces the loudspeaker. A good proportion of the sounds arrive directly from the loudspeaker and by reflection from the front wall, so that the sound sources appear pretty well spaced along the front. The acoustic conditions can be modified by treating the walls at x and y. Draperies hung at x gave improved results under some conditions.

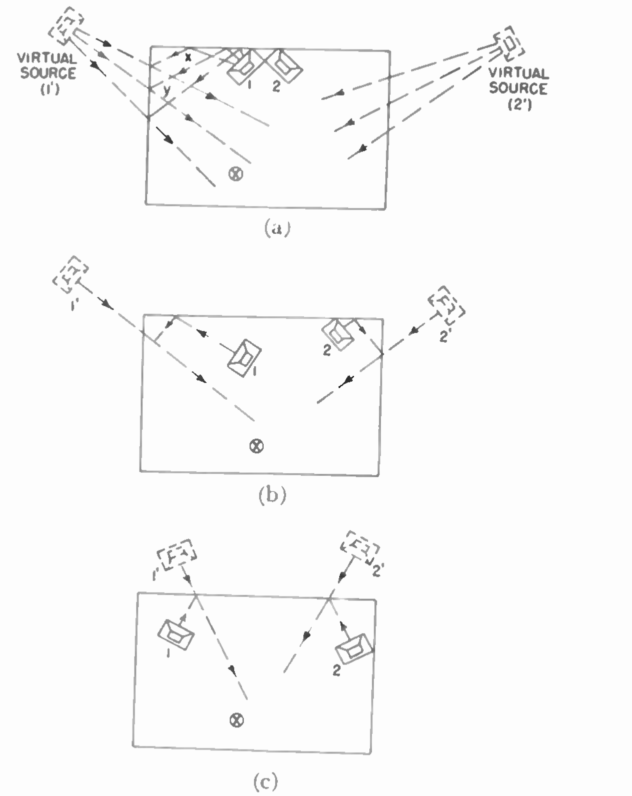


Fig. 9—Operation of reflective loudspeaker systems.

By changing the loudspeaker location, the virtual sources can be moved as shown by Fig. 9(b). They can be moved inside the side wall boundaries, if arranged as in Fig. 9(c). The arrangement need not be symmetrical, as in Fig. 9(a), but can be nearer to one corner, etc., depending on the interior decoration scheme, location of doors and windows, and other acoustical conditions. Relative input power to each loudspeaker should also be adjusted, and some extra emphasis of high frequencies may be desirable in order to compensate for absorption of the walls.

Microphone Placement

For a dihedral sound projection system, placement of the original pickup microphones is important. They could be spaced far apart, at the approximate location of the virtual loudspeaker images of Fig. 9(a), if the recording studio is large enough. Or, if the added reverberation is not detrimental, they could be substituted for the loudspeakers in the setup of Fig. 9(a). An excellent arrangement is shown in Fig. 10. Directional microphones are used, spaced apart about the same distance as the loudspeakers. A sound-absorbing baffle can be placed between the microphone for improved isolation. With pickup as in Fig. 10, the phase relations of reproduced sound from dihedral loudspeakers can be made correct, regardless of the relative acoustics of the record and listening rooms. An announcer or soloist can operate in front of one or both microphones with results that are very satisfactory.

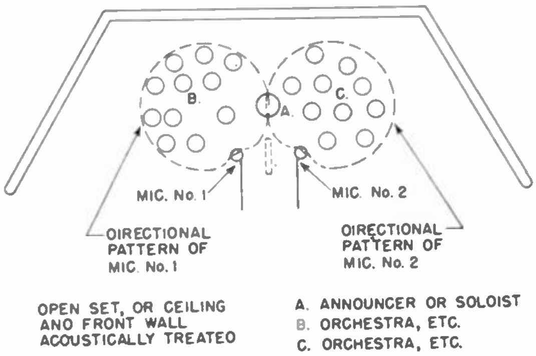


Fig. 10—Microphone placement for dihedral recording.

Home Stereophonic Unit

As a result of tests with the experimental stereophonic system, the design for a home model shown in Fig. 11 was evolved. The dihedral system of projection allows the machine to be built as a single unit. The drawers at the top house the stereophonic tape drive, the dual-channel amplifier, and a conventional-type radiophonograph combination. Two loudspeakers in bass-reflex housings are set at an angle and face toward the back. A grille-work on the face is for decorative purposes, although it is possible to put a third loudspeaker in front for a center channel.

There is a marked advantage in using the dihedral loudspeaker system, even with the conventional single-channel radio or phonograph. The sound seems to come from an area rather than a point, and is more full and more pleasant than with a single loudspeaker in front of the cabinet.

Whether there may be stereophonic AM or FM broadcasting sometime in the future is hard to predict. If there are a sufficient number of home stereophonic units, it is conceivable that large metropolitan areas would have at least one stereophonic broadcasting station. The second channel could be received by a separate external adapter unit, if a dual receiver were not already built into a large set.

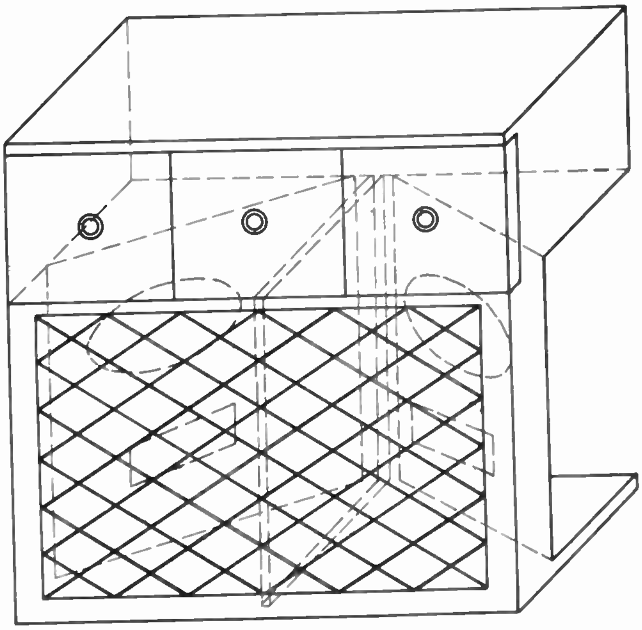


Fig. 11—Home stereophonic unit.

Conclusions

A home stereophonic system has been developed which provides new listening pleasure for home entertainment. The unit is usable in the average living room, and its cost is low enough to allow widespread use.

Transient-Response Equalization Through Steady-State Methods*

WILLIAM J. KESSLER†, ASSOCIATE, IRE

Summary—**This paper describes a steady-state method for matching the pulse-response characteristics of two or more networks using only a sinusoidal signal generator and a cathode-ray oscilloscope. Photographs of the patterns displayed on the screen of the oscilloscope, and analyses to permit rapid determination of the network adjustments to effect equal pulse-response characteristics, are presented.**

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Introduction

For some purposes, the relative transient response of two or more networks may be of greater interest than the absolute transient response of either network alone. Examples are found in the pulse-response characteristics of laboratory comparators[1] and instantaneous cathode-ray-tube direction

[1] R. A. Watson-Watt, J. F. Herd, and L. H. Bainbridge-Bell, "Applications of the Cathode Ray Oscillograph in Radio Research," His Majesty's Stationery Office, London, pp. 123–125; 1933.

finders[2] employed for the observation of atmospherics. Conceivably, pulses, rectangular waves, or the use of periodic wave forms whose Fourier transforms converge more slowly (sawtooth wave forms, for example) might be employed for the purpose of effecting identical transient response. However, experience has revealed that the resulting patterns displayed on the crt screen when using such methods are sufficiently complex to prevent a rapid analysis for determining the adjustments necessary to effect equal pulse-response characteristics. The method to be described circumvents this difficulty through the use of a steady-state sinusoid of varying frequency. This yields a simple elliptical pattern, delineated on the screen of the cathode-ray tube, which is subject to rapid analysis.

It will be recalled that the transient response of any linear passive network may be completely specified in terms of its sinusoidal steady-state behavior.[3] Referring to Fig. 1, it is clear that, if the networks are to exhibit

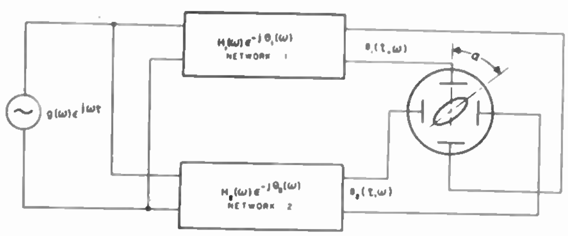

Fig. 1—Block diagram showing the sinusoidal signal generator, the networks to be matched, and the cathode-ray oscilloscope.

identical transient-response characteristics, $e_1(t, \omega)$ and $e_2(t, \omega)$ must be equal for all values of ω throughout the transmission range. The expression relating the angular position of the major axis of the elliptical pattern displayed on the screen of the cathode-ray tube and the steady-state-response functions is

$$\tan \alpha = [H_2(\omega)/H_1(\omega)]\epsilon^{-j[\theta_2(\omega) - \theta_1(\omega)]}$$

where $H(\omega)$ represents the scalar product of the amplitude functions associated with the individual sections, and $\theta(\omega)$ is the scalar sum of the individual phase functions.

If the steady-state functions are to be equal at all frequencies, then $\tan \alpha$ must remain real and equal to unity throughout the entire transmission range of the networks. Physically, this means that the angular position of the figure displayed on the screen must remain at an angle of 45°, or unit slope, and display no ellipticity as the frequency of the signal generator is swept through the transmission range of the networks which are employed.

[2] W. J. Kessler and H. L. Knowles, "Direction finder for locating storms," *Electronics*, vol. 21, pp. 106–110; May, 1948.
[3] E. A. Guillemin, "Communication Networks," John Wiley and Son, Inc., New York, N. Y. 1935; pp. 474–476.

The procedure employed to equalize the transient-response characteristics of any two networks is, then, to sweep slowly the frequency of the signal generator throughout the transmission ranges of the networks, and make such amplitude and phase adjustments as are necessary to display a closed line of unit slope for all frequencies.

Pattern Analyses and Experimental Results

The adjustments necessary to effect equal pulse-response characteristics for networks similar to that shown in Fig. 2 may be determined by analysis of the changing

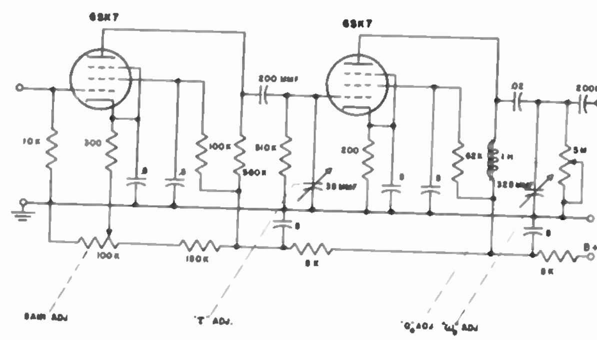

Fig. 2—Schematic of the networks employed to obtain the experimental data.

elliptical pattern as the frequency of the signal generator is varied. Interpretation of the changing elliptical pattern requires a knowledge of the electrical composition of the networks, and hence of the frequency characteristics of the amplitude and phase functions. The phase function $\theta(\omega)$ may be resolved into a number of components (depending on the number of degrees of freedom of the networks) which are, in general, different functions of frequency. Therefore, individual phase controls are required for each network parameter influencing the particular phase displacement. For networks including resonant as well as nonresonant sections in cascade, where each network is no more complex than that shown in Fig. 2, a minimum of four adjustments is necessary. They are: (1) a dissipation or damping (Q_0) adjustment; (2) a tuning (ω_0) adjustment; (3) a time-constant (τ) adjustment; and (4) a gain (G) adjustment, all as shown in Fig. 2.

The gain adjustments may be any convenient means for altering the amplitude responses, with the restriction that the phase changes introduced as a result of gain changes be negligible.

The influence of the relative amplitude and phase functions on the elliptical patterns displayed on the screen of the oscilloscope is best determined by examining the behavior of both the amplitude and phase functions as a function of frequency.

Figs. 3, 4, and 5 illustrate the relationship between the amplitude and phase functions for the resonant and

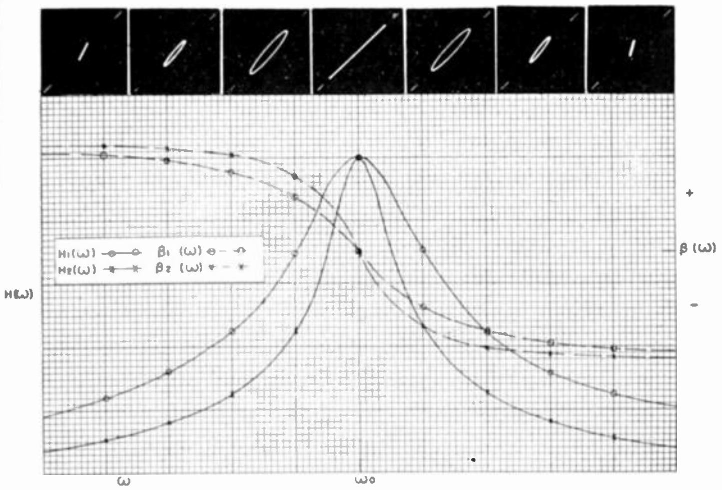

Fig. 3—Response curves of the amplitude and phase functions of the resonant sections, and corresponding photographs of the oscilloscope screen when $Q_{0_2} \neq Q_{0_1}$.

nonresonant sections of the two networks whose schematic is shown in Fig. 2. The photographs of the oscilloscope patterns correspond only to the particular relationships depicted by the accompanying curves. It is observed that the amplitude and phase curves of the charts describe only a portion of the networks, whereas the accompanying photographs refer to the networks in their entirety.

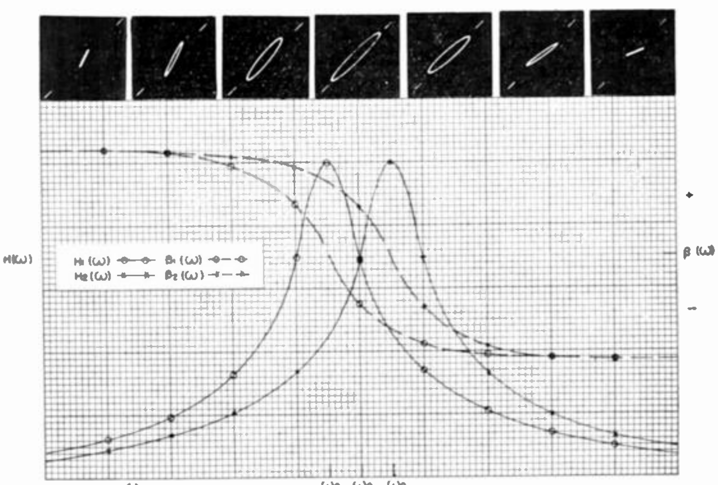

Fig. 4—Response curves of the amplitude and phase functions of the resonant sections, and corresponding photographs of the oscilloscope screen when $\omega_{0_2} \neq \omega_{0_1}$.

The phase displacements and amplitude responses as a function of frequency for the resonant sections of each network when $Q_{0_2} \neq Q_{0_1}$ are shown in Fig. 3. Also shown are the corresponding sequential photographs of the oscilloscope patterns as the frequency of the signal generator is slowly swept through the transmission range of the networks. An examination of the phase and amplitude-response curves shown in Fig. 3 reveals that the amplitude response of network 2 falls off more rapidly than that of network 1, and that the relative phase displacement becomes zero at resonance only. Further, it is noted that this observation holds regardless of the direction in which the frequency of the signal generator departs from resonance. Therefore, it appears to be a simple matter to predict that the position of the major axis of the elliptical figure will become more nearly vertical[4] and the figure more elliptical as the frequency of the signal generator departs from resonance in either direction. This appears to be verified by the corresponding photographs.

The phase displacements and amplitude responses of the resonant sections when $\omega_{0_2} \neq \omega_{0_1}$ are shown in Fig. 4. A similar examination reveals that, although the amplitude response of network 2 falls off *more rapidly* than that of network 1 *below resonance*, the response falls off *less rapidly* than that of network 1 *above resonance*, and that the relative phase displacement vanishes only for large frequency deviations. Thus, it appears that the position of the major axis of the eliptical pattern will swing about the unit slope position, becoming more nearly vertical or horizontal, depending on the direction of departure of the signal-generator frequency from resonance, and the pattern will display almost constant ellipticity.

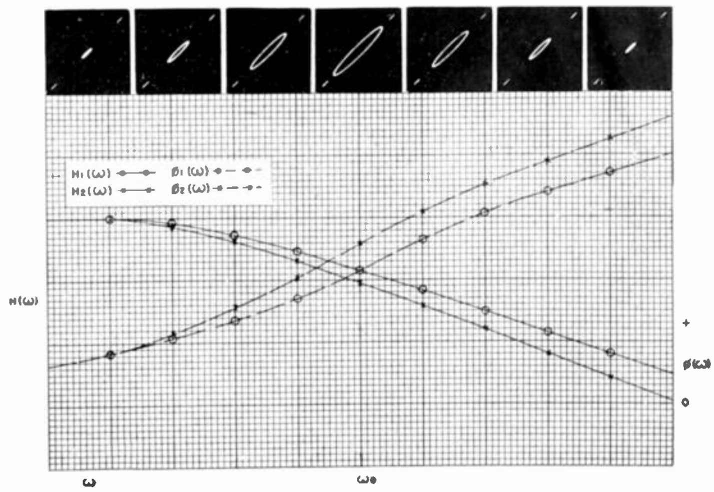

Fig. 5—Response curves of the amplitude and phase functions of the nonresonant sections, and corresponding photographs of the oscilloscope screen when $\tau_2 \neq \tau_1$.

The phase displacements and amplitude responses of the nonresonant sections when $\tau_2 \neq \tau_1$ are shown in Fig. 5. The amplitude difference introduced in such sections may be readily normalized by readjusting the gain of one channel. The relative phase displacement persisting should result in a figure of constant ellipticity whose slope remains fixed as the signal-generator frequency departs from resonance in either direction.

[4] The direction in which the major axis of the elliptical figure changes here depends only on the connections to the oscilloscope channels. Referring to Fig. 1, it is observed that channel 2, which is of the higher selectivity, is attached to the horizontal channel. Therefore, the slope becomes more nearly vertical for either direction of change of the signal generator.

Summary and Conclusion

Upon normalizing the network gain characteristics at the frequency of maximum response, the practical results of the foregoing section may be summarized as follows:

(1) A change in slope of the major axis of the elliptical figure, which is predominantly in one direction regardless of the direction in which the frequency of the signal generator departs from resonance, discloses a difference in selectivity between the corresponding resonant sections of the two networks. This may be corrected by any convenient adjustment, such as the introduction of losses into the resonant section of higher Q, until the changes in slope are of equal magnitude on each side of the unit slope position.

(2) A change in slope of the major axis of the elliptical figure which is of equal magnitude on each side of the unit slope position, but whose direction of change depends on the direction in which the frequency of the signal generator departs from resonance, discloses a difference in resonant frequencies between the corresponding resonant sections of the two networks. This may be corrected by changing the tuning adjustment of one channel until the slope of the major axis of the elliptical figure remains fixed as the frequency of the signal generator is changed in either direction.

(3) No appreciable change in slope of the major axis of the elliptical figure, but only constant ellipticity, discloses a constant phase difference between the two networks. This residual phase difference, which is the contribution of the nonresonant sections, may be removed by changing the time constant of the reactive network of the nonresonant sections in either channel by any convenient means, such as a variable capacitor, until the elliptical figure degenerates into a closed line. While this latter adjustment may be made at any fixed frequency, it is nevertheless advisable to sweep the frequency of the signal generator throughout the entire transmission range of the networks while making the adjustment, to make certain that the angular position of the figure remains fixed at 45° and does not develop appreciable ellipticity.

Since the amplitude and phase characteristics of physical networks are not independent but are closely related, it is found necessary to reset the amplitude controls after making each phase adjustment, so as to maintain the angular position of the elliptical figure at 45°. In addition, it should be recognized that, during the early phases of the adjustment process, the network conditions illustrated by Figs. 3, 4, and 5 may exist simultaneously. Efforts should then be concentrated on the removal of the predominating variations individually, until the adjustments are completed.

This method of transient-response equalization is particularly applicable to networks displaying comparatively narrow transmission ranges. For wide-band amplifiers or filter networks, any adjustments made well within the transmission band will hold over a wide frequency range. The remaining, and more critical, adjustment may then be carried out over those frequency regions where rapid amplitude and phase changes occur.

Although it may be possible to equalize the transient response characteristics of any two networks, it is nevertheless important to recognize the doubtful utility of the simple technique here described when networks displaying a large number of degrees of freedom are involved.

Contributors to Waves and Electrons Section

C. E. Barthel, Jr., was born in Gonzales, La., on July 13, 1911. He received the B.S. degree and the M.S. degree in physics from the Louisiana State University in 1932 and 1933, respectively, with highest academic honors. After serving as an instructor of physics at the Washington and Lee University, in Lexington, Va., for five years, he enrolled at the Iowa State College to pursue advanced graduate work, where he conducted research in the general field of photographic sensitometry, and received the Ph.D. degree in physics in July, 1940. The following year he conducted further research in photographic sensitometry and served as an instructor of physics at Iowa State College.

C. E. Barthel, Jr.

In July, 1941, Dr. Barthel joined the staff of the Naval Ordnance Laboratory in Washington, D. C. His original assignments were directed toward the solution of the Navy's critical degaussing problem, the protection of naval vessels against the threat of the magnetic mine; and within a few months he was placed in charge of the Magnetic Model Section of the Laboratory. In early 1944 Dr. Barthel was appointed personnel advisor to the officer-in-charge to assist in the development of personnel policies and personnel organization for the Laboratory. He was later appointed director of personnel, and served in that capacity until May, 1946. At the conclusion of the war, Dr. Barthel was awarded the Distinguished Civilian Service Award for his contributions to the degaussing program and to the development of the organization of the Naval Ordnance Laboratory.

Upon returning to scientific work, Dr. Barthel served as chief of the Electronic Computing and Control Division, and, subsequently, as acting chief of the Acoustics Division of the Naval Ordnance Laboratory.

In October, 1947, Dr. Barthel accepted the position of assistant chairman of physics research at the Armour Research Foundation, and in April, 1948, he was ap-

Contributors to Waves and Electrons Section

pointed chairman of physics research of the same organization. Dr. Barthel is a member of the American Physical Society, the American Association for the Advancement of Science, the American Association for Engineering Education, the Optical Society of America, and the Physics Club of Chicago, among other professional and honorary groups.

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Marvin Camras (S'41–A'42–SM'48) was born in Chicago, Ill., in 1916. He received the B.Sc. degree in electrical engineering from the Armour Institute of Technology in 1940, and the M.Sc. degree from the Illinois Institute of Technology in 1942. Since 1940, as a member of the staff of the Armour Research Foundation, he has done research on a variety of projects in the electronics department, including remote control, high-speed photography, magnetronstriction oscillators, and static electricity.

Marvin Camras

Mr. Camras has been active in the field of magnetic recording. Before the war he developed the Model-50 recorder, which was later used widely by the armed forces. He has worked on multichannel tape recording, on magnetic sound for motion pictures, and on binaural sound reproduction. He is a member of the Acoustical Society of America, the American Association for the Advancement of Science, the American Institute of Electrical Engineers, the Society of Motion Picture Engineers, Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

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Richard B. Dittmar was born in Cincinnati, Ohio, on March 25, 1914. He was graduated from Occidental College, Los Angeles, Calif., in 1936, with the B.A. degree in physics. Following his graduation, he joined the Metropolitan Water District of Southern California as a junior engineer, concerned with the inspection and construction of steel pipe and fabricated structural steel.

R. B. Dittmar

Continuing in the same line of work, in 1937 he joined the Western Pipe and Steel Company as a structural engineer. In September, 1938, Mr. Dittmar left that organization to become assistant chief engineer of Marshall Field and Company, Chicago, Ill., where he was responsible for the maintenance and operation of that company's buildings and facilities.

Shortly after the entry of the United States into World War II, Mr. Dittmar joined the staff of the Naval Ordnance Laboratory, and was assigned to the Mine Unit, where he served as procurement coordinator for that organization. In the spring of 1944, Mr. Dittmar was appointed vice-chairman of the Laboratory's Building Committee, and was concerned with the planning and construction of the new Naval Ordnance Laboratory plant at White Oak, Md. Later he was appointed plant engineer of the White Oak Laboratory. Mr. Dittmar, at present, is chief of the department of engineering of the Los Alamos Scientific Laboratory, Los Alamos, N. Mex., supervising the construction of permanent facilities for that organization.

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William J. Kessler (S'43–A'45) was born on February 28, 1917, at Roebling, N. J. From 1936 to 1942 he was engaged in the distribution and maintenance of broadcast receivers, communication and public address equipment. In 1943 he completed the ESMWT pre-radar and communications course conducted by the University of Florida, and was retained as instructor until the termination of the training program in 1944.

William J. Kessler

In 1944 Mr. Kessler joined the staff of the Engineering and Industrial Experiment Station of the University's College of Engineering to participate in the development of the AN/GRD-1A atmospherics direction finder for the Army Signal Corps and the proximity fuze under an OSRD contract. Mr. Kessler is the recipient of a certificate of recognition from the Office of Scientific Research and Development, and the Naval Ordnance Development Award.

Since 1946, Mr. Kessler has been engaged as project engineer in charge of the Experiment Station's research project relating to thunderstorm location by means of associated atmospherics.

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Allen H. Schooley (A'35–SM'47) was born in Terril, Iowa on December 16, 1909. He received the B.S. in electrical engineering from Iowa State College in 1931. In 1932 he received the M.S. degree in electrical engineering from Purdue University. Between 1932 and 1936, he did radio servicing, was a computer for the United States Coast and Geodetic Survey, and spent a year at the State University of Iowa doing graduate work in engineering and physics under a scholarship.

A. H. Schooley

From 1936 to 1940 Mr. Schooley was an engineer in the Advanced Development Section, Radiotron Division of the Radio Corporation of America, Harrison, N. J. While at RCA he designed and built the first miniature radio tube envelopes that are now extensively used in civilian and military electronic equipment.

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For a photograph and biography of John A. Victoreen, see page 211 of the February, 1949, issue of the Proceedings of the I.R.E

Abstracts and References

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ACOUSTICS AND AUDIO FREQUENCIES

016:534 602
References to Contemporary Papers on Acoustics—A. Taber Jones. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 712–719 and 882–887; September and November, 1948.) Continuation of 3286 of 1948.

531.49 603
Electromechanical Feedback—J. de Boer and G. Schenkel. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 641–647; September, 1948.) Describes how a mechanical impedance can be realized electrically.

534 604
Sound Transmission and Noise—A. J. King. (*Nature* (London), vol. 162, pp. 499–501; September 25, 1948.) Brief details of the papers read at a London Symposium held by the Acoustics Group of the Physical Society.

534.21 605
Interactions between a Plate and a Sound Field—R. D. Fay. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 620–625; September, 1948.) Discussion of interactions between flexural vibrations in the plate and the sound field in an ambient fluid medium. See also 611 below.

534.21 606
Radiation from a Diaphragm Struck Periodically by a Light Mass—M. Strasberg. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 683–690; September, 1948.) The radiation consists of a line spectrum with intensity maxima near the resonant frequencies of the diaphragm. The energy in the spectrum lies mainly below a frequency which is twice the reciprocal of the time of contact of the hammer at each impact. Expressions are developed for the efficiency of the system.

534.213 607
The Coupling of a Cylindrical Tube to a Half-Infinite Space—J. W. Miles. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 652–664; September, 1948.) "The radiation resistance of a semi-infinite space, as seen by a cylindrical tube through an aperture in an infinite plane, is computed." An asymptotic expression for the field at a distance from the aperture is developed, and reflection and transmission coefficients are determined in terms of the radiation admittance. Numerical results are shown graphically for the special case of a circular tube and an infinite slot; these results are compared with those obtained by Rayleigh.

534.213.4 608
On the Radiation of Sound into a Circular Tube, with an Application to Resonators—U. Ingård. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 665–682; September, 1948.) See 1 of February.

534.23:534.321.9 609
The Scattering of Ultrasonic Waves in Water by Cylindrical Obstacles—L. Bauer, P. Tamarkin, and R. B. Lindsay. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 858–868; November, 1948.)

534.231.3 610
Acoustic Impedance Matching by Means of Screens—F. H. Slaymaker and M. E. Hawley. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 802–807; November, 1948.) For operation in air, transmission increases of 9 db have been obtained by correctly placing suitable screens in front of a crystal transducer. The response curve of a sharply resonant transducer is only slightly modified; its directional characteristics are hardly affected. The results of experiments with woven and perforated screens of different shapes are discussed.

534.24 611
Reflection of Sound from Submerged Plates—W. J. Finney. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 626–637; September, 1948.) When high-frequency underwater sound strikes a submerged steel plate obliquely, "nonspecular" reflection occurs. The behavior of this reflection as a function of frequency, plate dimensions, and angle of incidence is studied experimentally. See also 605 above.

534.321.9:534.373 612
Absorption Measurements of Sound in Sea Water—G. J. Thiessen, J. R. Leslie, and F. W. Simpson. (*Canad. Jour. Res.*, vol. 26, no. 5, pp. 306–312; September, 1948.) The measurements were made with a diverging beam, at frequencies between 0.35 and 2.3 Mc. Absorption values are somewhat lower than Richardson's values (3474 of 1940) for fresh water, but conclusive comparison is not possible. Advantages of using long distances are discussed.

534.373 613
The Origin of Sound Absorption in Water and in Sea Water—L. N. Liebermann. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 868–873; November, 1948.) At frequencies below 100 kc the absorption in sea water is about 100 times that in pure water, whereas above 1 Mc the absorptions are similar. A possible explanation is given.

534.373 614
The Attenuation of Audible Sound in Fog and Smoke—V. O. Knudsen, J. V. Wilson, and N. S. Anderson. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 849–857; November, 1948.) Measurements in artificial fogs or smoke are in approximate agreement with the theories of Sewell, Epstein, and Oswatitsch.

534.373:534.321.9:546.212 615
Ultrasonic Absorption in Water in the Temperature Range 0°–80°C.—Smith and Beyer. (*See* 686.)

534.43:621.395.61 616
Electron Tube Phonograph Pickup—H. F. Olson and J. Preston. (*Audio Eng.*, vol. 32, pp. 17–20; August, 1948.) Description of a pickup in which the stylus is coupled to a tube plate. By using a conical instead of a cylindrical-rod plate, the effective mass and mechanical impedance are very greatly reduced. See also 2624 of 1947 (J. V.).

534.6 617
A Comparison of the Rayleigh Disk and the Acoustic Radiometer Methods for the Measurement of Sound-Wave Energy—J. Hartmann and T. Mortensen. (*Phil. Mag.*, vol. 39, pp. 377–394; May, 1948.) The two methods were compared with as high a degree of precision as possible. The results show the methods to be consistent if the constant in the formula for the pressure of sound waves incident normally on a wall has the value 2, a value which has been derived independently by various authors. The value $(1+\kappa)$ in Rayleigh's second formula thus appears to be incorrect.

534.6 618
Reciprocity Calibration of Vibration Probes—C. T. Morrow. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 826–829; November, 1948.)

534.612.4 619
Reciprocity Calibration of Primary Vibration Standards [transducers]—S. P. Thompson. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 637–640; September, 1948.) The reciprocity method for absolute calibration is shown to be reliable and accurate up to at least 700 cps. The design of a set of primary standards is discussed.

534.612.4 620
Reciprocity Free Field Calibration of Microphones to 100 kc/s in Air—I. Rudnick and M. N. Stein. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 818–825; November, 1948.)

534.75 621
The Effect of High Altitude on the Threshold of Hearing—H. W. Rudmose, K. C. Clark, F. D. Carlson, J. C. Eisenstein, and R. A. Walker. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 766–770; November, 1948.) Measurements at a simulated altitude of 35,000 feet, after correction for changes in the response of the apparatus, show that the shift in the average threshold of hearing is within ±2.5 db of sound-pressure level when the density of the air in the outer and middle ear is decreased to about a quarter of its sea-level value.

534.78 622
The Effects of High Altitude on Speech—K. C. Clark, H. W. Rudmose, J. C. Eisenstein, F. D. Carlson, and R. A. Walker. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 776–786; November, 1948.) Measurements were made at simulated altitudes up to 40,000 feet. Vowels and semivowels show a loss in mean square pressure with altitude roughly proportional to the logarithm of the density ratio, while some consonants are little affected.

534.78 623
The Masking of Tones by Repeated Bursts of Noise—G. A. Miller and W. R. Garner. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 691–696. September, 1948.) See also 2138 of 1948 (Miller and Taylor).

534.851 624
The Light-Pattern Meter—R. E. Santo. (Proc. I. R. E., vol. 36, pp. 1431–1433; November, 1948.) A meter which determines, to within 0.5 db, the amplitudes of sine waves recorded on disks. See also 1985 of 1947 (Hornbostel).

534.851:621.395.813 625
Simplified Dynamic Noise Suppressor—C. G. McProud. (*Audio Eng.*, vol. 32, pp. 22–24, 33; August, 1948.) A single reactance tube is connected directly across the magnetic pickup. It provides adequate low-frequency equalization to correct for average recording characteristics. For the original dynamic noise suppressor see 991 of 1947 (Scott) and 932 of 1948 (Scott.)

534.88 626
Sonic Navigation System—S. R. Rich and A. H. Rosen. (*Electronics*, vol. 21, pp. 92–97; November, 1948.) A hyperbolic navigation system for harbors and channels using underwater af pulses from pairs of transmitters. The correct course can be followed to within 50 yards by audio methods, or more accurately by using meter equipment.

621.395.61/.62 627
A 100 kc Underwater Magnetostrictive Transducer—L. Camp, R. Vincent, and F. du Breuil. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 611–615; September, 1948.) A diaphragm is driven by a set of 32 magnetostriction units excited in phase and with amplitudes graded so as to give a half-amplitude width for the main beam of 20°, with secondary lobes 30 db down. The laminations used are of 2 volts permendur, magnetized by a half-cycle surge current through the windings. For remanence operation, an efficiency of 50 per cent is possible with a maximum driving power of 40 watts. See also 628 below (Camp).

621.395.61/.62 628
Lamination Designs for Magnetostrictive Underwater Electroacoustic Transducers—L. Camp. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 616–619; September, 1948.) A theory of lamination design for transducers having a plane piston-like radiating face. Experimental data show how well the theory predicts operating characteristics. See also 627 above.

621.395.61/.62 629
Note on the Impedance Variations of an Electro-Acoustic Transducer in a Reflecting Field—S. Byard. (*Proc. Phys. Soc.*, vol. 61, pp. 478–480; November 1, 1948.)

621.395.61 630
Phase Characteristics of Condenser Microphones—F. M. Wiener. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 707; September, 1948.)

621.395.623.8 631
New Theater [cinema] Loudspeaker System—H. F. Hopkins and C. R. Keith. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 385–398; October, 1948.) A general discussion of the requirements for good reproduction and a description of a double system having a cross-over frequency of 800 cps and using sectoral horns to give a wide angle of coverage at high frequencies.

621.395.625 632
The Recording and Reproduction of Sound. Parts 18–22—O. Read: O. Read and R. Endall. (*Radio News*, vol. 40, pp. 49–51, 146; 48–49; 56–57; 89; 50–51, 158; and 48–50, 124; August to December, 1948.) Part 18: Factors affecting reproducers at af. Part 19: Design data for series and parallel filter networks and constant-resistance networks. Part 20: *RC* tone-control systems. Part 21: af correction. Part 22: Loudspeaker cabinets and baffles.

621.395.625 633
35-mm Magnetic Recording System—E. Masterson. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 481–488; November, 1948.) Discussion of the design and performance of conversion apparatus to adapt a well-known 35-mm sound recorder for magnetic recording.

621.395.625(083.74):621.317.79:621.395.813 634
Proposed Standards for the Measurement of Distortion in Sound Recording—(*See* 791).

621.395.625.3 635
Optimum High-Frequency Bias in Magnetic Recording—G. L. Dimmick and S. W. Johnson. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 489–499; November, 1948. Discussion, pp. 499–500.) "An experimental study was made of magnetic tapes and films produced by several manufacturers. The effects of bias current upon the frequency characteristic, the reproducing level, and the harmonic distortion are shown."

621.395.625.3 636
Magnetic Field Distribution of a Ring Recording Head—S. J. Begun. (*Audio Eng.* vol. 32, pp. 11–13, 39; December, 1948.) The components of the magnetic field acting on the recording medium are determined graphically; accuracy is adequate.

ANTENNAS AND TRANSMISSION LINES

621.315 637
High-Frequency Polyphase Transmission Line—C. T. Tai. (Proc. I. R. E., vol. 36, pp. 1370–1375; November, 1948.) Formulas for the characteristic impedance of the polyphase transmission line and single-phase multiwire line are found by the vector potential method.

621.392.26† 638
Analysis and Performance of Waveguide-Hybrid Rings for Microwaves—H. T. Budenbom. (*Bell Sys. Tech. Jour.*, vol. 27, pp. 473–486; July, 1948.) The rings are considered as re-entrant transmission lines, which are transformed into equivalent tee- or lattice-network sections. Determinantal methods of analysis are used. Experimental results, obtained from a carefully constructed sample of each of two specific types, agree satisfactorily with the theory. Another account noted in 2443 of 1948.

621.392.26† 639
Theory of Slots in Rectangular Wave-Guides—A. F. Stevenson. (*Jour. Appl. Phys.*, vol. 19, pp. 24–38; January, 1948.) Equations are developed for the field generated in a waveguide of arbitrary section by an assigned tangential electric field in the wall of the waveguide. The analogy with a transmission line is established, detailed formulas being given for the reflection and transmission coefficients and for the voltage amplitude generated in the slot by a given incident wave. The transmission coefficients can, in part, be calculated simply from energy considerations, and expressions are derived for the conductance of a slot when it is equivalent to a series or shunt element in a transmission line. Guide-to-guide coupling is considered and equations are developed for the voltage amplitudes in the various slots of an array.

621.392.26† 640
Scattering of Electromagnetic Radiation by a Thin Circular Ring in a Circular Wave Guide—P. Feuer and E. S. Akeley. (*Jour. Appl. Phys.* vol. 19, pp. 39–47; January, 1948.) The waveguide walls and the ring are assumed to be perfectly conducting. Approximate formulas are obtained for the $TE_{1,1}$ wave which give the scattering cross section, resonance maximum, and half-width as a function of the width of the ring.

621.392.26† 641
Remarks on Slow Waves in Cylindrical Guides—A. A. Oliner. (*Jour. Appl. Phys.*, vol. 19, pp. 109–110; January, 1948.) A letter indicating (a) a simpler method than that of Pincherle (1 of 1945) or Frankel (22 of 1948) for determining the values of dielectric constants and radii necessary for a given phase velocity, (b) the existence of curves which greatly reduce the work of calculation. See also 334 of 1948 (Bruck and Wicher).

621.392.52:621.396.662.3 642
Low-Pass Filters Using Coaxial Transmission Lines as Elements—Mode. (*See* 680.)

621.396.67 643
Characteristics of Helical Antennas Radiating in the Axial Mode—J. D. Kraus and J. C. Williamson. (*Jour. Appl. Phys.*, vol. 19, pp. 87–96; January, 1948.) A theoretical and experimental investigation of the radiating modes when the helix diameter is 0.2λ to 0.5λ and the pitch as high as 0.5λ. In the axial mode, the helix behaves as a beam antenna and the radiation is nearly circularly polarized; for a given helix, this mode can persist over a considerable frequency range. See also 650 below.

621.396.67 644
The Conical Dipole of Wide Angle—P. D. P. Smith. (*Jour. Appl. Phys.*, vol. 19, pp. 11–23; January, 1948.) A method of calculating the admittance of dipoles consisting of complete cones whose semi-angles lie between 0° and 90°. The theory uses the orthogonal properties of Legendre functions and their derivatives to make the tangential component of the outside field vanish over the spherical end surfaces of the dipoles and to make the inside and outside fields fit at the boundary sphere. An approximate formula is also developed which agrees reasonably even with that of Schelkunoff for the impedance of a thin conical dipole and with that of Stratton and Chu for the impedance of a spherical radiator.

621.396.67 645
Antenna Design for Television and F.M. Reception—F. A. Kolster. (Proc. I. R. E.,

vol. 36, p. 1363; November, 1948.) Correction to formula in 305 of March.

621.396.67:621.392.26† 646
The Field Surrounding an Antenna in a Waveguide—J. S. Gooden. (*Jour. IEE* (London), part III, vol. 95, pp. 346–350; September, 1948. Summary, *ibid.*, part I, vol. 95, p. 454; October, 1948.) An approximate method is described for obtaining the maximum electric field strength surrounding such an antenna. Formulas are given for the circular guide; results for rectangular waveguides are shown graphically.

621.396.67:621.396.97:621.396.812.3 647
An Antenna for Controlling the Nonfading Range of Broadcasting Stations—C. L. Jeffers. (Proc. I. R. E., vol. 36, pp. 1426–1431; November, 1948.) The antenna consists of two vertical elements, one at ground level and the other directly above it. By altering the ratio of the currents in the two sections, the angle above which the radiated energy is a minimum can be varied from 40° to 60°. Theoretical performance was checked by measurements on a scale model.

621.396.671 648
Mutual Impedance of Parallel Aerials—G. Barzilai. (*Wireless Eng.*, vol. 25, pp. 343–352; November, 1948.) Formulas are given for two vertical antennas of different lengths terminated at a perfectly conducting plane. Sinusoidal current distribution is assumed. Results obtained from the formulas are compared with those obtained by a graphical method of integration. Graphs illustrate the behavior of a driven antenna with a parasitic element of various lengths. See also 658 of 1948 (Cox.)

621.396.671 649
Rhombic Aerial Design Chart—R. H. Barker. (*Wireless Eng.*, vol. 25, pp. 361–369; November, 1948.) The equation for the angle of elevation at which the gain of the horizontal rhombic antenna is a maximum is solved by a graphical method for the vertical plane containing the major axis. The effects of variation of antenna dimensions, frequency, and height on the relative gain are calculated. The small correction due to the finite conductivity and dielectric constant of the earth is discussed.

621.396.677 650
Measured Impedances of Helical Beam Antennas—O. J. Glasser and J. D. Kraus. (*Jour. Appl. Phys.*, vol. 19, pp. 193–197; February, 1948.) The results of measurements at frequencies of 300 to 500 Mc are analyzed. The antennas are suitable for wide-band applications. See also 3033 of 1947 (Kraus) and 643 above.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.3:621.392 651
Simple Relations for Calculating Certain Transient Responses—W. J. Cunningham. (*Jour. Appl. Phys.*, vol. 19, pp. 251–256; March, 1948.) The response of a linear transmission system to a step voltage or an impulse is related directly to its steady-state response to a sinusoidal signal of variable frequency. Empirical equations connecting the two responses are given and discussed; they may be expected to be accurate within 25 per cent and are useful in preliminary design work or for checking exact but tedious calculations.

621.3.015.3:621.392 652
The Exact Solution for the Compensation of Transient Distortion in Networks—D. C. Espley. (*Onde Élec.*, vol. 28, pp. 461–462; December, 1948.) Summary only. Brief discussion of a method based on equivalent circuits.

621.3.015.3:621.392 653
The Transient Response of Damped Linear Networks with Particular Regard to Wideband Amplifiers—W. C. Elmore. (*Jour. Appl. Phys.*, vol. 19, pp. 55–63; January, 1948.) When the response to an applied unit step function consists of a monotonic rise to a final constant value, the delay time and rise time are defined in such a way that they may be simply computed from the Laplace system function of the network. The method is applied to low-pass multistage wide-band amplifiers.

621.3.094 654
Attenuation and Phase Distortion and their Influence on the Establishment of Television Signals—G. Fuchs and V. Baranov. (*Onde Élec.*, vol. 28, pp. 463–466; December, 1948.) Summary only. Formulas for the general solution of problems relating to the transient behavior of a transmission system are applied first to the case of transmission through a section of coaxial cable and then to a complete transmission system.

621.314.26:621.313.3 655
Parallel Operation of Aircraft Alternators Using Electronic Frequency Changers—O. E. Bowlus and P. T. Nims. (*Trans. AIEE*, vol. 66, pp. 31–38; 1947.) Preliminary development work. See also 2178 of 1948.

621.314.3† 656
Some Fundamentals of a Theory of the Transductor or Magnetic Amplifier—A. U. Lamm. (*Trans. AIEE*, vol. 66, pp. 1078–1085; 1947.) The combination of the dc presaturated reactor with the metal rectifier is studied assuming an idealized magnetization curve. Two typical connections are chosen as examples. Dynamic response and applications to voltage and current regulation are considered. A few German and Swedish references are given.

621.314.6 657
The Constancy of Small Rectifiers—D. G. Tucker and G. F. Machen. (*Jour. Sci. Instr.*, vol. 25, pp. 369–371; November, 1948.) The backward resistance of various types of rectifier is much more variable than the forward resistance, either from one rectifier to another of the same type or as regards change with temperature.

621.318.4.028.4 658
Data on the High-Frequency Resistance of Coils—W. F. Witzig. (*Trans. AIEE*, vol. 66, pp. 764–769; 1947.) From measured resistance values for several coils, an approximate method of determining the resistance of coils wound with flat strip or tubing is derived. Factors affecting resistance are discussed.

621.318.572 659
A Tripping Circuit for a Multi-Stage Surge Generator—E. L. White. (*Jour. Sci. Instr.* vol. 25, pp. 307–309; September, 1948.) A low-voltage impulse is applied to the grid of a thyratron and simultaneously to the sweep circuit of a cro. The thyratron transmits a delayed impulse to an extra sphere in the center of the first sphere gap of the generator, so that breakdown occurs at any desired instant.

621.318.572:539.16.08 660
Electronic Counters for Impulses—P. Naslin and A. Peuteman. (*Rev. Gén. Élec.*, vol. 57, pp. 417–431; October, 1948.) A decade system of thyratrons is described and many developments from the simple flip-flop circuit are discussed, including the Eccles-Jordan circuit, various binary decade systems and a ring decade system using pentodes. A generalized flip-flop circuit with 5 stable states uses 5 triodes with suitable interconnections. Applications of counters are outlined briefly.

621.319.4:621.315.614:621.315.59 661
The Dielectric Properties of Cellulose Insulation Impregnated with Semiconducting Liquids—Clark. (*See* 747.)

621.319.4:621.315.614.015.5 662
The Probable Breakdown Voltage of Paper Dielectric Capacitors—Brooks. (*See* 748.)

621.319.4(43) 663
German Radio Capacitors—S. J. Borgars. (*Electronic Eng.* (London), vol. 20, pp. 355–357; November, 1948.) A review of the characteristics and construction of capacitors with paper, ceramic, synthetic-mica, polystyrene, electrolyte, or air dielectric. Ceramics are extensively used, with glass or glazed ceramics for terminal seals and chlorinated naphthalene as an impregnant. A 4-gang variable capacitor uses die-cast Mg alloy for the main metal parts, with annular peripheral grooves in the rotor vanes for trimming purposes. In one type of variable capacitor, temperature changes cause equal axial displacements of the rotors and stators, so that the temperature coefficient is low. See also B.I.O.S. final reports nos. 226, 563, 567, 893, and 1459.

621.392:003.62 664
Circuit Symbols—L. H. Bainbridge-Bell. (*Wireless World*, vol. 54, pp. 437–438; December, 1948.) A general review of the British Standards Institution publication BS530: "Graphical Symbols for Telecommunications". Arguments in favor of retaining a large number of symbols are put forward. Nine new symbols not appearing in previous editions are explained. See also 2737 of 1948.

621.392.43 665
Wide-Band Matching by Means of Several Intermediate Elements—H. Aberdam. (*Onde Élec.*, vol. 28, pp. 474–481; December, 1948.) Mathematical analysis shows that matching by means of two intermediate transformers, such as $\lambda/4$ elements, diminishes very considerably the energy losses by reflection in a given frequency band. In a particular example considered, this reduction is about 90 per cent for an octave band of frequencies. This problem has been studied in Germany by O. Zinke (Grundlagen der Breitbandantennenanlagen) and W. Pauls (Berechnung und Aufbau von Breitbandleistungs-transformatoren.)

621.392.52 666
Network Analysis Involving Realizable Filter Functions—D. K. C. MacDonald. (*Phil. Mag.*, vol. 38, pp. 115–131; February, 1947.) The evaluation of the integrals obtained from the analysis of the response of a network to a given stimulus is difficult. The problem of the low-pass filter is discussed, and a family of "physically realizable" filter functions is obtained. Various combinations of this family will give approximations to practical filter functions.

621.396.611:537.291 667
On the Theory of Electron-Beam H. F. Oscillators—G. Ya. Myakishev. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1063–1068; August, 1948. In Russian.) An electron beam is subjected to a disturbance between two grids and the propagation of the resulting modulation of the charge density, current, and kinetic energy along the beam is investigated mathematically. Discussion of: (a) propagation of modulation with and without allowance for the interaction of electrons, and (b) the case of a beam of finite length, is based on an equation (proposed by Vlasov) determining the distribution of electrons.

621.396.611:621.316.729 668
Synchronization of Controlled Relaxation Oscillators—O. I. Butler. (*Phil. Mag.*, vol. 39, pp. 518–528; July, 1948.) Harker (3879 of 1938) has shown that synchronization can be realized for the linear-timebase type of oscillator. The phenomena of synchronization are

here investigated quantitatively. A method of synchronization materially different from Harker's is shown to be also practicable. A measure of the rigidity of the "synchronous lock" is obtained which allows the effect of changing conditions of operation to be more definitely assessed.

621.396.611:621.396.619.13 669

Frequency Modulation of Variable Frequency Oscillators—N. F. Vollerner. (*Radiotekhnika* (Moscow), vol. 3, pp. 47–55; July and August, 1948. In Russian.) "Howling" *RC* oscillators are considered and a brief analysis of the operation of a typical circuit is given, with experimental verification.

621.396.611.1:621.317.6 670

The Response of a Resonant System to a Gliding Tone—N. F. Barber and F. Ursell. (*Phil. Mag.*, vol. 39, pp. 345–361; May, 1948.). The response of an oscillatory system to a tone whose frequency slowly increases or decreases is discussed. The variation of amplitude near resonance depends on a single parameter θ. Such an oscillatory system can be used as a frequency analyzer (203 of 1947) whose resolving power is highest if θ has a certain value. Similar results are obtained when a tone of fixed frequency acts upon a resonant system whose natural frequency is slowly changed. See also 671 below.

621.396.611.1:621.317.6 671

Response of Linear Resonant Systems to Excitation of a Frequency Varying Linearly with Time—G. Hok. (*Jour. Appl. Phys.*, vol. 19, pp. 242–250; March, 1948.) "A general solution of this problem is obtained by means of Laplacian transforms. The resulting complex function is evaluated numerically, and universal response curves are presented in order to facilitate the application of the solution to simple or complicated electrical networks as well as to other resonant systems represented by equivalent networks." See also 670 above.

621.396.611.3 672

Parabolic Loci for Two Tuned Coupled Circuits—S. Chang. (Proc. I. R. E., vol. 36, pp. 1384–1388; November, 1948.) Under certain restrictions, the reciprocal of the response function, or its equivalent, leads to parabolic loci in the complex plane. Design methods for coupled circuits are based on the geometry of the parabola.

621.396.611.3 673

Coupled Circuits for High and Medium Frequencies—L. de Valroger. (*Rev. Tech. Comp.* (Franç) pp. 17–45; September, 1948. In French, with English Summary.) A general system of curves is developed which greatly facilitates the determination of the selectivity and phase distortion for a system of coupled circuits with any type of coupling and impedances of any value. The curves can be used for all cases where the pass band is not excessively large. One such system of curves is particularly useful for the coupled circuits of receivers. Large errors prohibit the use of the curves for the case of very large bandwidths; for this a new method of calculation is given, which gives fairly quickly and without approximations, the values of the various impedances of the circuits.

621.396.645 674

Fundamental Relations for Transmitter Amplifiers with Wide-Band H.F. Modulation and Using Ordinary Valves—J. Fagot. (*Onde Élec.*, vol. 28, pp. 376–378; October, 1948.) Summary only. Formulas for output power and power gain per stage are derived and applied to obtain improved performance of transmitter amplifiers by modification of either the output stage or the intermediate circuits.

621.396.645 675

On the Equivalence of H.F. and L.F. Amplifiers—S. I. Evtyanov. (*Radiotekhnika*, (Moscow), vol. 3, pp. 26–33; July and August, 1948. In Russian.) Tuned linear amplifiers are considered. The relationship between the instantaneous values of input and output voltage is compared with that between the complex envelopes of these quantities for a high-frequency amplifier, taking into account the frequency displacement of the input voltage from the resonance frequency of the amplifier. The condition for a low-frequency amplifier circuit to be equivalent to a high-frequency amplifier circuit is deduced. Equivalent circuits and transmission coefficients are tabulated for 3 different high-frequency circuits. Transient phenomena are discussed.

621.396.645:518.4 676

Graphical Analysis of Cathode-Coupled Amplifiers—H. A. Watson. (*Canad. Jour. Res.*, vol. 26, no. 8, pp. 340–346; August, 1948.) A method of design and gain calculation based on data obtained from the plate characteristic curves. The method can be used to predict performance and to determine the zero-signal operating conditions.

621.396.645:621.396.828.1 677

Heater Supplies for Amplifier Hum Reduction—F. W. Smith. (*Audio Eng.* vol. 32, pp. 26–27, 35; August, 1948.) The best solution of the hum problem is to use either dc or high-frequency ac for the tube heaters. Typical supply units are described.

621.396.645.36 678

The See-Saw Circuit Again—J. McG. Sowerby. (*Wireless World*, vol. 54, pp. 447–449; December, 1948.) Diagrams with explanations of three such circuits are given, together with a set of design curves for the tee see-saw and an appendix containing useful formulas. Various applications of these circuits are discussed. See also 2212 of 1948 (Cocking.)

621.396.662.3 679

Wide-Band Crystal Filter for Carrier Program Circuits—F. E. Stehlik. (*Bell Lab. Rec.* vol. 26, pp. 462–465; November, 1948.) The exacting requirements to be met by such filters are discussed. The filter consists of two complex lattice sections containing crystal elements and two ladder sections containing coils and capacitors. Each lattice section requires 22 crystal elements. Insertion loss and delay distortion characteristics are shown for a filter passing the lower sideband of an 88-kc carrier.

621.396.662.3:621.392.52 680

Low-Pass Filters Using Coaxial Transmission Lines as Elements—D. E. Mode. (Proc. I. R. E., vol. 36, pp. 1376–1383; November, 1948.) Four transmission-line low-pass filter designs are given which specify the mechanical dimensions required for constructing filters with pass bands as large as 4,000 Mc.

621.396.813:621.396.645 681

On the Reduction of Phase Distortion in Stagger-Tuned Amplifiers—J. Laplume. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 675–677; October 4, 1948.) All even harmonics will be practically annulled when the resonance frequencies of the different circuits are symmetrical with respect to the mean frequency. For the case of 2 tuned circuits, the third harmonic of the mean frequency can also be eliminated if $B/\Delta=2\sqrt{3}$, where $\Delta(=\frac{1}{2}|\eta_2-\eta_1|)$ is the difference between the mean frequency and that of either of the circuits, and $B(=\eta/Q)$ is the pass band at 3 db. The first harmonic not annulled is thus the fifth. With 3 circuits, both the third and fifth harmonics can be annulled. The first two circuits are again symmetrical with respect to the mean frequency, to which the third circuit is tuned. The pass band B_1 for the first two circuits and that (B_2) for the third should be such that $B_1/\Delta=2.097$ and $B_2/\Delta=2.647$. In both cases, the amplitude curve is rather rounded. See also 40 of February.

621.397.645 682

Pentriode Amplifiers—H. M. Zeidler and J. D. Noe. (Proc. I. R. E., vol. 36, pp. 1332–1338; November, 1948.) Description of two video-amplifier circuits in which the phase-shift and degenerative decrease in gain caused by inefficient screen-grid and cathode by-pass circuits are eliminated throughout the frequency range. Small mica capacitors are used where possible instead of bulky electrolytic ones.

GENERAL PHYSICS

53.081+621.3.081 683

Units—(*Bull. Soc. Franç. Élec.*, vol. 8, pp. 557–581; December, 1948.) Full discussion on 1006 of 1948 (Budeanu) and 1007 of 1948 (Grivet).

53.081+621.3.081 684

Electromagnetic Units and Definitions—G. Stedman. (*Wireless World*, vol. 54, pp. 406–409; November, 1948.) Discussion of the reasons underlying the changes in accepted standards noted in 2833 and 2834 of 1947.

530.162:519.2 685

The Restricted Problem of the Random Walk—A. N. Gordon. (*Phil. Mag.*, vol. 39, pp. 572–575; July, 1948.) Comment on 627 of 1944 (Silberstein). A simple solution is proposed which does not involve difference equations. This is then extended to any number of dimensions.

534.373:534.321.9:546.212 686

Ultrasonic Absorption in Water in the Temperature Range 0°–80°C—M. C. Smith and R. T. Beyer. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 608–610; September, 1948.) Experimental values for 6 frequencies between 12.25 and 40.50 Mc agree with Hall's calculated values (2774 of 1948) within the limits of experimental error.

537.212:621.392.029.64 687

The Electrostatic Field of a Point Charge inside a Cylinder, in Connection with Wave-Guide Theory—C. J. Bouwkamp and N. G. de Bruijn. (*Jour. Appl. Phys.*, vol. 19, p. 105; January, 1948.) Corrections to 78 of 1948.

537.291 688

Graphical Methods for Tracing Electron Trajectories—R. Musson-Genon. (*Onde Élec.*, vol. 28, pp. 469–473; December, 1948.) Methods giving results of different orders of approximation are discussed and an accurate method based on Taylor series, including terms of the 4th order, is described, together with a mechanical device with which the necessary determinations can be carried out both quickly and accurately.

537.291 689

Possible Fluctuations in Electron Streams Due to Ions—J. R. Pierce. (*Jour. Appl. Phys.*, vol. 19, pp. 231–236; March, 1948.) Theory predicts that disturbances in an electron stream containing ions will build up in the direction of electron motion. Experiment shows the existence of oscillations which roughly correspond to the theory.

537.525.029.64 690

Electrical Breakdown of a Gas between Coaxial Cylinders at Microwave Frequencies—M. A. Herlin and S. C. Brown. (*Phys. Rev.*, vol. 74, pp. 910–913; October 15, 1948.) Continuation of 3390 of 1948. Experimental values obtained for the ionization rate between parallel plates are applied to the computation of breakdown voltages of air between coaxial cylinders. Graphs of the values of this voltage as a function of air pressure are obtained, both theoretically and experimentally, for a coaxial cavity resonant for λ 9.6 cm. The agreement is sufficient to justify the original postulates.

538.1 691
Methods of Electromagnetic Field Analysis—S. A. Schelkunoff. (*Bell Sys. Tech. Jour.*, vol. 27, pp. 487–509; July, 1948.) A discussion of the fundamental conceptions underlying the application of electromagnetic field theory to practical systems. The points of contact between field and circuit theory are stressed and the properties of free space, transmission lines, antennas, and waveguides are discussed in terms of the field theory. The relationship between Kirchhoff's equations and Maxwell's field equations is indicated, and some differences between a network of lumped elements and a continuous network are explained by means of the complex impedance plane.

538.3.001.572 692
Theory of Models of Electromagnetic Systems—G. Sinclair. (Proc. I.R.E., vol. 36, pp. 1364–1370; November, 1948.) Discussion of the conditions necessary so that model measurements can be made on an absolute instead of a relative basis.

538.569.4 693
Minimum Detectable Absorption in Microwave Spectroscopy and an Analysis of the Stark Modulation Method—W. D. Hershberger. (*Jour. Appl. Phys.*, vol. 19, pp. 411–419; April, 1948.)

538.69 694
On the Influence of a Homogeneous Longitudinal Magnetic Field on Radiation [from Ra-E]—F. Ehrenhaft and R. Herzog. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 626–627; September 27, 1948.) The penetrating power of the radiation from a Ra preparation (mainly Ra-E) on the face of the south pole of an electromagnet was definitely increased by switching on the field. See also 1020 of 1948 (Erhenhaft).

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

521.15:523.7 695
On the Angular Momentum of the Sun—S. Lundquist. (*Ark. Mat. Astr. Fys.*, vol. 35, part 3, section A, 6 pp; September 21, 1948. In English.)

523.72.029.5:621.396.822 696
The Generation of Radio-Frequency Radiation in the Sun—M. Ryle. (*Proc. Roy. Soc. A*, vol. 195, pp. 82–97; November 12, 1948.) Under normal conditions, the radiation has random polarization, and the minimum intensity observed for frequencies below 200 Mc corresponds to a black-body source at about 10^6 °K. When sunspots occur, the intensity increases and corresponds to a source temperature of 10^9 to 10^{10} °K, and the radiation is circularly polarized. The experimental results are considered theoretically, and explained in terms of the acceleration of electrons in the solar atmosphere. Magneto-ionic theory indicates several regions in the solar atmosphere above a sunspot where intense absorption (and therefore, intense radiation) occurs. Radiation at the frequency of free gyration of electrons in the magnetic field is intense, but can only travel toward the center of the sun. A low-altitude high-density region of appreciable absorption coefficient can emit circularly polarized radiation corresponding to that of the ordinary wave of the magneto-ionic theory. A second region, situated at a greater height, absorbs the extraordinary component. Normally, the absorption coefficient in this region is insufficient for the radiation to approach the equilibrium intensity, but elevation of prominence material may greatly increase the intensity of the extraordinary radiation so that it predominates.

523.72.029.63:.64/621.396.822 697
Observation of a Solar Noise Burst at 9500 Mc/s and a Coincident Solar Flare—M. Schulkin, F. T. Haddock, K. M. Decker, C. H. Mayar, and J. P. Hagen. (*Phys. Rev.*, vol. 74, p. 840; October 1, 1948.) Solar noise bursts at 25, 50, 75, 110, 480, and 9,500 Mc were observed on July 29, 1948, simultaneously with the eruption of a very bright solar flare and with the commencement of an ionospheric disturbance. A microwave radiometer with a 10-foot paraboloid antenna, having a beam width of about 0.7°, was used for the observations on 9,500 Mc; the antenna temperature readings on the sun rose from about 2,145° K to over 4,000 °K within 20 seconds. Brief details are given of the personnel and apparatus involved in the other observations.

523.72.029.63:621.396.822:523.752 698
Observation of Remarkable Perturbations of Solar Radiation on Decimetre Waves—J. Houtgast and M. Laffineur. (*Compt. Rend Acad. Sci.*, (Paris), vol. 227, pp. 717–718; October 11, 1948.) The continuous record of solar radiation at λ 54.5 cm obtained with the help of a large (7.5-m) parabolic reflector, showed on September 17, 1948, an increase of intensity up to 3 times the normal mean value. The occurrence of this increase coincided with a chromospheric eruption observed at Greenwich. Short-wave communications were interrupted at the same time. Other perturbations occurred during a period of 1 and one-half hours on October 4, 1948.

523.746 "1749/1948" 699
Tables on Sunspot-Frequency for 1749–1948—E. M. Munro. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 241–246; September, 1948.)

523.746 "1947" 700
Final Relative Sunspot-Numbers for 1947—M. Waldmeier. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 265–267; September, 1948.)

523.746 "1948.01/.06" 701
Provisional Sunspot-Numbers for January to March [and April to June], 1948—M. Waldmeier. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 152 and 268; June and September, 1948.)

523.854:621.396.822.029.62 702
A New Intense Source of Radio-Frequency Radiation in the Constellation of Cassiopeia—M. Ryle and F. G. Smith. (*Nature*, (London), vol. 162, pp. 462–463; September 18, 1948.) Discussion of 80-Mc observations undertaken primarily to determine the polarization of the radiation from Cygnus (noted in 1926 of 1948) which appears to come from very intense discrete sources. Records are shown for the Cygnus source and the new source obtained with the two halves of the antenna system about 0.5 km apart and (a) parallel or (b) mutually perpendicular. As the amplitude of the interference pattern for case (b) is only 5 per cent of that for case (a), the radiation from both sources is randomly polarized. The radiation is possibly more analogous to that from the quiet sun than to that from sunspots.

538.12:521.15 703
Magnetic Field produced by the Rotation of a Gravitational Mass with a Volume Electric Charge—C. Salceanu. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 624–626; September 27, 1948.) From consideration of dimensions, S. Procopiu (704 below) finds that the ratio P/U in Blackett's formula can be expressed as Q/M, where Q is a quantity of electricity, in electromagnetic units, and M is a gravitational mass. This expression is here justified theoretically. From Blackett's value of $1.1+10^{-15}$ for P/U, the value of Q for the earth is found to be $1.32+10^{12}$ electromagnetic units, so that the charge per cm³ has nearly the same numerical value as the surface charge per cm.² The value of the ratio P/U is not of the same order of magnitude in the case of the spinning electron, because the electric charge is so large in comparison with the mass, but by quantizing the value of the magnetic moment of the rotating electron mass, it can be shown that the electron spin is equal to that of Bohr's magneton and the relation $P/U=Q/M$ holds good also for the electron. See also 3112 of 1947 (Blackett).

538.12:521.15 704
Magnetic Field of a Rotating Mass—S. Procopiu. (*Bull. Éc. Polyt.* (Jassy), vol. 3, pp. 453–458; January to June, 1948.) Discussion of a formula which represents the magnetism of the earth or the sun just as well as Blackett's formula; it also holds good for the spinning electron. See also 703 above (Sălceanu).

550.38 705
On the Significance of Geomagnetic Parameters Calculated from Observations in a Limited Area—G. Fanselau. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 163–165; June, 1948.) Summary of paper in *Z. Met.*, vol. 1, nos. 2/3 pp. 55–62; 1946. The analytical representation of geomagnetic observations can be performed either by means of a spherical-harmonic series (for the whole earth) or by a simple Taylor series (for a survey covering a limited area). The coefficients (parameters) of the general expansion are shown to be deducible from those of the Taylor series.

550.38:05 706
Geomagnetic and Geoelectric Literature in Two New German Periodicals—Macht. (*See* 916.)

550.38 "1947" 707
Five International Quiet and Disturbed Days for [each month of] the Year 1947—W. E. Scott. (*Terr. Mag. Atmo. Elec.*, vol. 53, p. 166; June, 1948.)

550.38"1948.01/.06" 708
Cheltenham [Maryland] K-Indices for January to March [and April to June], 1948—R. E. Gebhardt and P. G. Ledig. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 166–167 and 272; June and September, 1948.)

550.38"1948.01/.06" 709
K-Indices and Sudden Commencements, January to March [and April to June], 1948, at Albinger—H. Spencer Jones. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 167–168 and 303–304; June and September, 1948.)

550.384 710
The Abnormal Daily Variation of Horizontal Force at Huancayo and in Uganda—S. Chapman. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 247–250; September, 1948.)

550.384:523.73 711
Persistent Solar Rotation-Period of 26⅞ Days and Solar-Diurnal Variation in Terrestrial Magnetism—J. Olsen. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 123–134; June, 1948.)

550.385/.386 712
Geomagnetic Activity during 1946 Observed at Niemegk—G. Fanselau. (*Terr. Mag. Atmo. Elec.*, vol. 53, p. 162; June, 1948.) Summary of paper in *Z. Met.* vol. 1, no. 15, pp. 449–457; 1947. Daily tables from March to December, 1946. To be published annually in the future.

550.385"1948.01/.06" 713
Principal Magnetic Storms [Jan.–June 1948]—(*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 172–185 and 321–331; June and September, 1948.)

551.510.534 714
The Ozone Content of the Middle Stratosphere—R. Penndorf. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 162–163; June, 1948.) Summary of paper in *Z. Met.*, vol. 1, nos. 10 and 11, pp. 345–357; 1947.

551.510.535 715
Critical Frequency near $\tau=1$—T. L. Eckersley. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 155–161; June, 1948.) The normal and larger critical frequency of the extraordinary ray is in the region where the rays are lost. The extra critical frequency is in the neighborhood of $\tau=eH/(2\pi m\nu)=1$, where ν is the frequency. If τ_z is the vertical component of τ, theoretical reasons are given for expecting the extra critical frequency to occur when $\tau_z=1$. Existing experimental results are not sufficiently accurate to confirm or deny the theory, and further experiments should be undertaken all over the world with adequate power. See also 1282 of 1938 (Martyn and Munro) and 2185 of 1938 (Berkner and Booker).

551.510.535 716
A Study of the Ionospheric Data Obtained at Wuchang—Sept. 1946 through Dec. 1947—P. H. Liang, H. L. Lung, and S. Wang. (*Chin. Jour. Phys.*, vol. 7, pp. 115–131; June, 1948.) Analysis of routine hourly observations. The diurnal and seasonal variations of ionization density and virtual height are shown graphically for the E, F_1, and F_2 layers, and discussed briefly. Deviations from the normally recognized characteristics are discussed more fully. Sporadic-E ionization and its possible connection with meteors are also considered.

551.510.535 717
Magneto-Ionic Measurements at High Latitudes—J. C. W. Scott. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 109–122; June, 1948.) Discussion of measurements of F_2-layer critical frequencies at Clyde River, Baffin Land, only 8° from the geomagnetic pole. Large diurnal and seasonal variations occur in the magnetic field. In addition to these periodic changes, a drop of 20 per cent in apparent field occurred in February, 1946, and persisted till September, 1946. The variation of gyro-frequency with height is greater than would be expected from the inverse-cube law. The results are briefly compared with those obtained at Churchill (Manitoba), at Ottawa, and at College (Alaska).

551.510.535:525.624 718
Tides in the Upper Ionosphere—O. Burkard. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 273–277; September, 1948. In German, with English summary.) Examination of ionospheric critical frequency data for lunar tides has been made by the statistical periodogram method. After eliminating the daily solar variation, a semidiurnal M_2-tide has been detected. Washington data for 1945 gave an average period of 12.423 solar hours and amplitude 85 kc. These correspond to a relative pressure oscillation of 0.0342 for a height of about 300 km.

551.510.535:525.624 719
Lunar Tidal Oscillations in the Ionosphere—E. V. Appleton and W. J. G. Beynon. (*Nature*, (London), vol. 162, p. 486; September, 25, 1948.) Analysis of hourly values at Slough of F_2-layer critical frequency and of the height h_m of the maximum F_2-layer ionization. Semidiurnal lunar variations have been found in both quantities, but they are approximately in antiphase. The lunar diurnal height variations of the F_2 layer are unexpectedly different in phase from the corresponding E-layer height variations. An investigation of h'_{F_2} data indicates a phase maximum at a time intermediate between those found for the E layer and for h_m.

551.510.535:621.317.79 720
A Panoramic Ionospheric Echo Recorder—Stoffregen. (*See* 789.)

551.594.13 721
Factors Controlling the Atmospheric Conductivity at the Huancayo Magnetic Observatory—W. D. Parkinson. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 305–317; September, 1948.) Measurements of the density and rate of formation of small, intermediate, and large ions give results in agreement with Gish's theory of atmospheric conductivity variations. See also 1615 of 1941 (Gish and Sherman).

551.594.21/.22 722
Photographic Study of Lightning—J. H. Hagenguth. (*Trans. AIEE*, vol. 66, pp. 577–583; 1947. Discussion, pp. 583–585.) Discussion of apparatus used and characteristics of flashes. See also 1362 of 1948 (Malan and Schonland) and back references.

551.594.22:537.521 723
Impulse Characteristics of the Ground under Direct Discharges and with Pointed Electrodes—H. Norinder and G. Petropoulos. (*Ark. Mat. Astr. Fys.*, vol. 35, part 3, section A, 23 pp; September 21, 1948. In English.)

551.594.52(481) 724
Statistics of Heights of Various Auroral Forms from Southern Norway—C. Störmer. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 251–264; September, 1948.) Analysis of the results noted in 1785 of 1947.

LOCATION AND AIDS TO NAVIGATION

534.88 725
Bearing Deviation Indicator for Sonar—O. H. Schuck, C. K. Stedman, J. L. Hathaway, and A. N. Butz, Jr. (*Trans. AIEE*, vol. 66, pp. 1285–1295; 1947.) A review of the development of American asdic methods for submarine location. Switched or split-beam techniques using time-delay or sum-and-difference methods of comparison are described. Block diagrams of apparatus are given, with circuit details.

534.88 726
Submarine Detection by Sonar—A. C. Keller. (*Trans. AIEE*, vol. 66, pp. 1217–1230; 1947.) See also 2427 of 1947.

534.88 727
Sonic Navigation System—Rich and Rosen. (*See* 626.)

621.396.9 728
Using Air-Borne Radar to Increase Airline Safety—R. W. Ayer. (*Trans. AIEE*, vol. 66, pp. 1387–1395; 1947.) General discussion of requirements and of existing equipment for avoiding hills, other aircraft, dangerous storms, etc.

621.396.932:621.396.9 729
Radar Eyes bring Safety to Fog-Bound Liverpool—R. W. Hallows. (*Radio-Electronics* [hitherto *Radio Craft*], vol. 20, pp. 22–23; December, 1948.) See also 3415 of 1948.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788 730
A Combined Thermocouple and Cold-Cathode Vacuum Gauge—R. I. Garrod and K. A. Gross. (*Jour. Sci. Instr.*, vol. 25, pp. 378–383; November, 1948.) Design, construction, and performance details of a vacuum gauge having 2 elements in a common envelope. Pressures in the range 10^{-1} to 10^{-3} mm Hg are measured by a thermocouple gauge, and in the range 10^{-3} to 10^{-6} mm Hg by a cold-cathode discharge gauge which has a simple device for initiating the discharge at low pressures. The cause of the "reversal effect," common in thermocouple gauges, is also investigated.

531.788.7 731
An Investigation on Hot-Wire Vacuum Gauges: Part 2—H. von Ubisch. (*Ark. Mat. Astr. Fys.*, vol. 35, part 3, section A, 12 pp; September 21, 1948. In English.) Part 1: 436 of 1948.

535.37 732
Emission Spectra of Some Zinc Sulfide and Zinc-Cadmium Sulfide Phosphors—W. H. Byler. (*Jour. Opt. Soc. Amer.*, vol. 37, pp. 920–922; November, 1947.) Spectra of 30 such phosphors with varied Cd content and different activators are shown. These spectra suggest that the emission spectrum is not one broad band, but is the sum of contributions from a number of individual bands whose peak positions are invariant and characteristic of the base material rather than the activator. See also 734 below.

535.37 733
Temperature Dependence of the Emission Bands of Zinc Oxide Phosphors—F. H. Nicoll. (*Jour. Opt. Soc. Amer.*, vol. 38, p. 817; September, 1948.) In the temperature range 25° to 250° C the peak in the ultraviolet energy spectrum shifts about 1.2 Å per °C toward longer wavelengths, while the peak in the visible range shows no change.

535.37 734
Emission Spectra of Zinc Cadmium Sulfides—F. J. Studer and D. A. Larson. (*Jour. Opt. Soc. Amer.*, vol. 38, pp. 480–481; May, 1948.) Comment on 732 above. Byler's results appear to differ from those of other workers. The reason for this is not fully understood; it may be due to the photographic method used by Byler.

535.37:535.61-15 735
Infra-Red Stimulability of $CaSiO_3$:Pb and $CaSiO_3$:(Pb+Mn)—J. H. Schulman, R. J. Ginther, and L. W. Evans. (*Jour. Opt. Soc. Amer.*, vol. 38, pp. 817–818; September, 1948.) Infrared response, extending from 0.85 to 1.3 μ, was observed in the case of both phosphors after excitation by a low-pressure Hg-vapor lamp emitting 1849-Å radiation.

546.16:679.5 736
The Development of Fluorine Chemistry—H. J. Emeléus. (*Endeavour*, vol. 7, pp. 141–147; October, 1948.) Discussion of new techniques and polymer production.

548.5 737
New Crystals for Infrared Spectrometry—(*Jour. Frank. Inst.*, vol. 246, pp. 249–250; September, 1948.) A crystal containing about 42 per cent TlBr and 58 per cent TlI has been grown at the National Bureau of Standards. Methods of grinding and polishing such soft crystals have been developed and a prism with faces 1 and three-quarter inches by 2 and one-half inches and refracting angle of 26° has been made. This prism extends the wavelength range of an infrared spectrometer to 40 μ See also 2662 of 1948 (Chasmar).

549.514.51 738
A Determination of the Elastic Constants for Beta-Quartz—E. W. Kammer, T. E. Pardue, and H. F. Frissel. (*Jour. Appl. Phys.*, vol. 19, pp. 265–270; March, 1948.)

621.3(54):[620.193+620.197 739
Electrical Engineering Problems in the Tropics—R. Allan. (*GEC Jour.*, vol. 15, pp. 160–171; October, 1948.) Reprint of 2816 of 1948.

621.315.59 740
The Physics of Electronic Semiconductors—G. L. Pearson. (*Trans. AIEE*, vol. 66, pp. 209–214; 1947.) Present theories are outlined and correlated with quantitative experimental data obtained with typical materials.

621.315.59:535.61-15:621.383 741
The Effect of Room-Temperature Radiation on the Infra-Red Response of Lead Telluride Photoconductors—O. Simpson. (*Proc. Phys. Soc.*, vol. 61, pp. 486–487; November 1, 1948.)

621.315.59:546.289 742
Non-Rectifying Germanium—W. C. Dunlap, Jr., and E. F. Hennelly. (*Phys. Rev.*, vol. 74, p. 976; October 15, 1948.) Ge powder,

obtained by reduction of GeO_2 in a hydrogen furnace, was melted at a pressure of less than 10^{-4} mm Hg. The Ge thus obtained had practically no surface-rectification at a contact with metal. The rectification characteristics at 25°C and −196°C are shown graphically. It is suggested that the effect may either be intrinsic (not a result of inhomogeneity), or due to inhomogeneity on a scale small compared with the diameter of the point contact (0.0002 inch).

621.315.61 743
Effect of Moisture Content on the Dielectric Properties of Some Solid Insulating Materials at U.H.F.—S. K. Chatterjee. (*Indian Jour. Phys.*, vol. 22, pp. 259–264; June, 1948.) Variations of dielectric constant and power factor with moisture content of ebonite and fibre at frequencies from 214 to 750 Mc are discussed.

621.315.61:549.623.5 744
Electrical Properties of Indian Mica: Part 3—The Effect of Pre-Heating—P. C. Mahanti and S. S. Mandal. (*Indian Jour. Phys.*, vol. 22, pp. 7–13; January, 1948.) The power factors of Bengal ruby and Madras green muscovite micas of various qualities have been measured after heat treatment at various temperatures. Heat treatment for one-half hour at 200°C gives a minimum power factor.

621.315.612:546.431.82 745
Theory of the Dielectric Behavior of $BaTiO_3$—J. M. Richardson and B. T. Matthias. (*Phys. Rev.*, vol. 74, pp. 987–988; October 15, 1948.)

621.315.612:546.431.82 746
Dielectric Behavior of Single Domain Crystals of $BaTiO_3$—G. C. Danielson. B. T. Matthias, and J. M. Richardson. (*Phys. Rev.*, vol. 74, pp. 986–987; October 15, 1948.)

621.315.614:621.315.59:621.319.4 747
The Dielectric Properties of Cellulose Insulation Impregnated with Semiconducting Liquids—F. M. Clark. (*Trans. AIEE*, vol. 66, pp. 55–62; 1947. Discussion, pp. 62–63.) The abnormalities are described and discussed, and test results given. A new permalytic type of low-voltage capacitor using semiconducting impregnated paper is considered.

621.315.614.015.5:621.319.4 748
The Probable Breakdown Voltage of Paper Dielectric Capacitors—H. Brooks. (*Trans. AIEE*, vol. 66, pp. 1137–1144; 1947. Discussion, pp. 1144–1145.) Statistical evidence shows that large conducting particles may exist in the paper and bridge one or more layers, through chance orientation during manufacture. The probable voltage strength for a typical grade of paper is calculated.

621.318.2 749
Permanent Magnets—J. L. Salpeter. (*Jour. Brit. I.R.E.*, vol. 8, pp. 211–249; September to October, 1948.) Reprint of 3935 of 1947.

621.318.22+621.318.32 750
Magnetic Materials—G. FitzGerald-Lee. (*Electronic Eng.*, (London), vol. 20, pp. 351–353; November, 1948.) Brief details of modern materials suitable for various applications.

621.318.32 751
Hiperco—A Magnetic Alloy—J. K. Stanley and T. D. Yensen. (*Trans. AIEE*, vol. 66, pp. 714–718; 1947.) Mechanical and electrical characteristics are shown graphically. The effect of heat treatment and of composition variation is discussed. See also 3946 of 1947.

621.318.32:621.317.44 752
Study of Metals at High Frequencies with the Aid of Permeameters with Demountable Coils—I. Épelboim. (*Onde Élec.*, vol. 28, p. 444; November, 1948.) Corrections to 147 of February.

621.775.7:061.3 753
First International Powder Metallurgy Conference—(*Mettalurgia* (Manchester), vol. 38, pp. 227–230; August, 1948.) A brief report, with short accounts of some of the papers presented. These are to be published in their original languages, with summaries in English or German, by the Austrian Chemical Society. See also 123 of February.

621.793 754
Metallizing—A Versatile Method for Production and Maintenance Work—J. E. Wakefield. (*Materials and Methods*, vol. 28, pp. 86–90; September, 1948.) A review of modern methods, with discussion of properties of sprayed deposits. Applications include "printed circuits."

669.14:538.221 755
On the Variation of A.C. Permeability of Transformer Sheet Steels with D.C. Magnetization—B. M. Banerjee. (*Indian Jour. Phys.*, vol. 22, pp. 265–275; June, 1948.) Experiments show that the inverse of ac permeability at constant ac flux density varies almost linearly with the dc magnetization. Oscillograms of the hysteresis loops show that they are symmetrical and that the tips of the loops are bent toward the H-axis; the bending increases with dc magnetization.

669.14–41:538.221 756
Magnetic Sheet Steel—D. Edmundson. (*Engineer* (London), vol. 186, pp. 269–271; September 10, 1948.) A review of the present position in Britain, with special reference to steel for transformers and rotary machines. The properties of cold-rolled anisotropic steel and hot-rolled sheet steel with controlled impurities and grain size are compared with regard to their use in transformers. Losses in induction motors can be reduced by annealing stampings to remove damage caused by punching. Annealing is particularly useful in the case of small machines and can also be used to assist grain growth and for carburization.

MATHEMATICS

517.63 757
Application of the Laplace Transform in the Solution of Linear Integral Equations—L. B. Robinson. (*Jour. Appl. Phys.*, vol. 19, pp. 237–241; March, 1948.) Most of the operations used are illustrated by means of a solution of Abel's integral equation. Results obtained are compared with those of other authors.

518.5 758
An Electronic Differential Analyzer—J. S. Koehler. (*Jour. Appl. Phys.*, vol. 19, pp. 148–155; February, 1948.) The device will solve ordinary nonlinear nonhomogeneous differential equations. It is based on the relation between charge on a capacitor in a series resonant circuit and time, which is expressed by a linear second-order differential equation. The desired variations of the coefficients are obtained from a variable voltage generator whose output can be made to vary with time in accordance with any given curve. The solution is given on an oscillograph, with an accuracy within 4 per cent.

518.5 759
Electronic Digital Computers—F. C. Williams and T. Kilburn. (*Nature*, (London), vol. 162, p. 487; September 25, 1948.) A small experimental "universal" machine consisting essentially of (a) a store for information and orders, (b) various arithmetical organs, such as adders and multipliers, and (c) a control unit. The minimum set of facilities is provided, namely: (i) if x is any number in the store, $-x$ can be written into a central "accumulator" A or, x can be subtracted from what is in A, (ii) the number A can be written in an assigned address in the store, (iii) the content of A can be tested as to whether $x \geqq 0$ or $x < 0$; if $x < 0$ the order standing next in store is passed over, (iv) control can be shifted to an assigned order in the table, (v) the machine can be ordered to stop. The present store has only a capacity of 32 words each of 31 binary digits, and only simple arithmetical testing routines have been carried out.

518.5:512.25 760
An Electronic Simultaneous Equation Solver—E. A. Goldberg and G. W. Brown. (*Jour. Appl. Phys.*, vol. 19, pp. 339–345; April, 1948.) A number of high-gain amplifiers are interconnected by networks whose elements bear definite relationships to the known coefficients of the system of equations. See also 420 of March (Goldberg).

519.271 761
Systematic Sampling [of sequences of quantitative values]—F. Yates (*Philos. Trans.*, vol. 241, pp. 345–377; September, 14 1948.) New methods are evolved for estimating the systematic sampling error from short sections of sequences. Errors due to trend can be eliminated by means of end-corrections. The performance of systematic sampling is investigated theoretically for several functions and for some numerical sequences. The procedure to be adopted for material containing periodicities is discussed.

MEASUREMENTS AND TEST GEAR

531.764.5 762
The Evolution of the Quartz Crystal Clock—W. A. Marrison. (*Bell Sys. Tech. Jour.*, vol. 27, pp. 510–583; July, 1948. Bibliography, pp. 583–588.) A comprehensive review of (a) early methods and apparatus for timekeeping, and (b) the development of quartz oscillators of ever greater absolute frequency constancy and their incorporation in accurate time standards. Methods of comparing the performance of quartz clocks of the highest accuracy are described. Applications of such clocks and their future possibilities are discussed.

534.612.4:621.395.61.089.6 763
Microphone Calibrator—D. H. Bastin. (*Electronics*, vol. 21, pp. 106–109; November, 1948.) An instrument giving automatically a paper record of the frequency response or polar diagram of a microphone. The response from 30 to 1,000 cps is measured in a long sound-absorbing tube. Above 1,000 cps, the measurements are made in an ordinary room; pulse methods enable the record to be made before interfering waves reflected from the walls can affect the apparatus. Logarithmic amplifiers are used.

621.317.011.5:621.315.618 764
Measurements of Dielectric Constant and Dipole Moment of Gases by the Beat-Frequency Method—J. G. Jelatis. (*Jour. Appl. Phys.*, vol. 19, pp. 419–425; April, 1948.)

621.317.3.088 765
The Effect of Waveform on the Accuracy of Rectifier Type Instruments—A. Cunliffe. (*Jour. Sci. Instr.*, vol. 25, pp. 306–307; September, 1948.) Formulas are derived which show the way in which rms and full-wave rectifier-type ac instruments disagree when they are used with a wave form consisting of a fundamental plus a harmonic of order n. With the worst possible phasing conditions, the disagreement is a first-order effect if n is odd and a second-order effect if n is even.

621.317.332 766
Metal Optics at Centimetre Wave-Lengths: Part I—L. Speirs. (*Phil. Mag.*, vol. 39, pp. 105–116; February, 1948.) Theoretical discussion of two methods for investigating surface resistance of thin metal films at λ 1.25 cm. In the first method, a guided wave passes through the film and its support, and reflection and transmission coefficients are derived from

direct measurement of reflected and transmitted waves; in the second method, a resonant cavity is loaded with the film and the resonance frequency is measured before and after insertion.

621.317.34 767

A Highly-Selective Transmission Measuring Equipment for 12- and 24-Channel Carrier Systems—D. G. Tucker and J. Garlick. (*P.O. Elec. Eng. Jour.*, vol. 41, part 3, pp. 166–169; October, 1948.) The general principle of the system was discussed in 3181 of 1947 (Tucker).

621.317.35:621.396.619.16:621.396.813 768

Distortion in a Pulse Count Modulation System—A. G. Clavier, P. F. Panter, and D. D. Grieg. (*Trans. AIEE*, vol. 66, pp. 989–1004; 1947. Discussion, p. 1005.) Full paper. Summary noted in 2281 of 1948.

621.317.37:621.365.92 769

Dielectric Heating—The Measurement of Loss under Rising Temperature—J. B. Whitehead. (*Trans. AIEE*, vol. 66, pp. 947–949; 1947.)

621.317.372 770

Measurement of High *Q* Cavities at 10,000 Mc/s—R. W. Lange. (*Trans. AIEE*, vol. 66 pp. 161–166; 1947.) Possible methods are discussed, and the "heterodyne decrement" method is described in detail. An uhf pulse is applied to the cavity and the decay observed by means of a crystal mixer to which a standard beat frequency is also applied; the crystal then behaves as a linear modulator and an exact knowledge of its characteristic is, therefore, unnecessary.

621.317.384 771

Some Aspects of the Theory of Iron-Testing by Wattmeter and Bridge Methods—N. F. Astbury. (*Jour. IEE* (London), part II, vol. 95, pp. 607–616; October, 1948. Summary, *ibid.*, part I, vol. 95, p. 406; September, 1948.) The limitations of the dynamometer are discussed, with special reference to eddy-current, circuit phase-angle and loss compensation, leakage flux, and harmonic distortion. Null methods are outlined and a new circuit is given. Discussion of bridge methods is based on the concept of complex permeability and a method of eliminating copper losses is described. A distortion coefficient is defined.

621.317.431 772

A 60-Cycle Hysteresis Loop Tracer for Small Samples of Low-Permeability Material—D. E. Wiegand and W. W. Hansen. (*Trans. AIEE*, vol. 66, pp. 119–131; 1947. Discussion, pp. 131–133.) The basic components are: (a) a large (35-lb) exciting coil with a pickup coil at its center, (b) an amplifier and integrating circuit, and (c) a cro.

621.317.431:621.317.755 773

Quantitative Determination of Magnetic Properties by Use of Cathode-Ray Oscilloscope—J. Zamsky. (*Trans. AIEE*, vol. 66, pp. 783–787; 1947.) A detailed description of a method of displaying hysteresis loops on a cro.

621.317.44:621.318.32 774

Study of Metals at High Frequencies with the Aid of Permeameters with Demountable Coils—I. Épelboim. (*Onde Élec.*, vol. 28, p. 444; November, 1948.) Corrections to 147 of February.

621.317.6:534.232:681.85 775

Vibrators for Measurement of Response and Compliance of Phonograph Pick-Ups—H. A. Pearson, R. W. Carlisle, and H. Cravis. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 830–833; November, 1948.)

621.317.6:621.396.611.1 776

The Response of a Resonant System to a Gliding Tone—Barber and Ursell. (*See* 670.)

621.317.6:621.396.645.012 777

Very-Wide-Band Response-Amplitude Curve Tracer—M. A. Jullien. (*Onde Élec.*, vol. 28, pp. 388–390; October, 1948.) Summary only. The general conditions which such equipment should satisfy are stated and possible types, with different methods of FM, are discussed. A new curve tracer, developed in the C.F.T.H. laboratories, is described. This has a frequency excursion continuously adjustable from 0 to 200 Mc, the mean frequency being independently adjustable between 2 and 1,100 Mc. The apparatus uses the beats between a reflex klystron, oscillating in the 3-cm band, and a positive-grid triode with Lecher-line plate circuit operating at wavelengths around 9 cm. The signal output level is of the order of 0.1 volts. Operation is described in detail; a diagram shows the arrangement of the various parts.

621.317.715.5 778

New Method of Increasing the Voltage Sensitivity of Moving-Coil Galvanometers—J. Coursaget. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 673–675; October 4, 1948.) An arrangement applicable to any galvanometer is described, which enables the electromagnetic damping to be reduced while maintaining the original flux, so that a very high voltage sensitivity can be reached. The oscillation period can also be reduced. In a particular case, the period was reduced from 8.5 to 5 seconds and the critical resistance from 250 to 30 Ω, while the sensitivity was increased from 7.0×10^{-8} to 0.8×10^{-8}v/mm at 1 meter. See also 1085 of 1948 (Dupouy).

621.317.725 779

A Range of Kilovoltmeters for High D.C. Voltages [up to 500 kv]—F. W. Waterton (*Jour. Sci. Instr.*, vol. 25, pp. 304–306; September, 1948.) Each voltmeter comprises an oil-immersed voltage divider, with either an electrostatic indicator or a microammeter and very high resistance connected between one end and a tapping point. Errors in indication are caused by the voltage versus resistance and temperature versus resistance characteristics of the units used in constructing the divider. These units should all be of the same make and have the same nominal value.

621.317.725.027.7 780

Absolute Measurement of High Voltages by Oscillating Electrode Systems—E. Bradshaw, S. A. Husain, N. Kesavamurthy, and K. B. Menon. (*Jour. IEE* (London), part II, vol. 95, pp. 636–641; October, 1948. Discussion, pp. 641–644. Summary *ibid.*, part I, vol. 95, p. 411; September, 1948.) Full paper; summary noted in 2561 of 1948.

621.317.726.089.6 781

The Calibration of Ignition Crest Voltmeters—W. L. Davis and C. E. Warren. (*Trans. AIEE*, vol. 66, pp. 99–104; 1947. Discussion, p. 104.) Discussion of circuits and techniques for producing consistent voltage wave forms adjustable over a considerable range of peak voltage, rise time, and repetition rate, in order to investigate discrepancies between various types of voltmeters.

621.317.727:518.5 782

Potentiometers for Computing Circuits—R. W. Williams. (*Electronic Eng.* (London), vol. 20, pp. 358–360; November, 1948.) The error introduced by loading is discussed and illustrated by a numerical example. Temperature effects and potentiometers with graded windings are also considered.

621.317.728 783

A Note on the Measurement of Short-Duration Recurrent Voltage Impulses by Means of Spark Gaps—R. Cooper (*Jour. IEE* (London) part II, vol. 95, pp. 378–382; August, 1948. Summary, *ibid.*, part I, vol. 95, p. 404; September, 1948.) The calibration data noted in British Standards Institution publication B.S. 358:1939 for 2-cm spheres can be used for recurrent pulses of duration as short as 0.1 to 4 μ seconds. The breakdown voltage of the gaps was found to be independent of both pulse duration and recurrence rate if the gap was irradiated by 0.2 mg of Ra. For gaps between parallel-plate electrodes more than 2 mm apart, such irradiation did not affect the breakdown voltage for 1-μ second pulses with a repetition rate of 400 per second.

621.317.733:621.3.083.4 784

Electronic Null Detectors for Use with Impedance Bridges—H. W. Lamson. (*Trans. AIEE*, vol. 66, pp. 535–540; 1947.)

621.317.733:621.317.738 785

A Self-Balancing Capacitance Bridge—A. H. Foley. (*Trans. AIEE*, vol. 66, pp. 797–801; 1947.) Intended for testing mass-produced capacitors, scale indication being in the form of percentage deviation from the proper value. Accuracy claimed is about 0.1 per cent over a capacitance range of 1,000 to 1. A servomechanism balances the bridge in about 2 seconds.

621.317.733:621.392.26† 786

A Waveguide Bridge for Measuring Gain at 4000 Mc—A. L. Samuel and C. F. Crandell. (PROC. I.R.E., vol. 36, pp. 1414–1418; November, 1948.) The equipment is described and methods of reducing possible errors are discussed. The general method can be adapted for use in any desired frequency range.

621.317.761:621.3.015.33 787

A Pulse Deviation Meter—D. I. Lawson and E. R. Rout. (*Jour. Sci. Instr.*, vol. 25, pp. 309–311; September, 1948.) The full-scale reading of the most sensitive range corresponds to unit variation in pulse recurrence frequencies between 10,000 and 30,000 per second. The deviation of either the leading or the trailing edge of the pulse trains can be measured.

621.317.7 788

Advancements in the Design of Long-Scale Indicating Instruments—R. M. Rowell and N. P. Millar. (*Trans. AIEE*, vol. 66, pp. 155–160; 1947.) For another account see 3173 of 1948.

621.317.79:551.510.535 789

A Panoramic Ionospheric Echo Recorder—W. Stoffregen. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 269–271; September, 1948.) A frequency sweep of 1.4 to 14 Mc is made in 3 to 5 seconds and the complete curve of equivalent height versus frequency is displayed on a long-afterglow cathode-ray tube. Height and frequency calibration marks are included in the picture.

621.317.79:621.315.2 790

Pulse Echo Measurements on Telephone and Television Facilities—L. G. Abraham, A. W. Lebert, J. B. Maggio, and J. T. Schott. (*Trans. AIEE*, vol. 66, pp. 541–548; 1947. Discussion, p. 548.) See 2295 of 1948.

621.317.79:621.395.813:621.395.625(083.74) 791

Proposed Standards for the Measurement of Distortion in Sound Recording—(*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 449–466; November, 1948. Discussion, p. 467.) Draft proposals under consideration by a subcommittee of the American Standards Association Committee on Standards for Sound Recording.

621.317.79:621.396.615.12 792

Design of a Continuously Variable Audio Signal Generator—B. Bauer. (*Audio Eng.*, vol. 32, pp. 15–17, 43; November, 1948.) A stable, accurate unit for testing high-quality audio apparatus. Full circuit details are included.

621.317.79:[621.396.615.14+621.396.621 (083.74) 793

Standard Receiver and Generator for Ultra-High Frequencies—R. Cabessa and G. Phélizon. (*Onde Élec.*, vol. 28, pp. 423–432 and 482–486; November and December, 1948.) An account of two instruments developed at the Laboratoire Central de Télécommunications specially for uhf measurements on receivers, transmitters, and antennas. Some details are given of velocity-modulation tubes of a coaxial type with high-frequency stability and low signal-to-noise ratio. Simultaneous variation of the cavity tuning and the capacitive coupling is effected by axial adjustment of the central conductor. One of these tubes covers the wavelength range 15 to 30 cm, a second 8 to 15 cm, and a third 6 to 8 cm. The standard generator gives signals of wavelength from 6 to 30 cm, with square-wave AM up to 100 per cent, output power adjustable continuously from 1 μw to 10^{-14} w, and direct reading for receiver noise factor. The wavelength range of the standard receiver is at present 10 to 15 cm; this will shortly be extended to 6 to 30 cm. Its use for the following measurements is explained: (a) sensitivity of receivers, (b) signal-to-noise ratio of a modulated transmitter, (c) tracing of antenna radiation diagrams, and (d) absolute measurement of field strength.

621.317.79:621.396.822 794

A Direct-Reading Instrument for the Measurement of Noise Factor, and Its Application to the Testing of Microwave Mixer Crystals—L. A. G. Dresel, L. A. Moxon, and E. E. Schneider. (*Jour. Sci. Instr.*, vol. 25, pp. 295–298; September, 1948.) The signal generator is a noise source using a coaxial-line temperature-limited diode; the mean amplitude of the noise is substantially square-wave, modulated at 50 cps. This is fed into the mixer of a conventional receiver. An agc system is used to hold the maximum receiver output constant. The difference between the maximum and minimum receiver outputs is then read as a 50-cps voltage, on an indicator which can be calibrated in terms of noise factor.

621.317.79:621.397.62.001.4 795

Note on Television Test Equipment—Kniazeff. (*Onde Elec.*, vol. 28, pp. 391–394; October, 1948. Summary only.) Reasons are given for the use of square-wave pulses, and apparatus suitable for testing the various circuits of television receivers is discussed.

621.396.69.001.4:621.396.621 796

Selecting Components for Broadcast Receivers—G. D. Reynolds. (*Electronic Eng.* (London), vol. 20, pp. 307–313; October, 1948.) Long summary of IEE paper. Mechanical, electrical, and chemical tests are considered. There are three kinds of each: measurements, life tests, and peak-load or overload tests. The design of test equipment is discussed in the light of the limitations of both the component (or raw material) and the methods of test available, and specific examples are mentioned.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

535.336.2.05:621.389 797

Radio-Frequency Mass Spectrometer—(*Electronics*, vol. 21, pp. 124, 126; November, 1948.) Positive or negative ions formed near the cathode of a tube are accelerated into a rf field which is varied in frequency and amplitude so that only ions of one particular mass pass to the plate.

535.61-15 798

The Infra-Red Image Converter Tube—T. H. Pratt. (*Electronic Eng.* (London), vol. 20, pp. 274–278 and 314–316; September and October, 1948.) Operating principles are discussed, developments in Europe and in America are outlined and the special features of the German AEG tube, various RCA tubes and the English EMI (Electric and Musical Industries) tube are described. The EMI tube represents the best compromise between performance and complexity in the applications for which it was originally intended. Its pyrex envelope is only 5 cm in diameter and 4 cm long, with plane end windows 2 mm thick. It gives uniform resolution over the whole field of view. Examples of its application in various types of military equipment are described and some details are given of the Zamboni pile (862 below) developed for use with such equipment. Research and commercial applications are briefly discussed.

538.74:621.385.832 799

A Magnetic Compass with Cathode-Ray Sensing Element—W. H. Kliever and R. R. Syrdal. (*Trans. AIEE*, vol. 66, pp. 529–534; 1947.) Full account of an instrument briefly described in 3206 of 1947 (Squier).

539.16.08 800

On the Life of Self-Quenching Counters—S. S. Friedland. (*Phys. Rev.*, vol. 74, pp. 898–901; October 15, 1948.)

539.16.08 801

A Study of the Deterioration of Methane-Filled Geiger-Müller Counters—E. C. Farmer and S. C. Brown. (*Phys. Rev.*, vol. 74, pp. 902–905; October 15, 1948.)

539.16.08:621.318.572 802

Electronic Counters for Impulses—Naslin and Peuteman. (*See* 660.)

550.837.7:621.3.091:553.57 803

The Attenuation of Ultra-High Frequency Electromagnetic Radiation by Rocks—McPetrie and Saxton: Cooper. (*See* 819.)

620.179.16:534.321.9.001.8 804

Design and Application of Supersonic Flaw Detectors—D. C. Erdman. (*Trans. AIEE*, vol. 66, pp. 1271–1276; 1947.) A small quartz crystal converts electrical 15-Mc radiation into ultrasonic $\frac{1}{2}$-μ second pulses. These pulses are reflected from any flaw in a metal forging back to the crystal and re-converted to electrical energy. Block diagrams are given. Oil or glycerine ensures good acoustic coupling with the forging.

621.317.083.7:623.746.48 805

Telemetering Guided-Missile Performance—J. C. Coe. (PROC. I.R.E., vol. 36, pp. 1404–1414; November, 1948.) The functions required of a telemetering system and the conditions under which it must operate are discussed. Brief details are given of various forms of transducer, which convert physical into electrical quantities. Two of the more important types of telemetering systems are described: one involves FM of subcarriers; the other is a pulse-position modulation system.

621.365.92:621.317.37 806

Dielectric Heating—The Measurement of Loss under Rising Temperature—J. B. Whitehead. (*Trans. AIEE*, vol. 66, pp. 947–949; 1947.)

621.38 807

The Manchester Electronics Exhibition—(*Electronic Eng.* (London), vol. 20, pp. 296–297; September, 1948.) Brief descriptions of some of the exhibits.

621.38:621.316.718:655.324.5 808

Radar Technique in an Industrial Control—W. D. Cockrell. (*Trans. AIEE*, vol. 66, pp. 269–272; 1947.) A system for register control in printing. See also 789 of 1948 (Ludwig).

621.38.001.8:669.1 809

The Use of Electronic Instruments in Iron and Steel Making—S. S. Carlisle. (*Engineer* (London), vol. 186, pp. 450–451 and 476–477; October 29 and November 5, 1948.) Discussion of electronic techniques for measuring temperatures and small differential pressures, for CO and CO_2 estimation, etc. While electronic methods of amplification and detection enable very small quantities to be detected and measured, industrial requirements of ease of maintenance and satisfactory operation under severe conditions of temperature, dust, etc. are very stringent. Electronic methods are not necessarily the best available.

621.384:621.319.3 810

The Palletron, A New Electron Resonator and Its Proposed Application to the Generation of Potentials in the Million-Volt Range—A. M. Skellett. (*Jour. Appl. Phys.*, vol. 19, pp. 187–193; February, 1948.) In an electrostatic field of parabolic potential distribution, an electron will have simple harmonic motion. A gap at the center of the field provides a means of exchange of energy between electrons and the associated circuit. If energy is taken from the electrons, the device is an oscillator; if energy is given to them, the device may be used to generate a high dc voltage. Experimental results on a small model of the high-voltage generator are given and a proposed design for the million-volt range is briefly described. Summary noted in 3996 of 1947.

621.384.6:621.396.611.4 811

A Resonant Cavity Linear Accelerator—A. B. Cullen, Jr., and J. H. Greig. (*Jour. Appl. Phys.*, vol. 19, pp. 47–50; January, 1948.) A folded rectangular waveguide cavity resonant, at 2,800 Mc is used to accelerate electrons, in three stages, from an injected energy of 2 to 300 kev.

621.385.1.001.8:531.768.087 812

The Measurement of Acceleration with a Vacuum Tube—W. Ramberg. (*Trans. AIEE*, vol. 66, pp. 735–740; 1947.) For another account see 2528 of 1947.

621.385.833 813

Electron Lenses of Hyperbolic Field Structure: Part 1—R. Rüdenberg. (*Jour. Frank. Inst.*, vol. 246, pp. 311–339; October, 1948.) An electric field of hyperbolic structure focuses uniform parallel rays without aberration. A rigorous mathematical analysis is made possible by the independence of the field equations in the radial and axial directions. Trajectories of the electrons are calculated; the cardinal points are found and compared with those of a glass lens, and Newton's formula is shown to apply to the electron lens.

621.385.833 814

The Variation of Beam Angle with Modulation in Electron-Optical Immersion Systems—L. Jacob. (*Phil. Mag.*, vol. 39, pp. 400–408; May, 1948.) A mathematical proof of the dependence of beam angle on modulation, using certain simplifying assumptions. The theory is confirmed by experiment.

623.978+550.838]:538.71 815

Air-Borne Magnetometers for Search and Survey—E. P. Felch, W. J. Means, T. Slonczewski, L. G. Parratt, L. H. Rumbaugh, and A. J. Tickner. (*Trans. AIEE*, vol. 66, pp. 641–651; 1947.) See 220 of 1948.

PROPAGATION OF WAVES

538.566 816

A General Divergence Formula—H. J. Riblet and C. B. Barker. (*Jour. Appl. Phys.*, vol. 19, pp. 63–70; January, 1948.) A divergence expression is derived for the ratio of energy per steradian reflected from a smooth curved surface to that incident on the surface, when the source and the point of observation are both at finite distances from the reflecting surface. The wavelength is assumed small compared to the radii of curvature of the surface.

538.566 817

New Methods in Diffraction Theory—V. A. Fock. (*Phil. Mag.*, vol. 39, pp. 149–155; February, 1948.) Development and discussion of a method for general and practical solution of problems in diffraction of electromagnetic waves around obstacles of arbitrary shape. The basis of the method is that the transition from light to shadow on the obstacle's surface occurs in a narrow strip along the boundary of the geometrical shadow, the field in this strip is shown to depend only on the value of the field of the incident wave near the point considered and on the geometrical and electrical properties of the diffracting body. See also 2892 of 1947 (Booker and Walkinshaw).

538.566 818
Two Theorems Relative to the Propagation of Sinusoidal Waves in Stratified Media—F. Abelès. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 899–900; November 3, 1948.)

621.3.091:550.837.7:553.57 819
The Attenuation of Ultra-High-Frequency Electromagnetic Radiation by Rocks—J. S. McPetrie and J. A. Saxton: R. I. B. Cooper. (*Proc. Phys. Soc.*, vol. 61, pp. 482–483; November 1, 1948.) Comment on 3476 of 1948 (Cooper).

621.396.11:551.510.535 820
The Estimation and Forecasting of Short-Wave Propagation Conditions, with Special Reference to Naval Communications—L. E. Beghian. (*Jour. IEE* (London), part III, vol. 95, pp. 351–362; September, 1948; Summary, *ibid.*, part I, vol. 95, pp. 459–460; October, 1948.) Description of Admiralty techniques for using ionospheric data in the solution of high-frequency communication problems. The computation of ionospheric absorption is discussed and semi-empirical expressions are given for use for distances less than or in excess of 2,500 miles. Methods for determining the lowest usable high-frequency are based on these expressions. Ship-to-shore communication and the choice of optimum frequencies are also considered.

621.396.11.029.6 821
Radiation and Propagation of Electromagnetic Waves of Short Wavelength—G. Goudet: J. Voge. (*Ann. Télécommun.*, vol. 3, pp. 74–84, 113–125, 155–179, 182–208, and 233–256; March to July, 1948.) A survey of recent work, both theoretical and experimental, on cm and dm waves, with a bibliography of 137 important papers mentioned in the text. Part 1 discusses electromagnetic radiation theory and cm-wave equipment. Part 2 discusses reflection, refraction, diffraction, properties of the ionosphere, and the effect of meteorological conditions. Radar receives special attention.

621.396.11.029.6 822
The Effect of Ground Constants on the Characteristic Values of the Normal Modes in Non-Standard Propagation of Microwaves—C. L. Pekeris. (*Jour. Appl. Phys.*, vol. 19, pp. 102–105; January, 1948.) An investigation dealing with both vertically and horizontally polarized waves in the wavelength range 1 to 50,000 cm.

621.396.812.3:551.510.535 823
Fading of Short-Wave Radio Signals and Space-Diversity Reception: Part 1—S. S. Banerjee and G. C. Mukerjee. (*Phil. Mag.*, vol. 39, pp. 697–712; September, 1948.) Observations were made on signals from various short-wave stations, using a superheterodyne receiver with either a mirror galvanometer or an automatic recorder connected in the circuit of the diode second detector. The angle of arrival of downcoming waves was measured by Appleton and Barnett's method. Typical records are given and results are tabulated and correlated with ionospheric layer heights and electronic densities. For space-diversity reception, vertical separation of the antennas is more effective than horizontal.

621.396.11:551.510.535 824
N.B.S. Circular 462: Ionospheric Radio Propagation [Book Notice]—National Bureau of Standards. U. S. Government Printing Office, Washington, D. C. $1.00. (*Jour. Res. Nat. Bur. Stand.*, vol. 32, p. 123; October, 1948.) The physical and mathematical theory underlying electromagnetic wave propagation, and its relation to practical problems of radio communication, are discussed.

621.396.11:551.510.535 825
Radio Research Special Report No. 18: Application of Ionospheric Data to Radio Communication [Book Notice]—Department of Scientific and Industrial Research. H.M. Stationery Office, London, 1948, 1s. (*Govt. Publ.* (London), Daily List No. 254, p. 2; December 30, 1948.)

RECEPTION

621.396.621:621.396.619.13 826
On the Concept of Instantaneous Frequency—J. Laplume. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 722–724; October 11, 1948.) Discussion with particular reference to FM discriminators.

621.396.621(083.74)+621.396.615.12]: 621.317.79 827
Standard Receiver and Generator for Ultra-High Frequencies—Cabessa and Phélizon. (*See* 792.)

STATIONS AND COMMUNICATION SYSTEMS

621.39 828
Telecommunications for the 1948 Olympic Games—E. R. Smith and C. W. Sallnow. (*P.O. Elec. Eng. Jour.*, vol. 41, part 3, pp. 157–162; October, 1948.) An exchange with a multiple capacity of 2,800 was installed by the British Post Office. Lines were provided for administration, BBC programs, field events, teleprinter networks, and telephone services for public and press. See also 522 of March (Hotine).

621.395.44:621.315.052.63 829
A Carrier Telephone System for Rural Service—J. M. Barstow. (*Trans. AIEE*, vol. 66, pp. 501–507; 1947.) For another account see 2612 of 1948.

621.395.44:621.315.052.63 830
Application of Rural Carrier Telephone System—E. H. B. Bartelink, L. E. Cook, F. A. Cowan, and G. R. Messmer. (*Trans. AIEE*, vol. 66, pp. 511–517; 1947. Discussion, pp. 517–518.) Discussion of modifications required in the power circuits to permit carrier-frequency transmission. See also 829 above and back references.

621.396 831
Technical Problems of Military Radio Communications of the Future—J. Hessel. (Proc. I.R.E., vol. 36, pp. 1402–1403; November, 1948.) A communication system of adequate mobility, traffic capacity, and reliability is required. The factors prohibiting the present realization of such a system are discussed, and basic research problems outlined.

621.396.1 832
Copenhagen Frequency Allocations—(*Wireless World*, vol. 54, pp. 397–399; November, 1948.) New wavelengths for European broadcasting stations are listed. In some cases, directional antennas protecting particular regions must be used. Particulars are given of 8 additional BBC transmitter locations and of the way in which the BBC proposes to use the 14 wavelengths now allotted to Great Britain.

621.396.3:621.394.441 833
A Multi-Channel Radio Telegraph Equipment—G. N. Davison and R. J. Pickard. (*P.O. Elec. Eng. Jour.*, vol. 41, part 3, pp. 148–153; October, 1948.) The advantages of 2-tone and diversity operation are discussed briefly. A description is given of 2-tone voice-frequency equipment developed by the British Post Office for use by the Services on single-sideband high-frequency radio circuits using triple space-diversity reception. Satisfactory operation is obtained with signals varying from +5 to −45 db relative to 1 mw. The small time-constant of the agc circuit enables the receivers to follow deep and rapid fading.

621.396.44:621.315.052.63 834
A New Single-Side-Band Carrier System for Power Lines—B. E. Lenehan. (*Trans. AIEE*, vol. 66, pp. 826–830; 1947. Discussion, p. 830.) See 4027 of 1947.

621.396.619 835
Frequency Analysis of Modulated Pulses—S. H. Moss. (*Phil. Mag.*, vol. 39, pp. 663–691; September, 1948.) A mathematical analysis of the frequency spectra of modulated recurrent pulses of different types. The relationship between the form of the modulated carrier and the applied modulation wave form is studied, with particular reference to single- and double-tone wave forms. Five types of modulation are investigated: phase, frequency, and amplitude modulation of indefinitely narrow unit pulses, and symmetric and asymmetric width modulation of ideal rectangular pulses of unit amplitude. The results are summarized in tables.

621.396.619 836
Modulation in Communication.—F. A. Cowan. (*Trans. AIEE*, vol. 66, pp. 792–796; 1947.)

621.396.619 837
Composite Amplitude and Phase Modulation—O. G. Villard, Jr. (*Electronics*, vol. 21, pp. 86–89; November, 1948.) In this system, a carrier is modulated in both phase and amplitude. The phases of the af inputs to the two modulators differ by 90°. By adjusting the depth of modulation, the lower sidebands can be cancelled, leaving a single-sideband transmission with first-order sideband level corresponding to 100 per cent AM, but with appreciable second-order sidebands. A normal AM receiver can be used.

621.396.619.13:621.396.65 838
The Application of Heterodyne Modulation to Wide-Band Frequency-Modulated Television Relays—W. P. Boothroyd. (*Trans. AIEE*, vol. 66, pp. 1126–1130; 1947.) Heterodyne modulation consists in applying FM to a fixed carrier frequency and selecting the upper sideband. In a particular equipment considered, a FM band of 107 to 124 Mc was used with a carrier frequency of approximately 1,235 Mc furnished by a klystron. Apparatus design is discussed.

621.396.619.13:621.396.97:621.396.621 839
F.M. Broadcast Network With Radio Links—D. K. de Neuf. (*Communications*, vol. 28, pp. 12–15; October, 1948.) For another series of accounts see 494 of March (Sleeper) and cross references.

621.396.619.15 840
Frequency Shift Telegraphy—Radio and Wire Applications—J. R. Davey and A. L. Matte. (*Trans. AIEE*, vol. 66, pp. 479–493; 1947. Discussion, pp. 493–494.) See 2362 of 1948.

621.396.619.15:621.394.441 841
Frequency-Shift Keying—T. Roddam. (*Wireless World*, vol. 54, pp. 400–402; November, 1948.) Comparison with on-off keying methods.

621.396.619.16 842
Pulse Code Modulation—H. S. Black and J. O. Edson. (*Trans. AIEE*, vol. 66, pp. 895–899; 1947.) Full paper. Summary noted in 2363 of 1948.

621.396.619.16 843
The Philosophy of PCM—B. M. Oliver, J. R. Pierce, and C. E. Shannon. (Proc. I.R.E., vol. 36, pp. 1324–1331; November, 1948.) Some of the advantages of pulse-code modulation are discussed and the possible achievements of the system are compared with those of a FM system. In general, pulse-code modulation seems ideally suited for multiplex message circuits where good quality and high reliability are required.

621.396.619.16 844
Decoding in P.C.M.—R. L. Carbrey. (*Bell Lab. Rec*, vol. 26, pp. 451–456; November, 1948.)

621.396.619.16:621.395.43 845
Multiplex Telephony Systems with Impulse Modulation—S. van Mierlo. (*Tijdschr. ned. Radiogenoot.*, vol. 13, pp. 135–170; September, 1948. Discussion, p. 171. In Dutch, with English summary.) Discussion of: (a) the relative merits of pulse-amplitude, pulse-position, and pulse-code modulation systems, in which the pulses may be used for AM or FM of a carrier wave, (b) bandwidth and signal-to-noise ratios, (c) distributors, modulators, and demodulators, (d) two experimental pulse-position modulation systems, and (e) two commercial equipments now available.

621.396.619.16:621.396.813 846
Distortion and Band-Width Characteristics of Pulse Modulation—H. L. Krauss and P. F. Ordung. (*Trans. AIEE*, vol. 66, pp. 984–988; 1947.) The minimum allowable ratio of pulse repetition frequency to maximum af is expressed in terms of the distortion and the percentage modulation. The effects of pulse width, pulse shape, and percentage modulation on the required bandwidth are also discussed.

621.396.619.16:621.396.813:621.317.35 847
Distortion in a Pulse Count Modulation System—Clavier, Panter, and Grieg. (*See* 768.)

621.396.65 848
Indirect Microwave Relay System—In 219 of February please read R. P. Wakeman for R. R. Wakeman.

621.396.65:621.396.615.142.2 849
ST [studio/transmitter] Equipment Using Klystrons—In 220 of February insert M. Silver and H. French as authors.

621.396.65:[621.397.5+621.395.43 850
Use of Radio Links for the Transmission of Television and Multiplex Telephony Signals—J. Laplume. (*Onde Élec.*, vol. 28, pp. 396–397; October, 1948.) Summary only. Recent investigations have shown that for these purposes radio links have many definite advantages and are much cheaper than cable links.

621.396.65.029.64:621.397.743 851
A New Microwave Television System—J. F. Wentz and K. D. Smith. (*Trans. AIEE*, vol. 66, pp. 465–470; 1947.) Describes a point-to-point relay system using FM with center frequency between 3,900 and 4,400 Mc, and 4 to 5-Mc video signal. Filters for 2-channel operation on one antenna (paraboloid or lens) are provided. Block digarams and some test and performance figures are given. See also 1755 of 1948 (Durkee) and 1756 of 1948 (J.M.).

621.396.7 852
Criggion Radio Station—A. Cook and L. L. Hall. (*P.O. Elec. Eng. Jour.*, vol. 41, pp. 123–129; 1948.) For another account see 2371 of 1948 (West, Cook, Hall, and Sturgess).

SUBSIDIARY APPARATUS

621-526 853
Dimensionless Analysis of Servomechanisms by Electrical Analogy: Part 1—G. D. McCann and S. W. Herwald. (*Trans. AIEE*, vol. 66, pp. 111–118; 1947.

621-526 854
A Comparison of Two Basic Servomechanism Types—H. Harris. (*Trans. AIEE*, vol. 66, pp. 83 92; 1947. Discussion, pp. 92–93.)

621-526 855
The Analysis and an Optimum Synthesis of Linear Servomechanisms—D. Herr and I. Gerst. (*Trans. AIEE*, vol. 66, pp. 959–970; 1947.)

621-526 856
Stabilizing Servomechanisms—D. McDonald. (*Electronics*, vol. 21, pp. 112–116; November, 1948.) A circuit having a transfer function which is the inverse of the transfer function of the servomechanism can be used as a stabilization network. Such a circuit can be approximately realized by means of a feedback amplifier having in its feedback path a network whose transfer function is proportional to that of the servomechanism.

621-526 857
Stabilization of Carrier-Frequency Servomechanisms: Parts 1–3—A. Sobczyk. (*Jour. Frank. Inst.*, vol. 246, pp. 21–43, 95 121, and 187–213; July to September, 1948.)

621-526:061.3 858
Some Particularly Interesting Points brought out at the Congress on Servomechanisms, London, May, 1947—M. Naslin. (*Onde Élec.*, vol. 28, pp. 445–454; November, 1948.) Discussion of selected papers whose titles were noted in 4039 of 1947.

621.316.722 859
Solution of the General Voltage Regulator Problem by Electrical Analogy—E. L. Harder. (*Trans. AIEE*, vol. 66, pp. 815–825; 1947. Discussion, p. 825.)

621.316.722.1 860
Regulator for 400-c/s Inverter—C. A. Helber. (*Electronics*, vol. 21, pp. 90–91; November, 1948.) Describes stabilizing equipment for a 250-volt-ampere aircraft rotary converter. A voltage-regulator tube controls, through a thyratron and a saturable reactor, the ac input to the rectifier supplying the field excitation for the converter.

621.316.74 861
Some Aspects of Moderate Precision Temperature Control in Communication Engineering—M. P. Johnson. (*Jour. Brit. I.R.E.*, vol. 8, pp. 250–259; September and October, 1948.) Discussion with special reference to the design of master-oscillator ovens. Thermostats of the bimetallic and Hg type are considered and three electronic thermostats are treated in more detail, with applications to practical oven design. Experimental results show the influence of heat insulation in reducing the effects of ambient-temperature variations and also on thermostat ripple.

621.352.32 862
The Dry Voltaic Pile—A. Elliott. (*Electronic Eng.* (London), vol. 20, pp. 317 319; October, 1948.) A small, light, and reliable source of high voltage with particular application to infrared telescopes. Development of the pile from the laboratory experimental stage to mass production is discussed.

621.396.68:621.316.722 863
Stabilized Power Supplies: Part 2—Some Refinements and Modifications—M. G. Scroggie. (*Wireless World*, vol. 54, pp. 415–418; November, 1948.) Continuation of 231 of February.

621.396.69 864
Highlights of the "Super-Power" 8-Section Pylon—O. O. Fiet. (*Broadcast News*, no. 50, pp. 36–51; August, 1948.) Effective FM power up to 1,200 kw can be radiated.

TELEVISION AND PHOTOTELEGRAPHY

061.3:621.397.5 865
Television Conference, Paris, 25th–30th October 1948—(*Onde Élec.*, vol. 28, pp. 350 420; and 457 468; October to December, 1948.) Summaries of papers read at the Conference. For selected individual abstracts see this and other sections.

061.3:621.397.5 866
The Television Congress, Paris, 25th–30th October 1948—Y. Angel. (*Onde Élec.*, vol. 28, pp. 457 460; December, 1948.) A short account of the general organization.

061.3:621.397.5 867
The International Television Convention at Zurich—(*Electronic Eng.* (London), vol. 20, pp. 362–363; November, 1948.) The lectures included "Television and Outside Broadcast Practice in Great Britain," by T. H. Bridgewater, and an account by H. Thiemann of the development of the Swiss A.F.I.F. (Abteilung für industrielle Forschung) large-screen system based on the variations in the angle of refraction of light directed on to the surface of a thin oil film (see 296 of 1948). A demonstration of this method using the French ériscope camera gave pictures of cinema standard. The need for international standards was discussed.

621.397.2:551.509.2 868
Television Broadcasting of Meteorological Information—R. Clausse. (*Onde Élec.*, vol 28, pp. 358–360; October, 1948.) Summary only. Discusses the principles of and practical equipment for the rapid automatic transmission of meteorological charts.

621.397.3:778.534.4 869
Color-Television Film Scanner—B. Erde. (*Jour. Soc. Mot. Pic. Eng.*, vol. 51, pp. 351–372; October, 1948.) The film moves at a constant speed across a gate which is double the height of the picture frame. A fixed optical system of 6 lenses and filters and a rotating shutter project the successive primary-color pictures on to a nonstorage type of pickup tube. Compensated electronic scanning equipment allows for the continuous motion of the film.

621.397.331.2 870
"Knight" Scanning—P. M. G. Toulon. (*Onde Élec.*, vol. 28, pp. 412–416; October, 1948.) Summary only. Various methods of scanning are discussed and a method is described in which the order of scanning the elementary squares into which the picture is divided corresponds to a series of moves of a chess knight. The advantages of this method are enumerated. Definition for a 450-line system is comparable with that of a 1,000-line system using ordinary interlaced scanning.

621.397.5 871
A New Process for Television in Colour—Y. Angel. (*Onde Élec.*, vol. 28, pp. 353–354; October, 1948.) Summary only. The method proposed uses only a single analysis: that of a triple image formed by the juxtaposition of 3 primary monochromatic images produced by optical methods of decomposition of the colored image to be transmitted. In reception, the 3 primary images are observed through filters and an optical system which assures their proper superposition. Equipment suitable for such a system is discussed.

621.397.5 872
After the Television Congress—H. Chauvierre. (*Radio Franç.*, no. 12, pp. 1–4; December, 1948.) Comparison of television developments in America, Britain, and France shows that France at present is considerably behind, although the French super-ériscope will give images of a quality not surpassed by any other pickup.

The author considers the adoption of an

819-line standard (880 below) to be a mistake and that, as in Britain, the present standard should be retained for a specified period and then reviewed. Such a policy would allow manufacturers to concentrate on the quick production of large numbers of receivers of present designs.

621.397.5 873
Experimental Equipment for 729-line Television—J. L. Delvaux. (*Onde Élec.*, vol. 28, pp. 370–372; October, 1948.) Summary only. Reasons for the choice of 729 lines are given, with some particulars of equipment developed by the C.F.T.H.

621.397.5 874
Television and Its Industrial Outlook To-Day—P. Grivet. (*Onde Élec.*, vol. 28, pp. 381–383; October, 1948.) Summary only.

621.397.5:535.88:532.62 875
Large-Screen Television and the Eidophor Process—H. Thiemann. (*Onde Élec.*, vol. 28, pp. 409–411; October, 1948.) Summary only. For a full account of this process see 296 of 1948 and back references.

621.397.5:778.5 876
Numerical Values of the Definition of Cinematograph Films—Comparison with Television—J. L. Delvaux. (*Onde Élec.*, vol. 28, pp. 369–370; October, 1948.) Summary only.

621.397.5:778.53 877
Methods of Cinematographic Recording Using Television—Y. L. Delbord. (*Onde Élec.*, vol. 28, pp. 366–368; October, 1948.) Summary only. Methods with intermittent and with continuous motion of the film are discussed and an arrangement suitable for recording in color is outlined.

621.397.5:791.45 878
The Relations between Television and the Cinema—S. Mallein. (*Onde Élec.*, vol. 28, pp. 350–352; October, 1948.) Summary only.

621.397.5(083.74)+621.397.331.2 879
Proposals for the Standardization of Television in Italy, and New Electronic Generator for Television Synchronization—A. V. Castellani. (*Onde Élec.*, vol. 28, pp. 357–358; October, 1948.) Summary only. A system is proposed with transmission for about 1,200 lines interlaced, and reception either on 1,200 lines interlaced for large-screen receivers, with the quality of 35-mm film, or on 600 lines noninterlaced for home direct-viewing receivers.

A network of transmitting stations, mainly alpine and telecontrolled, would be connected with the producing centers by radio links, a single central station controlling the synchronization signals. A generator has been produced in which interlacing, stable and independent of supply-voltage variations, can be obtained with an even number of lines, while control is easy and operation certain.

621.397.5(083.74) 880
France has decided on her New Television Standard—Y. A. (*Onde Élec.*, vol. 28, p. 460; December, 1948.) Summary only. The number of lines is to be 819, a compromise between very high definition (about 1,000 lines) and medium definition (400 to 600 lines). The frequency band is to be 174 to 216 Mc and channel bandwidth 14 Mc. See also 872 above.

621.397.5(42) 881
British Television—(*Electronic Eng.* (London), vol. 20, September, 1948. Supplement.) An historical introduction and a brief illustrated discussion of the BBC service. See also 891 below.

621.397.5.09 882
Measurement of Phase Constants and Group Propagation Times in Television—A. P. A. Fromageot. (*Onde Élec.*, vol. 28, pp. 379–380; October, 1948.) Summary only. A quadripole ensures perfect transmission only when its attenuation is independent of the frequency and its phase constant is proportional to the frequency. The latter condition implies that group propagation time should be constant. Methods of measurement for the two quantities concerned are discussed.

621.397.6:621.316.345 883
TV Control Console Design—J. Ruston. (*Communications*, vol. 28, pp. 8–11, 33; October, 1948.) Discussion of master control requirements, video signal monitoring, bridging and terminating connections, modulation measurements, circuits, and mechanical layout.

621.397.61:621.396.619 884
Modulation of Television Transmitters and Cathode Excitation—H. H. Ernyei. (*Onde Élec.*, vol. 28, pp. 373–375; October, 1948.) Summary only. Grid modulation and cathode excitation are discussed with the help of equivalent circuits, which are again used in considering neutrodynes. A new equivalent scheme is given.

621.397.62 885
Postwar Television Receiver Design—D. W. Pugsley. (*Trans. AIEE*, vol. 66, pp. 453–458; 1947.) Illustrated discussion of recent improvements for both direct-view and projection receivers.

621.397.62:535.88 886
Large-Screen Projection of Television Images—A. Cazalas. (*Onde Élec.*, vol. 28, pp. 361–363; October, 1948.) Summary only. The characteristics of three RCA projection systems and of that of the Compagnie des Compteurs are tabulated and discussed. See also 1797 of 1948 and 238 of February (Maloff).

621.397.645 887
Pentriode Amplifiers—Zeidler and Noe. (*See* 682.)

621.397.645:621.385.4 888
Duplex Tetrode UHF Power Tubes—Smith and Hegbar. (*See* 906.)

621.397.7 889
Television Equipment for Broadcast Stations—W. L. Lawrence. (*Trans. AIEE*, vol. 66, pp. 443–452; 1947.) Description, with photographs and block diagrams, of a complete studio and transmitting equipment.

621.397.7 890
Some Problems arising from the Working of a Television Centre—H. Delaby. (*Onde Élec.*, vol. 28, pp. 364–365; October, 1948.) Summary only. Discussion of various technical and production problems. Financial questions are not considered.

621.397.7(42) 891
Television Development in Britain—(*Nature* (London), vol. 162, pp. 427–428; September 18, 1948.) The Television Advisory Committee (appointed on the recommendation of the Hankey Television Committee, summaries of whose report were noted in 3002, 3584, and 3585 of 1945) considers that possible improvements in quality are too slight to justify any change of the existing 405-line standard for several years. The London television station will, therefore, continue to operate on this standard. The Midlands station at Sutton Coldfield will do likewise; it should be in operation at the end of 1949, with a carrier frequency of about 60 Mc, antenna system on a mast 750 feet high, and a reception range of approximately 50 miles. Higher power than at Alexandra Palace will be used for both sound and vision carriers. Radio and cable links between London and Sutton Coldfield are being provided. A further station in the north of England is contemplated.

621.397.743:621.396.65.029.64 892
A New Microwave Television System—Wentz and Smith. (*See* 851.)

TRANSMISSION

621.396.61 893
A High-Level Single-Sideband Transmitter—O. G. Villard, Jr. (Proc. I.R.E., vol. 36, pp. 1419–1425; November, 1948.) Two high-power balanced modulators, biased to cutoff in the absence of an audio input and using tubes which behave substantially as constant-current sources, may be connected to a common tank circuit for the generation of single-sideband signals by the phase-rotation method. The efficiency obtainable with this arrangement approximates to that of a conventional linear amplifier. Simplicity, ease of adjustment, and power economy make this circuit suitable for applications where a certain amount of distortion and undesired sideband output can be tolerated.

621.396.619:621.396.615.141.2 894
Wide-Band Modulation with Magnetron—H. Gutton and J. Ortusi. (*Onde Élec.*, vol. 28, pp. 384–387; October, 1948.) Summary only. A new method of modulation, suitable for dm and cm waves, is based on the introduction, into the output waveguide, of an impedance which can be varied from 0 to ∞ by using the impedance variations of a whole-plate or a cavity magnetron. In the latter case, with negative reaction between the successive cavities, a modulation bandwidth 10 per cent of the carrier frequency has been achieved. The method can be used for carrier powers of several kw, though the video amplifier power need not exceed 1 watt.

621.396.8 895
Spectra of Quantized Signals—W. R. Bennett. (*Bell Sys. Tech. Jour.*, vol. 27, pp. 446–472; July, 1948.) To determine the number of quantized steps required to transmit a specific type of signal, the relation between distortion and step size must be found. Quantizing of magnitude only is discussed mathematically in terms of a "perfect step transducer". Combined quantizing of magnitude and time is discussed in terms of a theory of periodic sampling of signals. Results are shown graphically. Some experimental results obtained with a laboratory model of a quantizer are included.

VACUUM TUBES AND THERMIONICS

621.383 896
Concerning the Use of a 920 Double Photo-Cell in a Current Amplifier and Stabilizer—B. M. Banerjee and S. K. Sen. (*Indian Jour. Phys.*, vol. 22, pp. 43–50; January, 1948.) In a current-stabilizer circuit, a gas-filled double photo cell requires careful adjustment and is less sensitive than two separate photocells or a vacuum double photo cell.

621.383:621.396.822 897
Noise in Vacuum Phototubes at High Current Levels—R. F. Morrison. (*Electronics*, vol. 21, pp. 126, 168; November, 1948.) Measurements of output noise indicate that, for cathode currents from 50 μamp to 1 mamp, shot effect is the only important source of noise.

621.383:621.397.5 898
Electron-Multiplier Cell—Its Use in Television—A. Lallemand. (*Onde Élec.*, vol. 28, pp. 394–395; October, 1948.) Summary only. Multiplier cells are normally produced with 7 stages, rarely with 12, though production difficulties for 12 stages are little greater than for 7. Investigations have shown that with a 12-stage cell having an Sb-Cs photocathode, for a source at 2400°K, a photocurrent is obtained equal to the dark current for a flux of 10^{-9} lumen. Such a cell is particularly suited for responding to the

long waves from a cathode-ray tube. Its dark current is so small that cooling is quite unnecessary.

621.383.4 899
Lead Selenide Photo-Conductive Cells—J. Starkiewicz. (*Jour. Opt. Soc. Amer.*, vol. 38, p. 481; May, 1948.) Thin layers of PbSe can be activated in vacuo so that they have marked photoconductive sensitivity at room temperature. The process is analogous to that for PbS noted in 3709 of 1947 (Sosnowski, Starkiewicz, and Simpson) and 443 of 1948 (Sosnowski), but is complicated by the fact that Se, formed during activation, tends to remain in the layer.

621.383.5 900
The Efficiency of the Barrier Layer Photo-Cell—R. A. Houstoun. (*Phil. Mag.*, vol. 39, pp. 902–910; November, 1948.) The highest efficiency obtained for approximately monochromatic light for 5 commercial photo cells was 6.4×10^{-3}, and for white light 6.2×10^{-3}, in contrast to manufacturers' claims of 50 per cent.

621.385 901
Radio Valve Practice—(*Electronic Eng.* (London , vol. 20, pp. 321–324; October, 1948.) Long summary of British Radio Valve Manufacturers' Association recommendations for obtaining optimum performance. The importance of consulting the manufacturer before including tubes in unusual circuits is emphasized.

621.385:513.761.5 902
On the Similitude of Valves—A. Martinot-Lagarde. (*Onde Élec.*, vol. 28, pp. 440–444; November, 1948.) A general method of treatment is presented which is essentially based on Vaschy's demonstration of the central theorem of dimensional analysis. The method is applied to discussion of output, amplification, plate current, cathode emission, etc., as functions of frequency, tube dimensions, and applied voltages.

621.385:621.397.61 903
Theoretical Methods of Study and Recent Realizations of Transmitting Valves for Television—G. Lehmann (*Onde Élec.*, vol. 28, pp. 398–401; October, 1948.) Summary only. The special features of some triodes, tetrodes, and pentodes for powers up to 200 kw and frequencies to 2,000 Mc are briefly considered. Design methods based on dimensional analysis (3821 of 1946) can take account of transit-time and space-charge effects. Fundamental relations for triodes are tabulated. For tubes operating continuously, or nearly so, at frequencies below 600 Mc, the principal limiting factor is the plate dissipation, while above 600 Mc it is the cathode emission. Suitable design formulas for these two cases are given.

621.385.029.64:621.397.6 904
On the Help which Some Recent Ideas Concerning U.H.F. Valves Can Give in Television—R. Warnecke and P. Guénard. (*Onde Élec.*, vol. 28, pp. 417–420; October, 1948.) Summary only. Discussion, with special reference to television, of (a) high-power triodes and tetrodes, (b) v.m. tubes, (c) magnetrons, (d) klystrons, and (e) distributed amplification.

621.385.1:621.397.61 905
Some Recent Transmitting Valves for Television—J. Becquemont. (*Onde Élec.*, vol. 28, pp. 355–357; October, 1948.) Summary only. Discussion of modern constructional developments, with particulars of tubes with output of 500 watts at 600 Mc.

621.385.4:621.397.645 906
Duplex Tetrode U.H.F. Power Tubes—P. T. Smith and H. R. Hegbar. (Proc. I.R.E., vol. 36, pp. 1348–1353; November, 1948.) The design and development of wide-band power tubes are considered, with particular reference to television applications. Methods of obtaining the required performance are discussed qualitatively. A 5-kw 300-Mc liquid-cooled internally neutralized double tetrode is described.

621.385.832:621.318.572 907
Electrostatically Focused Radial-Beam Tube—A. M. Skellett. (Proc. I.R.E., vol. 36, pp. 1354–1357; November, 1948.) Discussion of the principle of operation and the construction of a tube in which the electron beam is focused and rotated by means of internal electrostatic fields. The tube has a single cathode with twelve control-grid and plate elements fixed in a circle around it. It thus forms an electronic distributor with many applications as a high-speed switch. See also 3167 of 1944.

621.396.615.141.2 908
High-Power Interdigital Magnetrons—J. F. Hull and A. W. Randals. (Proc. I.R.E., vol. 36, pp. 1357–1363; November, 1948.) The operation of a pillbox-cavity interdigital magnetron in the cavity mode is described. Stability of operation has been achieved by the addition of cathode decoupling chokes. Continuous outputs up to 500 watts with an efficiency of 70 per cent have been obtained with $\lambda \simeq 10$ cm.

621.396.615.141.2 909
The Rising-Sun Magnetron—S. Millman and A. T. Nordsieck. (*Jour. Appl. Phys.*, vol. 19, pp. 156–165; February, 1948.) Full paper: summary noted in 296 of 1947. See also 293 of 1947 (Fisk, Hagstrum, and Hartman).

621.396.615.141.2 910
The Resonant Modes of the Rising-Sun and Other Unstrapped Magnetron Anode Blocks—N. M. Kroll and W. E. Lamb, Jr. (*Jour. Appl Phys.*, vol. 19, pp. 166–186; February, 1948.) Full paper: summary noted in 297 of 1947.

621.396.615.142 911
Transit Time Effects in Output Fields—T.S. Popham. (*Wireless Eng.*, vol. 25, pp. 353–360; November, 1948.) An analysis of the mechanism whereby energy is transferred from a modulated electron beam to an electric field by means of a pair of electrodes. The oscillatory-field energy produced at large transit angles remains comparable with that produced at small angles. The effect of secondary electron radiation from one electrode is discussed and a critical value of transit angle is derived above which there is a considerable increase in the suppression of secondary radiation.

621.396.645:537.311.33:621.315.59 912
Characteristics of Amplifying Crystals—S. Y. White. (*Audio Eng.*, vol. 32, pp. 18–19; November, 1948.) Discussion of design and performance characteristics of oscillators incorporating the transistor. See also 913 to 916 below and back references.

621.396.645:537.311.33:621.315.59 913
Clarification of Germanium Triode Characteristics—S. Y. White. (*Audio Eng.*, vol. 32, pp. 19–21, 44; December, 1948.) A review of present information. It is not yet possible to produce Ge triodes in quantity at a competitive price, and they are unlikely yet to displace ordinary tubes. Potential advantages and present limitations are considered. See also 265 of February, 581, 582 of March, and 912 above.

621.396.645:537.311.33:621.315.59 914
The Transistor, or the Return of the Crystal—L. Chrétien. (*TSF Pour Tous*, vol. 24, pp. 260–262; October, 1948.) An account based on that given in *Electronics*, vol. 21, pp. 68–71; September, 1948. (D. G. F. and F. H. R.). See also 582 of March (White) and back references.

621.396.645:537.311.33:621.315.59 915
Transistor: the Crystal Amplifier and Oscillator—J. P. Arnaud. (*Rev. Telegr.* (Buenos Aires), vol. 37, pp. 715–720; October, 1948.) A general description, with theory, given in a lecture to the Buenos Aires section of the IRE. References are given to all pertinent publications to date. See also 582 of March (White) and back references.

621.396.828.1 916
Reducing Hum in Pentodes—I. Zakarias. (*Electronics*, vol. 21, pp. 170, 178; November, 1948.) The operating condition of a pentode affects the relative hum current appearing in the plate circuit; this hum current can sometimes be made to vary about zero magnitude. Experimental curves showing relative hum current for various pentodes and operating conditions are included.

621.396.615.141.2:621.396.619.23 917
Pulse Generators [Book Review]—G. N. Glasoe and J. V. Lebacqz. McGraw-Hill, New York, N. Y. and London, 1948, 722 pp., $9. (Proc. I.R.E., vol. 36, p. 1396; November, 1958.) Volume 5 of the Radiation Laboratory series. The book might have been entitled "Magnetron Modulators". The various types of pulse generator used with magnetrons, and their associated components, are discussed in detail. For another review see *Nature* (London), vol. 162, p. 754; November 13, 1948.

MISCELLANEOUS

05:550.38 918
Geomagnetic and Geoelectric Literature in Two New German Periodicals—H. G. Macht. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 169–171; June, 1948.) *Zietschrift fur Meteorologie* is a new German journal which has replaced *Met. Z. Deutsche Hydrographische Zeitschrift* has also replaced *Annalen der Hydrographie und Maritimen Meteorologie.* Brief details of editorial organization of these two journals are given, with abstracts of selected articles. See also 705, 712, and 714 above.

061.3:01 919
The Royal Society Scientific Information Conference, 21 June–2 July 1948: Report and Papers Submitted. [Book Notice]—The Royal Society, London, 723 pp., 25s. Full report of the Conference noted in 590 of March.

061.3:[621.395.06+620.193 920
C.C.I.(F.) Meeting, Stockholm, June 1948—(*P.O. Elec. Eng. Jour.*, vol. 41, part 3, p. 154; October, 1948.) A brief review of the discussions on the European switching plan and on cable corrosion.

061.3:621.396 921
Fifth Plenary Meeting of the C.C.I.R., Stockholm, July, 1948—(*P.O. Elec. Eng. Jour.*, vol. 41, part 3, pp. 155–156; October, 1948.) Titles of 13 international study groups are listed. Recommendations are very briefly discussed.

621.3 922
Fundamentals of Electrical Engineering [Book Review]—V. P. Hessler and J. J. Carey. McGraw-Hill, London, 241 pp., 45s. (*Wireless Eng*, vol. 25, pp. 370–371; November, 1948.) A background of electricity and magnetism from a physics course is assumed. The book deals solely with fundamentals, and there is no mention of electrical machines or thermionic tubes. Alternating current is not considered.

621.396.029.6 923
Radio at Ultra-High Frequencies: Vol. 2 [Book Notice]—A. N. Goldsmith, A. F. Van Dyck, R. S. Burnap, E. T. Dickey, and G. M. K. Baker (Eds). Radio Corporation of America, Princeton, N. J., 1948, 485 pp., $2.50 in United States, $2.70 elsewhere. Covers the period 1940 to 1947. Summaries of the papers contained in vol. 1, (1930 to 1939) are also included. Many of the papers have already appeared in *RCA Rev.* and other publications; here they are arranged under broad subject headings.

SECTION MEETINGS

Baltimore

"Narrow-Band Automatic Radio Compass," by K. F. Umpleby and D. M. Heller, Bendix Radio Division; January 25, 1949.

Cincinnati

"Printed Electronic Circuits," by R. L. Henry National Bureau of Standards; January 18, 1949.

Columbus

"Antenna for Television," by D. C. Cleckner and A. Joost, Antenna Research Laboratories, Inc.; January 12, 1949.

"Electric Coil and Insulation Manufacturing," by D. Stafford, National Electric Coil Company; January 21, 1949.

Connecticut Valley

"Transitors," by J. N. Shive, Bell Telephone Laboratories; January 20, 1949.

Dallas-Fort Worth

Election of Officers; January 11, 1949.

Dayton

"The Electron Wave Tube," by A. V. Haeff, Naval Research Laboratory; February 10, 1949.

Detroit

"Applications of Subminiature Tubes," by R. K. McClintock, Sylvania Electric Products Inc.; January 21, 1949.

London (Ontario)

"Sound Waves, Their Properties and Their Uses," by T. D. Northwood, National Research Council; January 14, 1949.

Los Angeles

"A New Long-Playing Disk Recording System," by C. A. Boggs, Columbia Records, Inc.; December 14, 1948.

"Double Stream Amplifier," by W. B. Hebenstreit, Hughes Aircraft; January 18, 1949.

"Analogue Computer for Linear Antenna Arrays," by T. T. Taylor, Hughes Aircraft; January 18, 1949.

Montreal

"The Design of FM Equipment for the 160-Mc Band," by F. H. Margolick, Canadian Marconi Company; January 19, 1949.

"Engineers and Canadian Labor Legislation," by F. S. Howes, McGill University; February 8, 1949.

"GCA—Ground-Controlled Approach," by W. E. Ekblaw, Jr., Gilfillan Brothers; February 8, 1949.

New Mexico

"Phase Shift Networks in Single Sideband Transmission," by R. Hoffman, United States Army; January 21, 1949.

North Carolina-Virginia

"Raydist," by C. E. Hastings, Hastings Instrument Company; October 15, 1948.

"Electrical Network Analyzers," by K. R. Spangenburg, Naval Research Laboratory; November 19, 1948.

(Continued on page 36A)

SECTION MEETINGS

(Continued from page 34A)

Ottawa

"The Technique of Television Sound," by R. M. Tanner, Northern Electric Company Ltd.; January 13, 1949.

"Telephone and Telegraph Multiplex on VHF Circuits," by C. B. Fisher, Radio Engineering Products Ltd.; February 3, 1949.

Philadelphia

"Present and Future Trends in Television Receivers," by D. G. Fink, *Electronics;* February 2, 1949.

Rochester

"Radio Propagation above 100 Megacycles," by P. E. Lannan, Stromberg-Carlson Company; January 20, 1949.

Salt Lake

"Brains, Computing Machines, and Electronics," by L. A. Woodbury, University of Utah Medical School; January 24, 1949.

San Francisco

"The Engineer's Role in International Conferences," by R. V. Howard, National Association of Broadcasters; January 5, 1949.

"Engineering Research in the Development of Western Industry," by J. E. Hobson, Stanford University Research Institute; January 27, 1949.

"A Carnival of Measurements," by J. M. Whittenton, General Electric Company; January 31, 1949.

Syracuse

"New Developments," by W. Hausz, J. F. McAllister, and R. F. Shea, General Electric Company; January 6, 1949.

"An Evaluation of the Application of New and Old Techniques to the Improvement of Magnetic Recording Systems," by L. C. Holmes, Stromberg-Carlson Company; February 3, 1949.

Toledo

"Technical Advances in the Reproduction of Motion Picture Sound," by H. L. Neuert, Altec Service Corporation; January 17, 1949.

"House of Magic," by W. Hoverman and R. Verbiliss, General Electric Company; February 7, 1949.

Twin Cities

"The Microwave Dielectrometer," by D. G. Jelatis, Central Research Laboratories, Inc.; January 4, 1949.

SUBSECTIONS

Amarillo-Lubbock

"WBAP-TV Television Installation," by C. E. Houston, Texas Technical University; January 18, 1949.

Election of Officers; January 18, 1949.

Hamilton

"Modern B. C. Directive Arrays," by B. de F. Bayly, Bayly Engineering Company; January 17, 1949.

Northern New Jersey

"Broad-Band Microwave Lenses," by W. E. Kock, Bell Telephone Laboratories, Inc.; January 19, 1949.

"Broad-Band Transmission System," by S. Hopfer, Polytechnic Research and Development Corporation; January 19, 1949.

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Here's the Recorder You asked for!

The best features of Presto's dual motor gear drive *with* the overhead mechanism and turntable of the famous Presto 6-N.

YES, engineers have often asked us for a compact, economical yet high-quality recorder. Now you may have it in the Presto 66-G for standard and microgroove recording.

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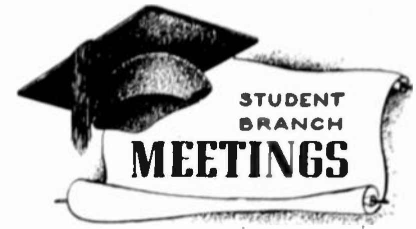


UNIVERSITY OF CALIFORNIA, IRE—AIEE BRANCH

Election of Officers; January 31, 1949.

UNIVERSITY OF FLORIDA, IRE—AIEE BRANCH

"A Generator for Predetermined Wave Forms," by J. D. Wells, Student, University of Florida; February 8, 1949.

IOWA STATE COLLEGE, IRE—AIEE BRANCH

"My Trip to Mexico," by W. L. Cassell, Iowa State College; January 19, 1949.

STATE UNIVERSITY OF IOWA, IRE BRANCH

"Student Talks," by T. Babcock and E. L. Jesse, Students, State University of Iowa; January 19, 1949.

Election of Officers; January 25, 1949.

"Microwaves in Telephony," by C. D. Peebler, Northwestern Bell Telephone Company; February 9, 1949.

LAFAYETTE COLLEGE, IRE-AIEE BRANCH

"Tesla Coils," by R. W. Karcher, Student, Lafayette College; February 10, 1949.

"Little Known Facts in the History of Electricity Up to 1866," by F. Blanchard, Lafayette College; February 10, 1949.

MASSACHUSETTS INSTITUTE OF TECHNOLOGY, IRE-AIEE BRANCH

Election of Officers; January 12, 1949.

MISSOURI SCHOOL OF MINES, IRE-AIEE BRANCH

"Construction of an FM Broadcasting Station," by W. R. Chapin and R. Nickles, Station KWDG; February 2, 1949.

NEW YORK UNIVERSITY, IRE BRANCH

"Color Television," by G. E. Anner, New York University Faculty; January 20, 1949.

NORTH CAROLINA STATE COLLEGE, IRE BRANCH

"Telemetering in Aeronautical Research," by C. A. Taylor, Langley Memorial Aeronautical Laboratory; February 16, 1949.

"Experimental Functions as Applied to Electrical Engineering," by J. W. Cell, North Carolina State College; January 26, 1949.

NORTHWESTERN UNIVERSITY, I.R.E.-AIEE BRANCH

Election of Officers; January 11, 1949.

"Careers in Power Utilities," by M. C. Kimberly, Commonwealth Edison Company; January 25, 1949.

OREGON STATE COLLEGE, IRE-AIEE BRANCH

"High-Voltage Insulators," by C. R. Kingsbury, Ohio Brass Company; January 27, 1949.

PURDUE UNIVERSITY, IRE BRANCH

"Intercarrier Sound Systems for Television," by S. W. Seeley, Radio Corporation of America; January 28, 1949.

RUTGERS UNIVERSITY, IRE-AIEE BRANCH

"Advantages of Prize Paper Competition," by P. S. Creager, Rutgers University; February 8, 1949.

THE UNIVERSITY OF TENNESSEE, IRE BRANCH

"High-Fidelity Audio Systems," by C. M. McCracken, University of Tennessee Faculty; January 18, 1949.

"Your Telephone Voice of the Future," by L. H. Calloway, Southern Bell Telephone and Telegraph Company; February 1, 1949.

UTAH STATE AGRICULTURAL COLLEGE, IRE BRANCH

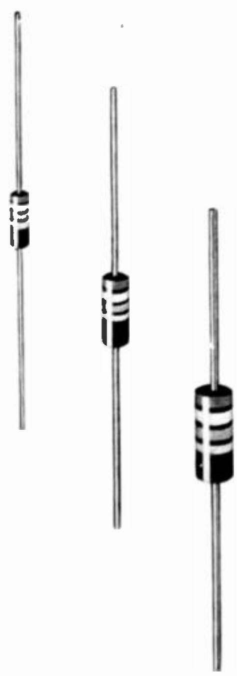
"The Electron Microscope," by G. Cochran, Utah State Agricultural College; February 9, 1949.

A NEW ADDITION TO THE ALLEN-BRADLEY PLANT FOR PRODUCING A-B FIXED AND ADJUSTABLE RESISTORS

A REPORT to the Radio Industry about Allen-Bradley Radio Resistors

Radio manufacturers have discovered that fixed resistors of run-of-mine quality will not meet the requirements of television circuits.

This situation has, overnight, created an unprecedented demand for the top quality and stability found in Allen-Bradley fixed resistors. In spite of weekly shipments of many millions of Bradleyunits in ½-watt, 1-watt, and 2-watt ratings, the current demand still far exceeds the capacity of the Allen-Bradley radio resistor department . . . and our customers are unhappy with our resistor deliveries.



But a large addition to the Allen-Bradley factory is under way. Much additional resistor production machinery is under construction. However, it will take time before these facilities will be ready for the increased production of Bradleyunits.

Meantime, Allen-Bradley is working twenty-four hours per day—seven days per week—to produce the maximum output of Bradleyunits. It is physically impossible to do more at this time.

During this stringency, we appeal to the radio industry to restrict its use of our products to applications in which Bradleyunit quality performance is absolutely essential.

A-B Quality

ALLEN-BRADLEY COMPANY, 114 W. GREENFIELD AVE., MILWAUKEE 4, WISCONSIN

MEMBERSHIP

The following transfers and admissions were approved and will be affective as of April 1, 1949:

Transfer to Senior Member

Alter, A. R., 1524 68 Ave., Philadelphia 26, Pa.
Boyd, L. K., 1206 Wilson Ave., Santuree, Puerto Rico
Disney, V. H., 2295 Summit, Columbus, Ohio
Fernandez, M., Headquarters Military Air Transport Service, Washington, D. C.
Inglis, A. F., 161 Danbury St., S. W., Washington, D. C.
Jackson, C. H., 31 N. Granada St., Arlington, Va.
Miller, C. E., 1132 Melrose Ave., Glendale 2, Calif.
Overton, B. H., Box 115, Shalimar, Fla.
Schuck, O. H., 4711 DuPont Ave., S., Minneapolis, Minn.
Spencer, B. F., 120 Wilson St., Garden City, L. I., N. Y.
Vogelman, J. H., 676 Westwood Ave., West End, N. J.

Admission to Senior Member

Harder, E. L., 1204 Milton Ave., Pittsburgh 18, Pa.
Kirchenbauer, C. C., 2907 Powhattan Dr., Toledo, Ohio
Liberman, A., 1512 S. Pulaski Rd., Chicago 23, Ill.
Power, D. W., 16 Inwood Rd., Summit, N. J.
Steinberg, E. B., 639 Summer St., Stamford, Conn

Transfer to Member

Auerbach, I. L., 2106 Delancey St., Philadelphia 3 Pa.
Beach, O. R., 1869 W. 38 Pl., Los Angeles 37, Calif.
Bullock, G. M., 21 Danbury St., S. W., Washington 20, D. C.
Clyne, W. E., 521 New Customhouse, Denver 2, Col.
Dakin, O. C., 240 Sutherland Dr., Toronto, Ont., Canada
Fernane, J., Federal Communications Commission, 838 U. S. Court House, Kansas City 6, Mo.
Fleming, J. W., 80 Myrtle Ave., Edgewater, N. J.
Furst, R. E., 520 Cornelia Ave., Chicago 13, Ill.
Gable, N. P., 40–42 Lawrence St., Flushing, L. I., N. Y.
Greenbaum, W. H., 17 Poplar St., Elmsford, N. Y.
Howe, W. E. W., 3940 Second St., Washington 20, D. C.
Hyder, H. R., III, c/o Department 63, Bendix Radio Division, Baltimore 4, Md.
Levin, K. L., 316 Fairfield Ave., Kenmore 17, N. Y.
Mills, C. A., 30 W. Oakland Ave., Oaklyn, N. J.
Muschamp, R. A., 103 Sandra Lane, North Syracuse, N. Y.
Rich, A. P., 1855 Dome Ct., San Pedro, Claif.
Russell, W. W., 146 Poplar Ave., Elmhurst, Ill.
Schwarz, R. J., Dept. of Electrical Engineering, Columbia University, New York 27, N. Y.
Sherbin, L. E., 1112 Gerard Ave., New York, N. Y.
Stephenson, H. B., 102 Byron Ave., Buffalo 17, N. Y.
Trout, C. C., 528 N. Eastern Ave., Indianapolis 1, Ind.
Wilson, L. W., 4015 Front St., San Diego 3, Cal.
Wyman, R. C., 5711 31 Ave., Hyattsville, Md.

Admission to Member

Ahmed, M. A. H., Electrical Dept., Faculty of Engineering, Cairo, Giza, Egypt
Boutros, O. L., 1629 E. 31 St., Kansas City 3, Mo.

(Continued on page 42A)

MEMBERSHIP

(Continued from page 40A)

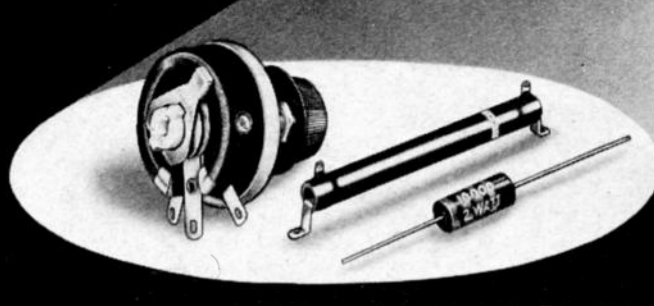
Burch, E. C., 5544 Riggs, Overland Park, Kan.
Daubenheyer, H. W., 6817 Barr Rd., Washington 16, D. C.
Dickinson, R. H., 902 Albany St., Schenectady 7, N. Y.
Dinardo, J. A., 107 W. Sixth St., Emporium, Pa.
Everts, W. J., Chittenango, R. D. 1, N. Y.
Fonda, J. C., 3132 Tyson Ave., Philadelphia 24, Pa.
Hayes, H. J., Jr., Dept. of Electrical Engineering, Engineering Bldg., Princeton University, Princeton, N. J.
Herz, A. J., 22 W. Monroe St., Chicago 3, Ill.
Jefferson, W. E., 302 South St., Halifax, Nova Scotia
Joga Rao, H., No. 8 Murugesa Mudali St., T. Nagar, Madras South India
Kolkedy, J. G., 441 Mellon St., S. E., Washington 20, D. C.
Lavender, R. W., 81 Beethoven St., Binghamton, N. Y.
Mudaliyar, K. N. D., Engg. Supr. Carr., 6 Nataraja Pillai St., Kaveripak, North Arcot, South India
Razdowitz, A., 593 Riverside Dr., New York 31, N. Y.
Rorholt, B. A., E.R.L., Vanserg Bldg., 12 Divinity Pl., Cambridge, Mass.
Rutherford, C. R., Defense Research Laboratory, Box 1, University Station, Austin, Tex.
Skolnik, B. J., 1450 Veteran Ave., West Los Angeles Calif.
Skorup, G. E., 49 Manor Ave., Oaklyn, N. J.
Waters, W. F., 801 Telephone Bldg., Dallas 2, Tex.
Witow, M. L., 130 Cass St., Dayton 2, Ohio

The following admissions to Associate grade were approved and will be effective as of March 1, 1949:

Alcorn, C. L., 25 Arnold Pl., Dayton 7, Ohio
Aycock, E. H., 225 N. Mayo St., Rocky Mount, N. C
Bajaj, P. L., The Vithal Nivas, Block A No. 34, Jacob Circle, Bombay 11, India
Banham, J. C., Gransha Rd., Ballybeen, Dundonald, Belfast, N. Ireland
Barnsdale, E. A., The Risley Club, Risley, Warrington, Lancs., England
Bechtold, H. P., Jr., R.F.D. 365A, John St. Wantagh, N. Y.
Belz, A. R., Herb Lew Rd., Warrington, Pa.
Benda, D., 6804 S. Halsted St., Chicago 21, Ill.
Biagi, A. D., 956 E. 25 St., Paterson 3, N. J.
Blohm, R. W., 4149 N. Bernard St., Chicago 18, Ill.
Bloom, F. J., 10 Hartley Ave., Mount Vernon, N. Y.
Bowman, R. B., KPAN, Hereford, Tex.
Boyd, T. J., Jr., 4213 Oglethrope St., Hyattsville, Md.
Bradley, J. L., 215 N. Bowling Green Way, Los Angeles 24, Calif.
Brody, H., 355 Stockton St., Brooklyn 6, N. Y.
Brogan, J. M., 396 Union St., Jersey City, N. J.
Bruun, D. R., B.O.Q.-B, Naval Ordnance Test Station, China Lake, Calif.
Burger, F. J., 89 Colonial Pkwy., Manhasset, L. I., N. Y.
Carr, R. W., 640 Briar Pl., Chicago 14, Ill.
Castaneda, R. A., 446 W. 38 St., New York 18, N. Y.
Chadney, H. B., 1066 Evergreen Ct., New Milford, N. J.
Chandler, C. F., 29A Caselli Ave., San Francisco 14, Calif.
Chittenden, K. M., 341 Elm St., Oxnard, Calif.

(Continued on page 44A)

Your electronic equipment is no better than its smallest component . . .



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Here is the most extensive line of rheostats offered today . . . 10 sizes, from 25 to 1000 watts, with many resistance values. All-ceramic construction. Windings are locked in vitreous enamel.

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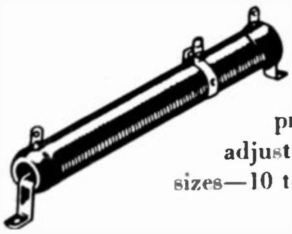
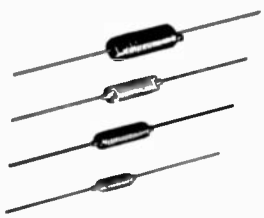
Resistors are wire wound on a ceramic core, rigidly held in place, insulated, and protected by vitreous enamel. Even winding dissipates heat rapidly —prevents hot spots. Many types, in ratings from 5 to 200 watts.

DIVIDOHM ADJUSTABLE RESISTORS



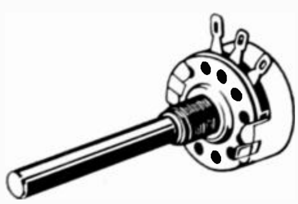
Used as multi-tap resistors or voltage dividers. Narrow strip of exposed winding provides contact surface for the adjustable lug. Available in seven sizes—10 to 200 watts.

RADIO FREQUENCY PLATE CHOKES



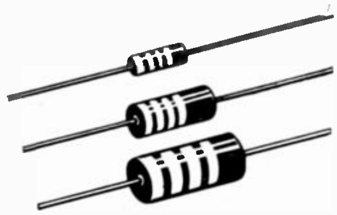
Single-layer wound on low power factor steatite or molded plastic cores. Seven stock sizes cover range 3 to 520 mc. Two units rated 600 ma; all others 1000 ma.

* MOLDED COMPOSITION POTENTIOMETER



A 2-watt molded composition unit with good margin of safety. It is unaffected by heat, cold, or moisture. Resistance element is a thick, solid-molded ring.

* MOLDED COMPOSITION RESISTORS



Small and sturdy, these "Little Devil" units come in ½, 1, and 2-watt sizes. 10 Ohms to 22 megohms. Tol. ± 10% and ± 5%.

** So that two exceptionally high-quality products will be universally obtainable, Ohmite Manufacturing Company, in co-operation with the Allen-Bradley Company, has arranged for the Type AB (Allen-Bradley Type J) control and Little Devil Molded Composition Resistors (Allen-Bradley Types EB, GB, and HB) to be available from stock at Ohmite distributors.*

Write for Catalog No. 40

OHMITE MANUFACTURING CO.
4860 Flournoy St., Chicago 44, Ill.

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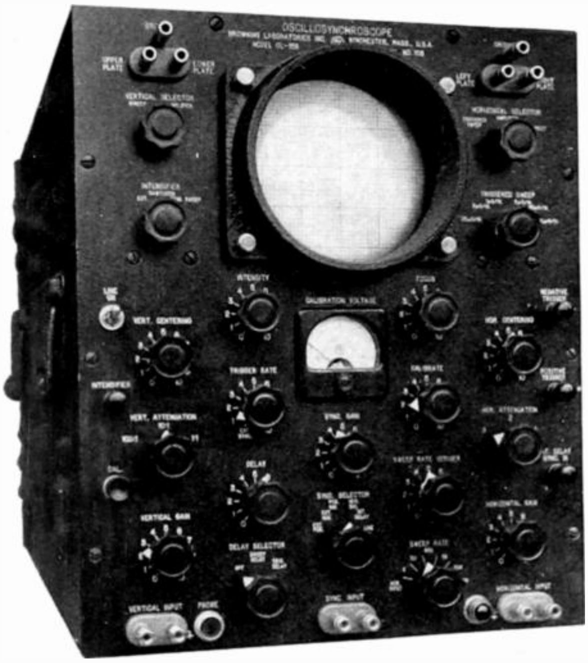
(Continued on page 46A)

Browning

OSCILLOSYNCHROSCOPE
MODEL OL-15B



Combining the functions of

OSCILLOSCOPE and SYNCROSCOPE

An Outstandingly Versatile Instrument

Applicable to—

- **TELEVISION**
- **FACSIMILE**
- **PULSE MODULATION**
- **RADAR**
- **NUCLEAR PHYSICS**
- **COMMUNICATIONS**

GENERAL FEATURES

Five-inch cathode ray tube operating at 4,000 volts accelerating potential. Ordinarily supplied with Pl phosphor, others available on special order.

Vertical amplifier flat within 3 db. from 5 cycles to 6 megacycles. One inch deflection with .05-volt RMS input.

Horizontal amplifier flat within 1 db. from 5 cycles to 1 megacycle.

Built-in calibrating system for determining wave amplitude. No external meter needed.

Deflection plates and intensity grid available directly at front panel terminals.

No waiting for trace to reappear after adjusting gain or applying DC component to input.

Low capacitance, high impedance probe supplied for minimizing test circuit disturbance.

Reasonably symmetrical waves permit full screen vertical deflection.

Contained in single cabinet, weighs less than 100 pounds.

AS AN OSCILLOSCOPE

Linear sawtooth sweeps continuously variable from 5 to 500,000 per second in conjunction with the excellent vertical amplifier outlined. Permits observation of RF waves and envelopes to above 6 megacycles. Because of the extended ranges of the amplifiers and sweep generator, oscilloscopic capabilities are correspondingly increased over standard oscilloscopes.

AS A SYNCHROSCOPE

An internal trigger generator continuously variable from 200 to 5,000 cycles can be used to excite external equipment as well as the sweeps. The trigger can be made by panel control to lead or lag the start of the sweep by amounts up to 1,000 microseconds, making it possible to phase any part of a pulse or transient onto the screen for measurement. Sweep speeds of $\frac{1}{4}$, $\frac{1}{2}$, 1, 5, 20, and 200 microseconds per inch provide convenient image time expansion for detailed observation. As the sweep generator will sweep once for each incoming pulse, single transients or pulses occurring at irregular intervals can be observed or photographed.

For more detailed information write for descriptive Bulletin RO-449

• COMPANION INSTRUMENTS •

SWEEP CALIBRATOR MODEL GL-22

For accurately calibrating sweeps. Markers are provided at 1/10, $\frac{1}{2}$, 1, 10, and 100 microsecond intervals which may be applied as deflection or as intensity modulation. May be triggered directly from OL-15B. Write for bulletin RC-449.

FAIRCHILD OSCILLO-RECORD CAMERA

For permanent records of waveforms on 35 mm. film. Single frames or variable continuous motion permit recording of all phenomena. Various lenses, magazines, etc. available. Easily set up with OL-15B. Write for bulletin RF-449.

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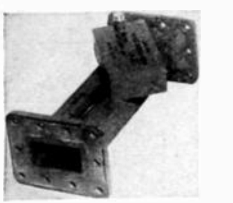
MICROWAVE PLUMBING

10 CENTIMETER



WAVEGUIDE DIRECTIONAL COUPLER, 27 db. Navy type CABV-47AAN, with 4 in. slotted section as shown $42.50
SQ. FLANGE to rd choke adapter, 18 in. long OA 1½ in. x 3 in. guide, type "N" output and sampling probe $32.00
"S" BAND CRYSTAL MOUNT, gold plated, with 2 type "N" connectors $12.50
POWER SPLITTER: 726 Klystron input, dual "N" output $5.00
MAGNETRON TO WAVEGUIDE coupler with 721-A duplexer cavity, gold plated $27.50
10 CM WAVEGUIDE SWITCHING UNIT, switches 1 input to any of 3 outputs. Standard 1½" x 3" guide with square flanges. Complete with 115 vac or dc arranged switching motor. Mfg. Raytheon, CRP 24AAS. New and complete $150.00
10 CM END-FIRE ARRAY POLYRODS .... $1.75 ea.
"S" BAND Mixer Assembly, with crystal mount, pick up loop, tunable output $3.00
721-A TR CAVITY WITH TUBE. Complete with tuning plungers $12.50
10 CM. McNALLY CAVITY Type SG $3.50
WAVEGUIDE SECTION, MC 445A. rt. angle bend. 5½" ft. OA. 8" slotted section $21.00
10 CM OSC. PICKUP LOOP, with male Homedell output $2.00
10 CM DIPOLE WITH REFLECTOR in lucite ball, with type "N" or Sperry fitting $4.50
10 CM FEEDBACK DIPOLE ANTENNA, in lucite ball, for use with parabola $8.00

⅞" RIGID COAX—⅜" I.C.

RIGHT ANGLE BEND, with flexible coax output pick-up loop $8.00
SHORT RIGHT ANGLE bend, with pressurizing nipple $3.00
RIGID COAX to flex coax connector $3.50
STUB-SUPPORTED RIGID COAX, gold plated 5' lengths. Per length $5.00
RT. ANGLES for above $2.50
RT. ANGLE BEND 15" L. OA $3.50
FLEXIBLE SECTION. 15" L. Male to female $4.25
MAGNETRON COUPLING to ⅞" rigid coax, with TR pickup loop, gold plated $7.50

⅞" RIGID COAX—¼" I.C.

⅞" RIGID COAX. Bead Supported ....$1.20 per ft.
SHORT RIGHT ANGLE BEND $2.50
Rotating joint, with deck mounting $15.00
RIGID COAX slotted section CU-60/AP $5.00

1.25 CENTIMETER

"K" BAND FEEDBACK TO PARABOLA HORN, with pressurized window $30.00
MITRED ELBOW cover to cover $4.00
TR/ATR SECTION choke to cover $4.00
FLEXIBLE SECTION 1" choke to choke $5.00
ADAPTER, rd. cover to sq. cover $5.00
MITRED ELBOW and S sections choke to cover ..$4.50
PICKUP Loop. Type "N" output 10 cm $2.75
TR Box Pick Up Loop 10 cm. $1.25

****

MICROWAVE TEST EQUIPMENT

W. E. I 138. Signal generator, 2700 to 2900 Mc range. Lighthouse tube osciUator with attenuator & output meter. 115 VAC input, reg. Pwr. supply. With circuit diagram $50.00
3 cm. Wavemeter: 9200 to 11,000 mc transmission type with square flanges $15.00
3 cm. stabilizer cavity. transmission type $20.00
3 cm. Wavemeter, Micrometer head mounted on X-Band guide. Freq. range approx. 7900 to 10,000 Mc. $78.00
THERMISTOR BRIDGE: Power meter I-203-A. 10 cm. mfg. W.E. Complete with meter, interpolation chart, portable carrying case as shown $72.50
TS 12/AP VSWR METER. Consisting of Slotted Line, and all additional fittings and ampl. for 3CM operation. New and complete $450.00
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****

MICROWAVE GENERATORS

AN/APS-15A "X" Band compl. RF head and modulator. incl. 725-A magnetron and magnet, two 723A/B klystrons (local osc. & beacon), 1B24 TR, rcvr-ampl, duplexer, HV supply, blower, pulse xtmr. Peak Pwr. Out: 45 KW apx. Input: 115. 400 cy. Modulator pulse duration .5 to 2 micro-sec. apx. 13 KV Pk Pulse. Compl. with all tubes incl. 715-B, 829B, RKR 73, two 72's. Compl. pkg., new $210.00
APS-15B. Complete pkg. as above, less modulator $150.00
"S" BAND AN/APS-2. Complete RF head and modulator, including magnetron and magnet, 417-A mixer, TR, receiver, duplexer, blower, etc., and complete pulser. With tubes, used, fair condition $75.00
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Modulator-motor-alternator unit for above ....$ 75.00
Receiver-rectifier power unit for above ....$ 25.00
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STEP DOWN TRANSFORMER: Pri: 440/220/110 volts a.c. 60 cycles, 3 KVA. Sec. 115 v. 2500 volt ins. 12" x 12" x 7" $40.00

PLATE Transformer: Pri: 117 v. 60 cy Sec. 17,000 v @ 144 ma. w/ choke. Oil filled. Size 26" x 20" x 13" AMERTRAN ...$120.00

"Communications"

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MAGNETRONS

TUBE	FREQ. RANGE	PK. PWR. OUT.	PRICE
2J31	2820-2860 mc.	265 KW.	$25.00
2J21-A	9345-9405 mc.	50 KW.	$25.00
2J22	3267-3333 mc.	265 KW.	$25.00
2J26	2992-3019 mc.	275 KW.	$25.00
2J27	2965-2992 mc.	275 KW.	$25.00
2J32	2780-2820 mc.	285 KW.	$25.00
2J37			$45.00
2J38 Pkg.	3249-3263 mc.	5 KW.	$35.00
2J39 Pkg.	3267-3333 mc.	87 KW.	$35.00
2J40	9305-9325 mc.	10 KW.	$65.00
2J49	9000-9160 mc.	58 KW.	$85.00
2J55 Pkg.	9345-9405 mc.	50 KW.	$35.00
2J61	3000-3100 mc.	35 KW.	$65.00
2J62	2914-3010 mc.	35 KW.	$65.00
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5J30			$39.50
714AY			$25.00
718DY			$25.00
720BY	2800 mc.	1000 KW.	$50.00
720CY			$50.00
725-A	9345-9405 mc.	50 KW.	$25.00
730-A	9345-9405 mc.	50 KW.	$25.00

Klystrons: 723A/B $12.50; 707B W/Cavity $20.00
417A $25.00 2 K.41 $65.00

MAGNETRON MAGNETS

GAUSS	POLE DIAM.	SPACING	PRICE
4850	¾ in.	⅝ in.	$12.50
2500	1⅝ in.	1 11/16 in.	$12.50
1500	1⅝ in.	1½ in.	$12.50
D161392*	1⅝ in.	1 5/16 in.	$12.50

* Mfr's Number.

TUNABLE PKGD. 'CW" MAGNETRONS

QK 61 2975-3200 mc. QK 62 3150-3375 mc.
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New, Guaranteed ....Each $65.00

3 CENTIMETER PLUMBING

(STD. 1" x ½" GUIDE, UNLESS SPECIFIED)

"X" BAND PREAMPLIFIER, consisting of 2-723 A B local oscillator-beacon feeding waveguide and TR/ATR Duplexer sect. incl. 60 mc IF amp $47.50
Random Lengths wavegd. 6" to 18" Lg. ...$1.10 Ft.
WAVEGUIDE RUN 1⅛" x 1/₂" guide, consisting of 4 ft. section with Rt. angle bend on one end 2" 45 deg. bend other end $8.00
WAVEGUIDE SECTION 1⅛" x ½" choke to choke 4 ft. long. $10.00
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15 DEG BEND 10" choke to cover $4.50
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Rotary Joint, choke to choke with deck mounting $17.50
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CIRCULAR CHOKE FLANGES solid brass ....$.55
SQ. FLANGES FLAT BRASS ....ea. $.55
APS-10 TR/ATR DUPLEXER section with additional IRIS Flange $10.00
CU 105/APS 31 Directional coupler 25 lb. $15.00
CU 103/APS 33 Directional coupler 25 lb. $15.00
FLEXIBLE WAVEGUIDE ....$4.00/FT.
"X" BAND calibrated attenuator $85.00
SHIELDED KLYSTRON Tube Mounts with rough attenuator outputs $90.00
FLEXIBLE SECTION 2½" cover to cover $5.00
TRANSITION WAVEGUIDE ⅜" x ⅞" x 1" x ½". 14" L. ....ea. $8.00

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D-170396 (bead) ...$.95
D-167613 (button) .$.95
D-166228 (button) .$.95
D-164699 for MTG. in "X" band Guide $2.50
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UG 86 U ....$.95
UG 254/U ....$.75

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D-167176 ....$.95
D-168687 ....$.95
D-171812 ....$.95
D-171528 ....$.95
D-168549 ....$.95
D-162482 ....$3.00
D-166277 ....$2.50
D-162356 ....$1.50
D-16187A $2.85
D-163075 ....$1.25
3A(12-43) $1.50
D167020 ....$3.00

PULSE EQUIPMENT

MODULATOR UNIT BC 1203-B

Provides 200-4,000 PPS. Sweeptime: 100 to 2,500 microsec. in 4 steps, fixed mod. pulse, suppression pulse, sliding modulating pulse, blanking voltage, marker pulse sweep voltages, calibration voltages, fil. voltages. Operates 115 vac. 50-60 cy. Provides various types of voltage pulse outputs for the modulation of a signal generator such as General Radio #804B or #804C used in depot bench testing of SCR 695, SCR 595, and SCR 535. New as shown $125.00
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TPS-3 PULSE MODULATOR. Pk. power 50 amp. 24 KV (1200 KW pk); pulse rate 200 PPS, 1.5 microsec; pulse line impedance 50 ohms. Circuit—series charging version of DC Resonance type. Uses two 705-A's as rectifiers. 115 v. 400 cycle input. New with all tubes $49.50
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G.E. #25E5-1-350-50P2T, 25 KV, 5 sections, "E" circuit, 1 microsecond pulse length, 350 PPS, 50 ohms impedance $45.00
G.E. #6E3-5-2000-50P2T, 6KV, "E" circuit, 3 sections, .5 microsecond, 2000 PPS, 50 ohms impedance .$6.50
G.E. #3E (3-.84-810; 8-2.24-405) 50P4T: 3KV, "E" CKT Dual Unit: Unit 1, 3 Sections, .84 Microsec, 810 PPS, 50 ohms imp.; Unit 2, 8 Sections, 2.24 Microsec, 405 PPS, 50 ohms imp. $6.50
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7.5E3-3-200-6PT. 7.5 KV, "E" Circuit, 3 microsec, 200 PPS, 67 ohms imp., 3 sections $12.50

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D-168184: .5 microsec. up to 2000 PPS, 1800 ohm term $4.00
D-170499: .25/.50/.75/. microsec. 8 KV. 50 ohms imp. $18.50
D-165997: 1¼ microsec. $7.50

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G.E.K.-2745 ....$39.50
G.E.K.-2744-A $39.50
W.E. #D166173 Hi-Volt input transformer, W.E. Impedance ratio 50 ohms to 900 ohms. Freq. range: 10 kc to 2 mc. 2 sections parallel connected, potted in oil $36.00
W.E. KS 9800 Input transformer. Winding ratio between terminals 3-5 and 1-2 is 1.1:1, and between terminals 6-7 and 1-2 is 2:1. Frequency range: 380-520 c.p.s. Permalloy core $6.00
G.E. #K2731 Repetition Rate: 635 PPS, Pri. Imp: 50 Ohms. Sec. Imp: 450 Ohms. Pulse Width: 1 Microsec. Pri. Input: 9.5 KV PK. Sec. Output: 28 KV PK. Peak Output: 800 KW. Bifilar 2.75 Amp. $64.50
W.E. #D169271 Hi Volt input pulse Transformer $27.50
G.E. K2450A Will receive 13KV. 4 micro-second pulse on pri., secondary delivers 14KV. Peak power out 100KW GE $45.00
G.E. #K2748A Pulse Input, line to magnetron ..$36.00
#9280 Utah Pulse or Blocking Oscillator XFMR Freq. limits 790-810 cy-3 windings turns ratio 1:1:1 Dimensions 1 13/16 x 1 1/8 x 19/32" $1.50

400 CYCLE TRANSFORMERS

352-7273: Pri: 115V. 400 cy. Sec: 6.3V, 2.5 Amp; 6.3V, .06 Amp; 6.3V, .9 Amp; 5V, 6 Amp; 700 VCT, 2-5U4's. For APS-15. T201 ....$4.75
352-7176: Pri: 115V. 400 cy. Sec: 6.3V, 20 Amp; 6.3V, .5 Amp; 6.3V, .5 Amp; 320V (2-6x5's). For APS-15. T202 ....$5.25
352-7278: Pri: 115V. 400 cy. Sec: 2.5V, 1.75 Amp. 3500V (2x2). For APS-15. T203 (Anode #2 5FP7) $5.85
352-7070: Pri: 118V. 440 cy. Sec. 2.5V, 2.5 Amp; 2.5V, 2.5 Amp: (2000V. Ins.); 6.3V, 2.25 Amp; 1200V. Tpd. at 1000 and 750V. P/o AN/APS 15 $4.95
#7469105: Pri: 115V. 400 cy. Sec: Tpd. to give 742.5V. 50 MA; 709V, .0477 A. 671V, .045 A. $2.95
M-7474319: Pri: 115V. 400 cy. Sec: 6.3V, 2.7 Amp; 6.3V, 66 Amp; 6.3V, .21 Amp. $2.95
32332: Pri: 115V. 400-2400 cy. Sec: 400 Vct. 35 MA; 6.4V, 2.5 Amp; 6.4V, .15 Amp. $2.25
332-7138M: Pri: 115V. 400-2400 cy. Sec: 640V, .5 MA; 2.5V, 1.75 Amp. $3.85
352-7179: Pri: 115V. 400-2400 cy. Sec: 6.5V, 12 Amp. Ct. 250V, 100 MA; 5V, 2 Amp. $3.50
#9069: Pri: 115/80V. 400-2600 cy. Sec: 650 Vct. 50 MA; 6.3 Vct, 2 Amp; 5 Vct, 2 Amp. $2.15
352-7096: Pri: 115/80V. 400-2400 cy. Sec: 2.5V, 1.75 Amp, 3 Kv. Ins.; 5V, 3 Amp; 6.5V, 6.5 Amp; 6.5V, 1.2 Amp. $3.95
KS 9607: Pri: 115V. 400-2400 cy. Sec: 734 Vct. 177 MA, 1710 Vct. 177 MA. $5.95
D-166333: Pri: 115V. 400-2400 cy. Sec: 6.3V, 0.9 Amp, 7.7V, 0.365 Amp. $2.79
GE #7471957: Pri: 100/110/120/130V. 400-2400 cy. Sec: 2.5V, 20 Amp. 11V ins. $4.85
D-163254: Pri: 115V. 400 cy. Sec: 6.3V, 12 Amp; 6.3V, 2A; 6.3V, 1A. P/O AN/APQ5 ....$5.85
KS-9685: Pri: 115V. 400-2400 cy. Sec: 6.4Vct, 7.5 Amp; 6.4V, 3.8 Amp; 6.4V, 2.5 Amp. $4.35
using 2J27 magnetron oscillator, 250 KW peak input. 707-B receiver-mixer $150.00
Modulator-motor-alternator unit for above ....$ 75.00
Receiver-rectifier power unit for above ....$ 25.00
PLATE XFMR: Pri: 115V. 400 cy. Sec: 9800V. or 8600V. @ 32 MA. DC. $12.50

OIL CONDENSERS

1.5 mfd. 6000vdc. $12.50
.25 mfd. 20,000 vdc. $17.50
10 mfd. 1000 vdc. .$1.79
15 mfd. 1000 vdc. .$2.25
1 mfd. 6000 vdc. ..$6.50
3 x 10 mfd. sync. cond. 90v. 60 cy. $4.95

All merchandise guaranteed. Mail orders promptly filled. All prices, F.O.B. New York City. Send Money Order or Check. Only shipping charges sent C.O.D. Rated Concerns send P.O.

COMMUNICATIONS EQUIPMENT CO.

131-I Liberty St., New York, N.Y. Phone: Digby 9-4124 ATT Mr. Chas. Rosen

MEMBERSHIP

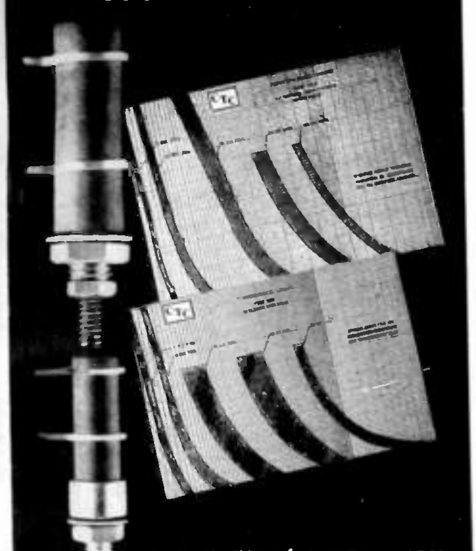
(Continued from page 46A)

Wilkinson, R. L., 408 York Rd., Baltimore, Md.
Williams, R. W., 3097 Walnut St., Huntington Park, Calif.
Yondorf, E. G., 615 Cator Ave., Baltimore 18, Md.
Zion, H., 162-23 73 St., Flushing, L. I., N. Y.

The following transfers to Associate grade were approved to be effective as of January 1, 1949:

Anthony, J. P., 3546 Ozark, Houston 4, Tex.
Arnold J. B., 209-27 Hillside Ave., Apt. S, Queens Village, L. I., N. Y.
Barnes, P. M., Sandia Base Branch, Albuquerque, N. Mex.
Battaglino, V., Puan 383, Buenos Aires, Argentina
Belfi, J. L., 83 MacDougal St., New York 12, N. Y.
Boulet, L., Department of Engineering, Laval University, Quebec, Que., Canada
Burke, M. H., R.F.D. 1, Box 25, Farmingdale, N. J.
Carlson, C. G., 11206 91 St., Edmonton, Alta., Canada
Darr, J. E., 402 Ross Ave., Wilkinsburg 21, Pa.
Dobler, J. W. H., Jacob Ford Village, Bldg. 10, Apt. 1A, Morristown, N. J.
Doel, D., 358 Kellogg Park, Portland 2, Ore.
Falk, A. K., 1330 Carleton St., Port Huron, Mich.
Freyman, R. W., 418 Kiva St., Los Alamos, N. Mex.
Giordano, A. F., 112 Third Ave., Newark 4, N. J.
Gough, L. E., Box 548, State College, N. Mex.
Groves, Q. D., 26 Wellington Rd., Merrick, L. I., N. Y.
Hiatt, W. C., R.F.D. 2, Hope, Ind.
Howie, J. S., 38 Nelson St., Gordon, N. S. W., Australia
Kaplan, H., 5553 W. Congress St., Chicago 44, Ill.
Krainin, S., 222-07 141 Ave., Laurelton 13, N. Y.
Lashier, H. M., Emmanuel Missionary College, Berrien Springs, Mich.
Lepanto, P. J., 35-28 95 St. Jackson Heights, L. I., N. Y.
Lewis, M. B., 3051 Ocean Ave., Brooklyn 29, N. Y.
Malone, W. C., Jr., 8071 Mountain Blvd., Oakland 5, Calif.
Mattison, G. H., 3215 Corlear Ave., New York 63, N. Y.
Montgomery, R. A., Rm. 239, Bldg. 37, General Electric Co., Schenectady, N. Y.
Moore, R. E., 17 Glen Ave., Scotia, N. Y.
Nevitt, R. G., 613 So. Western Pkwy., Louisville 11, Ky.
Nichols, F. K., Department of Electricity, U. S. Military Academy, West Point, N. Y.
Paul, W., 1458 University Ter., Ann Arbor, Mich.
Peckhart, M. A., R.F.D. 1, Fort Wayne, Ind.
Pimenoff, V. J., Department of Physics, University of New Brunswick, Fredericton, N. B., Canada
Powell, M., Jr., 896 Idlewild Dr., Madison, Tenn.
Preziosi, F. W., 7466 St. Denis St., Montreal, Que., Canada
Reed, D. E., Box 90, R.F.D. 2, San Luis Obispo, Calif.
Rubin, W. L., 3549-12 Ave., Brooklyn 18, N. Y.
Salsbury, D., 44 Harvest Lane, Hicksville, L. I., N. Y.
Schupp, G. A., Jr., 40 Cedar St., Worcester, Mass.
Seidler, R. L., 179 N. Main St., Spring Valley, N. Y.
Simon, A., 354 Bergen Ave., Jersey City, N. J.
Sperrazzo, V. J., 161 Troutman St., Brooklyn, N. Y.
Spieker, L. J., Box 347, Brainerd, Minn.
Summer, C. F., Jr., Box 401, R.F.D. 3, Turtle Creek, Pa.
Turner, W., 265 Ridgeway Ave., Rochester, N. Y.
Vacca, L. N., 43 Midland Pl., Newark 6, N. J.
Wilson, G. G., 2858 E. 194 St., New York 61, N. Y.
Wulfsberg, K. N., 5 Hooker St., Allston, Mass.

THESE EXTRA SMALL COIL FORMS FIT INTO TIGHT PLACES



Graphs show frequency ranges covered by the five standard sizes of LSM and LS3 Coils. Write for full-size copies.

If small space is your problem — as in peaking coils in strip amplifiers, chokes, R. F. coils, oscillator coils, single-turned I. F. coils, etc.—you'll find space-saving one of many advantages in CTC Slug Tuned Coil Forms.

Coil forms are of quality paper-base phenolic, high frequency grade. Mounting bushings and ring-type terminals are brass, the bushings cadmium plated and terminals silver plated. Necessary mounting hardware is supplied.

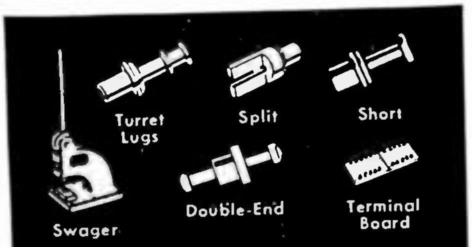
DIMENSIONS

LSM — Extreme small size; only 27/32" high when mounted; coil form, ¼" diameter; mounts in single #18 hole; mounting bushing has 8-32 thread.

LS3 — Moderate small size; 1⅛" high when mounted; coil form, ⅜" diameter; mounts in single ¼ hole; mounting has ¼-28 thread.

WINDINGS

CTC LSM and LS3 Coil Forms are available unwound or in any of five standard windings: 1, 5, 10, 30 and 60 megacycles. They are also wound to specifications. (Standard slug is high-frequency type.) CTC will custom-engineer special coils of practically any size and winding . . . Let us talk over your requirements.



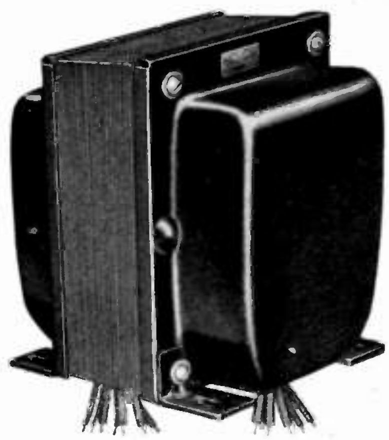
Custom or Standard
The Guaranteed
Components

CAMBRIDGE THERMIONIC CORPORATION
456 Concord Avenue, Cambridge 38, Mass.

EAR and EYE TUNED

TELEVISION TRANSFORMERS

This Acme Electric 500 V.A. Power Supply Transformer for television receivers, has been carefully engineered to provide the exact electrical characteristics required for larger sets. Hum-free operation has been attained thru both riveting and bolting core and varnish impregnating entire unit.



This larger V.A. capacity transformer, permits manufacturers to use only one transformer instead of two.

From standard laminations, sizes and standard mounting cases Acme Electric engineers can design exactly the transformers you need to improve your product. We invite your inquiry.

ACME ELECTRIC CORP.
444 Water St. Cuba, N.Y., U.S.A.

Acme Electric
TRANSFORMERS

MICRODIMENSIONAL WIRE & RIBBON

Made in almost all ductile metals and alloys; or we will draw wire from your own metals.

Your inquiry, with engineering specifications is invited.

SIGMUND COHN CORP.
44 GOLD ST. NEW YORK
SINCE 1901

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . .

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.

I East 79th St., New York 21, N.Y.

SENIOR RADIO ENGINEERS

Must have good fundamental training, experience in circuit design and test of government-type portable and mobile transmitters and receivers. Write fully to Government Contract Dept., Pilot Radio Corporation, 3706-36th Street, Long Island, City, N.Y.

ELECTRONIC ENGINEERS

College graduates with 3-5 years of development engineering experience in circuit design. Well versed in magnetic circuits, non-linear circuit operation and electronic theory. Send résumé and all particulars to Personnel Department, General Precision Laboratory, Inc., Pleasantville, New York (30 miles north of New York City).

ENGINEERS AND PHYSICISTS

Unusual opportunity for engineers or physicists familiar with SCR-584 radar, electronic computers, or synchronization and timing. Desirable beach location. Many advantages. For application forms write Rt. 1, Box 118, Oxnard, California.

VACUUM TUBE ENGINEER

Vacuum tube engineer (starting salary $6235.20) to head well equipped and well staffed vacuum tube development shop in Government laboratory. Requires engineer or physicist with several years broad experience in vacuum tube development and construction and a genuine interest in developing new tube making techniques. Box 559.

ELECTRONIC ENGINEER

To work on challenging control projects of an electrical, electro-mechanical and electronic nature. Good pay with excellent future with an established New England company of top rating for young man with necessary background who has initiative and imagination. Apply Box 560.

DESIGN ENGINEERS

We have several immediate openings for design or development engineers. Mechanical and electrical engineers with considerable experience in design of instruments or control preferred. Excellent opportunities with leading manufacturers of automatic controls. Attractive salary. Location: Minneapolis. Contact J. A. Johnson, Employment office, Minneapolis-Honeywell Regulator Company, 4th Avenue and 28th Street, Minneapolis 8, Minnesota.

RADIO PROJECT ENGINEER

Graduate engineer with at least 5 years recent experience in design and development of low power oscillator and amplifier circuits as used in signal generators in the very high frequency range. Must be

(Continued on page 51A)

POSITIONS OPEN

(Continued from page 50A)

familiar with theoretical concepts and calculations of the circuit components, as well as practical design and layout work. Applicant should have initiative and supervisory ability and be capable of assuming full responsibility for project. Apply by letter only. Address Personnel Dept. Federal Mfg. & Engineering Corp., Brooklyn 5, N.Y.

ELECTRONIC AND SALES ENGINEERS

Recently established and rapidly expanding company has openings for Junior Electronic Engineers in development and field engineering. Also openings for sales engineers in south, southeast, Pennsylvania and New England. Attractive remuneration. Company specializes in industrial electronics, instrumentation and control. Write giving full details to Fielden Electronics, Inc., Huntington Station, N.Y.

ELECTRONICS RESEARCH ENGINEER

Electronics research engineer experienced in electronics as well as engineering sciences and physics for research and development work in connection with motor and generator control equipment. Excellent opportunity with large manufacturer in New York City. Please write, giving complete education and experience. Replies will be held in confidence. Our employees know of this ad. Box 561.

ENGINEERS

Location, Phoenix, Arizona. Excellent working conditions. Housing available. Motorola Inc. announces a Research Laboratory devoted to armed service contract and company research in microwave, mobile communications, supervisory control, telemetering, miniaturization, and aviation electronics. Only fully qualified experienced inventors, engineers and scientists should apply. Send detailed statement of education and experience to Motorola Inc., D. E. Noble, 4545 Augusta Blvd. Chicago 51, Ill.

SENIOR AND JUNIOR ENGINEERS

Senior and junior engineers needed with experience on SCR-584 radar or similar equipment. Location about 50 miles from Los Angeles. Electronic Engineering Co., 2008 W. 7th Street, Los Angeles 5, Calif.

ELECTRONICS ENGINEER

Radio and industrial electronics instructor for two years technical college-Extension Division of Georgia Institute of Technology, Atlanta, Georgia. Write The Technical Institute, Chamblee, Georgia.

ELECTRONICS ENGINEERS, ELECTRO-MECHANICAL DESIGNERS, MATHEMATICIAN, PHYSICIST, LAYOUT DRAFTSMEN, AERODYNAMIST (AERONAUTICAL ENGINEER)

Men needed immediately for permanent positions on an experimental, development, and production program of complex electronic and electro-mechanical equipment. Work covers computers, servos, amplifiers, instrumentation, small mechanisms in aircraft simulation for complex training equipment. Applicants must have college degree, equivalent experience or

(Continued on page 52A)

Electronic Engineers

BENDIX RADIO DIVISION
Baltimore, Maryland
manufacturer of
RADIO AND RADAR EQUIPMENT
requires:

PROJECT ENGINEERS

Five or more years experience in the design and development, for production, of major components in radio and radar equipment.

ASSISTANT PROJECT ENGINEERS

Two or more years experience in the development, for production, of components in radio and radar equipment. Capable of designing components under supervision of project engineer.

Well equipped laboratories in modern radio plant . . . Excellent opportunity . . . advancement on individual merit.

Baltimore Has Adequate Housing

Arrangements will be made to contact personally all applicants who submit satisfactory resumes. Send resume to Mr. John Siena:

BENDIX RADIO DIVISION
BENDIX AVIATION CORPORATION
Baltimore 4, Maryland

Electron Tube Engineers

Positions open for

1. **EXPERIENCED MICRO-WAVE DEVELOPMENT ENGINEERS**
2. **JUNIOR MICROWAVE ENGINEERS**

   Experience not needed but college degree and scholastic ranking in upper ¼ of class necessary.

3. **TUBE PROCESS ENGINEERS**

These positions are open at a New England Manufacturing Plant specializing in microwave electron tube development and manufacture.

Salary commensurate with experience and ability—insurance plan—bonus plan—paid vacations. Please furnish complete résumé of education, experience, and salary required. Box 562.

The Institute of Radio Engineers
1 East 79th St. New York 21, N.Y.

Radio and Radar Development and Design Engineers

Openings for experienced men at

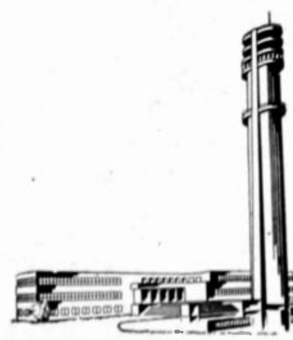
HAZELTINE ELECTRONICS CORPORATION
Little Neck, L.I., N.Y.

Please furnish complete resume of experience with salary expected to:
Director of Engineering Personnel

(All inquiries treated confidentially)

Wanted . .
ELECTRICAL ENGINEERS

Our modern, progressive organization has a number of desirable openings for qualified graduate engineers with specialized experience in the Communication and Aerial Navigation Fields. Program includes development of FM, AM, TV and Micro-Wave Transmitters and Receivers, Pulse Code Modulation Systems, Radar, Vacuum Tubes and Direction Finders. MS and PhD Comm. majors considered.



Address Inquiries to Personnel Manager—

Federal Telecommunication Laboratories, Inc.
500 Washington Ave. Nutley, N.J.

POSITIONS OPEN

(Continued from page 51A)

both. Apply Personnel Mgr. Link Aviation, Binghampton, N.Y. Specify particular opening, qualifications, salary expected. Interviews will be arranged.

ELECTRO-ACOUSTIC ENGINEER

An attractive opportunity for an experienced engineer with thorough electro-acoustic training (transducers) in a small, sound rapidly growing concern. Permanent position with future opportunity as research director. Write giving full details to Box 563.

Positions Wanted By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants, and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The Institute publishes free of charge notices of positions wanted by I.R.E. members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The Institute necessarily reserves the right to decline any announcement without assignment of reason.

TELEVISION ENGINEER

B.S.T.E. Age. 25. Married. First class radiotelephone license. Desires position in television station or development work. Trained in operation and maintenance of R.C.A. Image Orthicon, DuMont equipment and very high frequency techniques. Box 205 W.

JUNIOR ENGINEER

Graduate 2 year course television engineering. Married. Age 25. First class FCC license. Trained in all phases of television studio work. Desires position in television broadcasting field. Box 206 W.

JUNIOR ENGINEER

Syracuse University. B.E.E. June 1948. Age 26. Married with no children. Desires work with power company or motor manufacturing company located in the east. Prefer training program if possible. Interested in transmission and mathematical design. Box 207 W.

ENGINEER

University of Minnesota, communications major. B.E.E. with distinction, August 1948. 2½ years electronics experience in U.S. Army. Desires position in production or electronic development. Will work anywhere in U.S. Box 208 W.

(Continued on page 53A)

Positions Wanted

(Continued from page 52A)

ELECTRONICS ENGINEER

Will graduate March 1949 Iowa State College B.S.E.E. in communications. Married. Age 25. First class Radio Telephone license. Some servicing experience. Desires position in radio or electronics anywhere in U.S. Box 209 W.

ENGINEER

B.S.E.E. Northeastern University, Boston 1947. Two years experience as Navy radio technician with Navy radar and communication equipments. Valuable experience in UHF antenna and radiation research and development at Naval Air Test Center, Patuxent River, Maryland. Desires position in research and design of antennae. Box 210 W.

JUNIOR ENGINEER

B.S. Television Engineering, American Television Institute of Technology, January 1949. Age 25. Married, no children. Three years experience on Navy radar. Desires position in microwave research. Anywhere in U.S. Box 211 W.

JUNIOR ENGINEER

R.C.A. Institute's graduate seeks position in research and development, or production field in New York area. Age 29. Married. Studying for engineering degree at night. Air Force officer, 2½ years experience as instructor in bomb-sight and auto-pilot theory and operation. Box 212 W.

TEACHER

Candidate for M.A. in physics August 1949. 10 years teaching. High school and Army radio school. Also Army and broadcast radio experience. Age 38 with family. Prefer college teaching in southwest. Box 220 W.

TELEVISION ENGINEER

B.E.E., Age 25, Married. 3 years radar design and maintenance, including 1½ year Navy ETM instructor; 2½ years television development and production. Available June. Prefer California. Box 221 W.

ELECTRONIC ENGINEER

B.E.E. 1943; 3 semesters postgraduate work. 5 years experience in electronic instrumentation in connection with rockets and guided missiles. Also remote guidance and control mechanisms, servos and digital computers. Interested in position involving design and development along similar lines. Box 222 W.

ELECTRONIC PHYSICIST

B.S. and graduate work in physics. 1 year electronic laboratory experience, 1 year full time and 1 year part time additional professional experience plus 4 years Navy radar maintenance. Prefer position in west or mid-west. Box 223 W.

JUNIOR ENGINEER

Graduating in radio engineering, March 1949. B.S. Age 24, married, no children. 3 years naval experience as radio technician, with accent on Airborne radar counter-measures. Desires position in electronic circuit design and development. Will consider foreign position. Box 224 W.

ENGINEER

B.S.E.E. University of Wisconsin. 2 years Army communications experience and 1 year experience in electronic and control circuit design. Age 25. Single. Desires position in production or electronic and control circuit development. Box 225 W.

(Continued on page 54A)

Positions Wanted

(Continued from page 53A)

RADIO PHYSICIST

B.Sc. May 1949, radio physics option of honor mathematics and physics, University of Western Ontario. 3 years radar in Canadian Navy. Age 27, married. Desires development, research or interesting position with good future. Box 243 W.

ELECTRONIC ENGINEER

B.S.E.E. June 1949 University of Cincinnati, Age 30. Experience 2½ years co-op in development and test components, video and I.F. amplifiers, servo-mechanisms, instrument landing systems. Working knowledge of machine tools and machine shop practice. Desires position in research or development. Member of Eta Kappa Nu. Box 244 W.

ENGINEER

B.A. Sc., E.E., April 1949, U.B.C.; age 27, married; appreciable experience as technician in F.M. and microwave radio relay, audio, mobile and harbour control F.M. with some design and development. Desires development or research engineering. Preference microwave. D. Nuttall, 2870 W. 13, Vancouver, B.C. Canada.

ENGINEER

Interested in television and its industrial application. Columbia University 1949, M.S. in I.E. City College 1948 B.S. in E.E Engineer in training State of New York Prefers position in New York area. Box 245 W.

ELECTRONIC ENGINEER

B.E.E. February 1949 Cornell University. Age 24. Air Corps electronics officer, training at M.I.T., Harvard, Yale. Experience in electronics and power work, Interested in industrial electronics field. Box 246 W.

JUNIOR ENGINEER

Graduate R.C.A. Institutes; age 30; married. 3 years electrical engineering major in electronics, continuing studies for B.E.E. at night. Desires work as Junior Engineer in design and development of communication equipment under Senior Engineer. 4 years radio servicing experience preceding 3½ years U.S. Army radio servicing. 1st. class Radiotelephone, 2nd class Radiotelegraph, and Class B amateur F.C.C. Licenses. Box 247 W.

ELECTRONIC ENGINEER

B.S. in E.E 1948 University of Illinois. Experience with reputable firm in circuit design. Desires position anywhere in the U.S. Box 248 W.

PHYSICIST

Graduate of Rensselaer Polytechnic Institute. 28 years of age. 3½ years radar and communications army experience 2 years research and development on communications systems, multiplex systems, servo systems. Experienced in all phases of executing government contracts. Desires position in New York City. Box 249 W.

ENGINEERING—ADVERTISING

Engineering-advertising is a rare combination. I have extensive copy experience in technical advertising literature in large company; strong technical manual writing; knowledge advertising production; BSEE; engineering experience; high references. Young, personable, like people. Desires change, opportunity to grow with firm and help firm grow. Box 250 W.

(Continued on page 55A)

Positions Wanted

(Continued from page 54A)

SALES OR FIELD ENGINEER

Young, hard hitting research and design engineer seeks inquiries from firms needing addition to sales or field engineering staff. Desires change to position which will allow combination of engineering and top sales talents. Have solid experience in all phases of electronics and communications. Also quality control and teaching experience along with B.E.E. degree. Married, 1 child. Will consider Baltimore, Md. or vicinity. Box 226 W.

SALES ENGINEER

Strong technical background, good sales personality. Desires position in New York City vicinity. Box 227 W.

News—New Products

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

(Continued from page 28A)

of the three voltage pedestals produced by the photocell is amplified in a common amplifier circuit and then applied to the vertical deflection plates of an oscilloscope. The sweep voltage is applied to the horizontal plates of the oscilloscope in such a manner that termination of each viewing cycle occurs at the same time that the scanning disk alternates from one aperture to another.

The indications of the three voltage pedestals appear as three horizontal lines on the face of the oscilloscope due to the persistence of the screen, combined with a rapid scanning rate.

New Wide Range Logarithmic Attenuator

The Kay-Lab Logaten, a wide-range logarithmic attenuator with an output proportional to the logarithm of its input for a range of 50 db, has been developed by **Kalbfell Laboratories, Inc.**, Dept. P, 1076 Morena Blvd., San Diego 10, Calif.

Frequency response of the attenuator is flat from dc to 500 kc, and only minor errors are found up to several megacycles. By plugging a Logaten into a vacuum-tube voltmeter, oscilloscope, or recorder, these devices can be made to read linearly in decibels. In its octal-plug form, the Logaten is suitable for inclusion in amplifiers. It is adaptable to many applications involving electronic computers through the well-known principle of multiplication and division by logarithms. It also finds application wherever data having a very great dynamic range must be recorded on a single sensitivity range of an instrument. In acoustical reverberation studies, it converts the ordinary exponential decay curve into a straight line.

The Logaten is a network of nonlinear circuit elements adjusted to give an output voltage which is accurately proportional to the logarithm of input voltage. Both input and output impedances are 10,000 ohms. Its case is $1\frac{1}{2}$" in diameter by $2\frac{3}{8}$" long.

(Continued on page 56A)

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 55A)

New Catalogs

• • • A new loose-leaf catalog illustrating relays, transformers, coils, terminals, etc., has just been released and is available without cost to those who inquire on their firm stationery, from **Comar Electric Co.,** 2701 Belmont Ave., Chicago 18, Ill.

• • • In September, 1948, this column published an item describing the Twin-Trax Recorder manufactured by **Amplifier Corp. of America,** 398-1 Broadway, New York 13, N. Y. Now this firm is offering a 12-page booklet compiled from an analysis of 5,000 letters answering the most-often-asked questions about the recorder. Entitled "99 Questions," this folder will be sent free on request.

• • • A folder on mechanical development apparatus for laboratory and developmen engineers has been released by a new firm, **Servomechanisms, Inc.,** Old Country & Glen Cove Rds., Mineola, L. I., N. Y.

• • • A six page folder in color describing various mounts for antenna installation by **Metalace Corp.,** 2101 Grand Concourse Bronx, 35. New York.

(Continued on page 57A)

News—New Products

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

(Continued from page 56A)

New Microwave Power Meter

A new microwave power meter, known as the -hp-430A, designed to automatically indicate power developed in a standard barreter, is now being marketed by **Hewlett Packard Co.,** 395 Page Mill Rd., Palo Alto, Calif.

Power level is read directly on a 4" square meter face. No calculation or knob-twisting is necessary once range selection and zero set are made. The -hp- 430A is automatically self-balancing, and may be used over any frequency, depending on the associated barreter and mount.

This meter consists of an ac bridge, one arm of which is a barreter. The bridge is in balance with zero rf power in the barreter. As rf power is applied to the barreter, an equivalent ac (audio) power is automatically removed. Thus the bridge remains in balance. A vacuum-tube voltmeter reads the change in audio power level. This meter, calibrated in milliwatts, gives a direct indication of the rf power in the barreter.

The indicating meter is calibrated in dbm in addition to the linear milliwatt calibration.

The five power ranges are selected on a convenient front-panel switch.

From a technical standpoint, the -hp 430A covers a power range of from 0.02 to 10 milliwatts. Ranges are related in 5-db steps, and continuous readings are available from −20 dbm to +10 dbm (0 dbm equals 0.001 watt). This power range may be extended by the use of attenuators or directional couplers. Accuracy of the meter is within ±5% of full-scale readings.

The -hp- 430A microwave power meter measures 12" wide by 9" high, and is 9" deep. It is powered by a 115-volt 60-cps ac source.

New Beta-Gamma Counter

An aluminum beta-gamma counter, RMA type designation 1B85, designed to replace the thin-walled glass tubes previously used in laboratory and field radiation-measuring instruments, was placed on the market a few months ago by **The Victoreen Instrument Co.,** 5608 Hough Ave., Cleveland 3, Ohio.

(Continued on page 58A)

News—New Products

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

(Continued from page 57A)

The thyrode 1B85 operates at 900 volts. The plateau length is not less than 200 volts, and the plateau slope does not exceed 3% per 100 volts. Nominal recovery time is 100 microseconds. The minimum operating life is 10^8 counts per minute with life-test end-point plateau 850 to 950 volts. The wall of 30 mg/cm² aluminum provides greater uniformity.

The new design features claimed by the manufacturer are: more nearly uniform and reproducable characteristics; plug-in base for practical water-tight probe or chassis mounting; and ribbed walls for greater mechanical strength.

(Continued on page 60A)

MAXIMUM PERFORMANCE

Jim Lansing Signature Speakers will provide an almost unbelievable realism. The experience gained through a quarter of a century of leadership in the sound reproduction field has gone into their development and design. For maximum dynamic range and frequency response compare Jim Lansing Signature Speakers before you buy.

MODEL D-130

Designed especially for music systems and public address use. Has exceptionally high efficiency. Recommended for operation and frequencies from 60 to 6500 C.P.S. with a maximum usable range of 40 to 15000 C.P.S.

MODEL D-1002
TWO WAY SYSTEM

Designed especially for FM Monitoring and high quality home sound reproduction. Housed in a beautiful console type cabinet.

Write for Descriptive Catalog containing complete specifications.

SEE YOUR JOBBER OR SEND DIRECT

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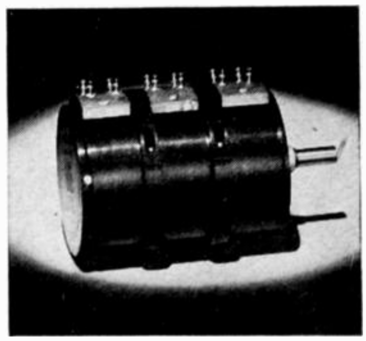
ACCURATE PHASING WITH T.I.C. METHOD OF GANGING PRECISION POTENTIOMETERS

MECHANICAL SPECIFICATIONS

- Precision machined aluminum base and cover 2" diameter, 1" depth.
- Precision phosphor bronze bushing.
- Centerless ground stainless steel shaft.
- No set screws.
- Mechanical rotation—360°.
- Clamping method of ganging permits individual adjustment of angular position.
- Temperature range —85° to +165°F.
- Rotational Life—At least 1 million complete cycles of revolution.

ELECTRICAL SPECIFICATIONS

- Winding—both linear to 0.2% and non-linear to 1% accuracies.
- Paliney contact to winding; two-brush rotor take-off assembly with precious metal contacts.
- High, uniform resolution provided by our method of winding non-linear resistances.
- Electrical rotation maximum 320°.
- All soldered connections (except sliding contacts).



(3 gang RV-2)

Write today for bulletins on other T.I.C. products: Z-Angle Meter . . . R.F. Z Angle Meter . . . R-F Power Oscillator . . . Translatory Variable Resistors . . . Slide Wire Resistance Boxes . . . Phase Angle Meter.

This general line of precision potentiometers was developed in collaboration with the Fire Control Section of the Glenn L. Martin Company.

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Only with CO-AX

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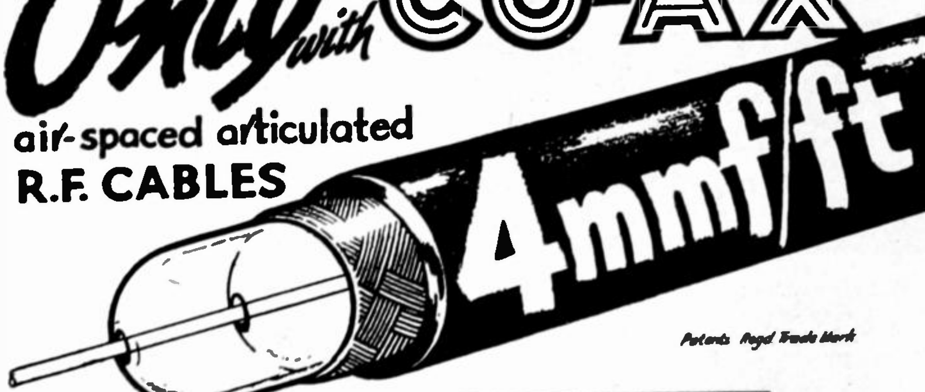

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We are specially organised to handle direct enquiries from overseas and can give

IMMEDIATE DELIVERIES FOR U.S.A.

Billed in Dollars. Settlement by your check. Transaction *as simple as any local buy.*

LOW ATTEN TYPES	IMPED OHMS	ATTEN db/100ft at 100 Mc/s	LOADING Kw	O.D."	
A 1	74	1.7	0.11	0.36	HIGH POWER FLEXIBLE
A 2	74	1.3	0.24	0.44	
A 34	73	0.6	1.5	0.88	

LOW CAPAC TYPES	CAPAC mmf/ft	IMPED OHMS	ATTEN db/100ft 100Mc/s	O.D."	
C 1	7.3	150	2.5	0.36	
PC 1	10.2	132	3.1	0.36	PHOTOCELL CABLE
C 11	6.3	173	3.2	0.36	
C 2	6.3	171	2.15	0.44	
C 22	5.5	184	2.8	0.44	
C 3	5.4	197	1.9	0.64	
C 33	4.8	220	2.4	0.64	V.L.C. ★
C 44	4.1	252	2.1	1.03	

★ *Very Low Capacitance cable*

TRANSRADIO LTD
CONTRACTORS TO H.M. GOVERNMENT
138A CROMWELL ROAD · LONDON SW.7 ENGLAND
CABLES: TRANSRAD. LONDON.

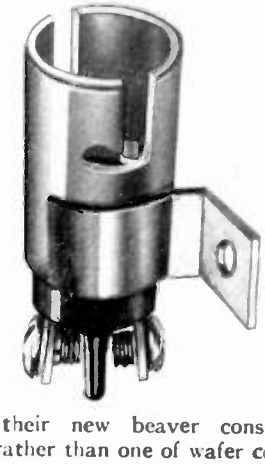
News—New Products

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

(Continued from page 58A)

Beaver Construction Socket for GE Pilot Lamp

For those companies using the GE 10-watt 115-volt double-contact bayonet-base pilot lamp, the **Cole-Hersee Co.**, 20 Old Colony Ave., Boston 27, Mass., recom-



mends their new beaver construction socket, rather than one of wafer construction, because a wafer will not make proper contact if the two solder spots on the base of the lamp are not exactly the same height.

(Continued on page 61A)

News—New Products

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(Continued from page 60A)

Suggested fabrications include the 2354 with straight-sided bayonet-base double-contact beaver construction, 2098 with flanged top, and 2384 with side bracket. Custom variations can be supplied if desired.

Terminals can be furnished with wire leads, instead of screws and lock washers.

Recent Catalogs

• • • A 12 page booklet entitled "Precision Investment Casting for Engineers" has been compiled to serve as an aid to design engineers, and is available from **Arwood Precision Casting Corp.**, 70 Washington St., Brooklyn, 1. N. Y.

• • • Two bulletins: No. 83, which describes rf inductors for broadcast phasing and tuning networks, and No. 85, which explains the type 1900 automatic dehydrator for use on coaxial transmission-line systems, may be obtained from **Andrew Corp.**, 363 E. 75th St., Chicago 19, Ill.

• • • Two bulletins, No. 485, describing the models LS100 and LS1000 decade-scaling laboratory counter sets, and No. 486, describing the model SM3 portable radiation survey meter, by **El-Tronics, Inc.**, 2647 N. Howard St., Philadelphia 33, Pa.

INDEX AND DISPLAY ADVERTISERS

DISPLAY ADVERTISERS

"Madame X" was the code name, during research and development, for an entirely new system of recorded music . . . perfected by RCA.

The remarkable background of **"Madame X"**

Now the identity of "Madame X," the *unknown* in a long search for tone perfection, has been revealed. From this quest emerges a completely integrated record-playing *system* —records and automatic player—the first to be entirely free of distortion to the trained musical ear . . .

The research began 11 years ago at RCA Laboratories. First, basic factors were determined—minimum diameters, at different speeds, of the groove spiral in the record—beyond which distortion would occur; size of stylus to be used; desired length of playing time. From these came the mathematical answer to the record's *speed*—45 turns a minute—and to the record's size, only 6⅞ inches in diameter.

With this speed and size, engineers could guarantee 5⅓ minutes of distortion-free performance, and the finest quality record in RCA Victor history!

The record itself is non-breakable vinyl plastic, wafer-thin. *Yet it plays as long as a conventional 12-inch record.* The new RCA Victor automatic record changer accommodates up to 10 of the new records—1 hour and 40 minutes of playing time—and can be attached to almost any radio, phonograph, or television combination.

Not only records are free of surface noise and distortion—the record player eliminates faulty operation, noise, and cumbersome size. Records are changed quickly, quietly . . . RCA Victor will continue to supply 78 rpm instruments and records.

This far-reaching advance is one of hundreds which have grown from RCA research. Such leadership adds *value beyond price* to any product or service of RCA and RCA Victor.

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To Our Customers:

We take pleasure in announcing the purchase of the Faradon Capacitor Division of the Radio Corporation of America.

Cornell-Dubilier acquired by the purchase the good will and trademark of "Faradon", the inventory, tools, dies, molds, equipment, instruments, designs, processes, and patent licenses. We have moved the Faradon equipment to our plants and are presently manufacturing the complete line of Faradon capacitors previously manufactured by the Radio Corporation of America.

Cornell-Dubilier transmitting capacitors and Faradon capacitors will be sold as separate lines, as Faradon capacitors are not always interchangeable with those of Cornell-Dubilier. Orders for Faradon capacitors, using the Faradon part numbers, may be mailed to our Sales Office at South Plainfield, New Jersey.

The high quality for which both Faradon and Cornell-Dubilier have been known for the last four decades will be meticulously maintained. The addition of the Faradon line will greatly improve our services, particularly to the broadcast stations and for those engaged in the specialty electronic fields.

The continued confidence of our customers in our product has made possible the acquisition of this additional outstanding line.

Sincerely yours,

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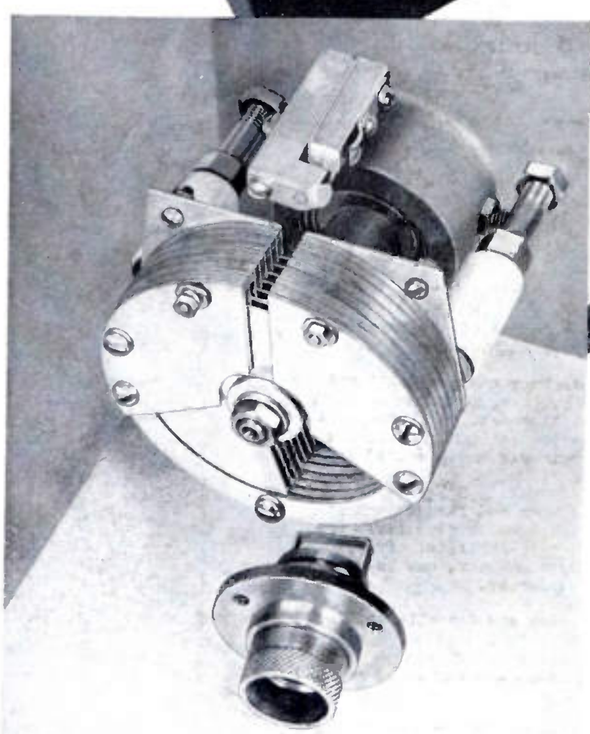
CAPACITORS • AUTO VIBRATORS • TV AND FM ANTENNAS • POWER CONVERTERS

100 to 500 Mc OSCILLATOR



The tuned circuit of the Type 857-A Oscillator is our well-known Butterfly type. The difficulty of sliding contacts in any part of the oscillator circuit is avoided in this unique construction. The photograph above shows the output coupling loop and output jack. Coupling can be changed from maximum to almost zero by rotating the output jack.

THIS oscillator, designed for use as a power source for general laboratory measurements and testing, covers the frequency range of 100 Mc to 500 Mc. With its associated power supply it is small, lightweight and compact. The entire range is covered with a single-dial frequency control with a slow-motion drive equipped with an auxiliary scale.

FEATURES

- Dial calibrated directly in megacycles to an accuracy of ± 1%
- Vernier dial with 100 divisions, covering the oscillator range in ten turns
- Output through a coaxial jack with provision for varying coupling
- Output of ½-watt at 500 Mc
- Electron-ray tube in power supply to indicate grid current and furnish indication of oscillation
- Filament and plate power furnished by the Type 857-P1 Power Supply which is furnished with the oscillator

TYPE 857-A U-H-F OSCILLATOR (with power supply) . . **$285**

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