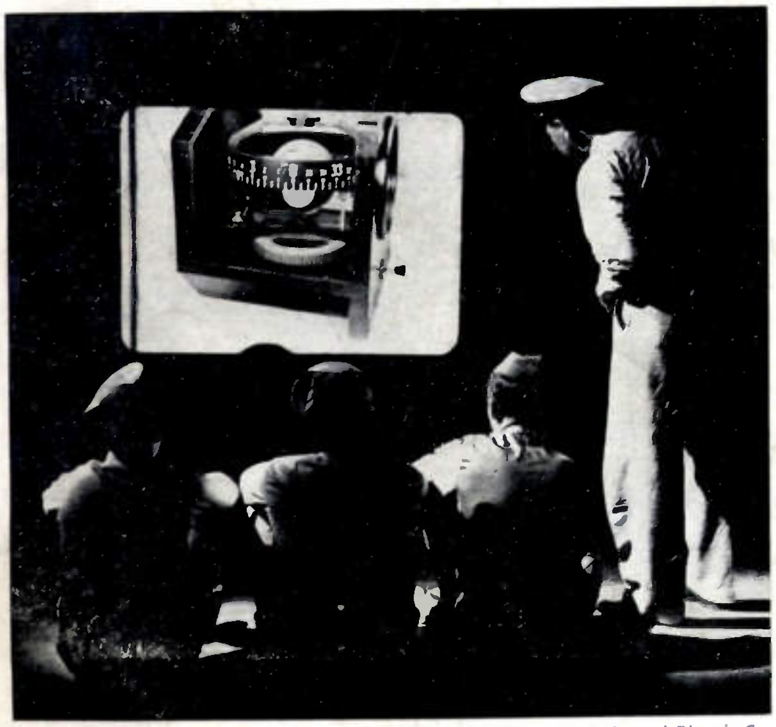
Proceedings of the I·R·E

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

January, 1949
Volume 37 Number 1



U. S. Navy and General Electric Co.

TELEVISION CLASSROOM

At Sands Point, Long Island, lectures are presented to Naval personnel, by television, from special studios four miles away. Comparison with standard instructional methods are in progress.

PROCEEDINGS OF THE I.R.E.

Waves and Electrons Section

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The Institute of Radio Engineers

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with the most extensive line of production-standardized, self-quenching

RADIATION COUNTER TUBES

AMPEREX has the most complete line of standardized types of radiation counter tubes that are actual production line models. If you are working on anything which requires radiation counter tubes, chances are that Amperex can fit you neatly with a tube from our regular line. Save time . . . save money . . . write today for detailed Amperex literature.

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COMING

March 7-10, 1949

The Radio Engineering Show
of the
I.R.E. National Convention
Grand Central Palace, New York City

Last March 14,459 engineers and technically interested people attended the sessions and exhibits of the IRE National Convention and Radio Engineering Show. They saw the worlds largest exhibit of:

- The newest Transmitters, Station Equipment
- Advances in Vacuum Tubes and all kinds of radio-and-electronic components.
- Test Equipment in which a survey proves 68% of the audience showed interest.
- Oscilloscopes, Meters, Central Equipment.
- Audio Systems, Amplifiers, Tape Recorders
- Television Transmitters, Cameras, Receivers
- Materials and Tools for electronic manufacturing —a complete industry picture.

In the coming Convention you can see more, learn more, meet more IRE engineers in four days than ever before. Plan to attend!

Registration, Members, $1.00 Non-Members, $3.00

PROCEEDINGS OF THE I.R.E., January, 1949, Vol. 37, No. 1. 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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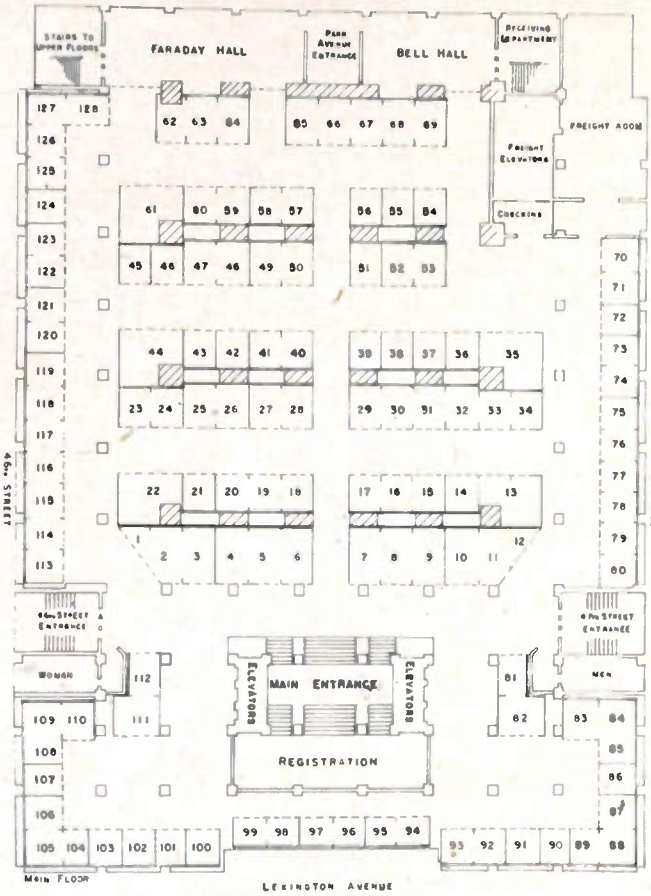
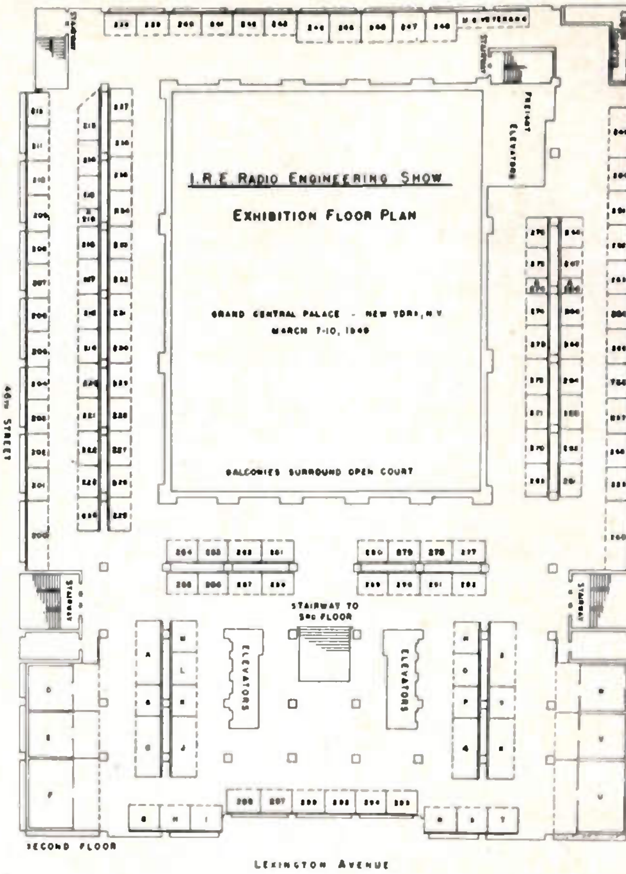
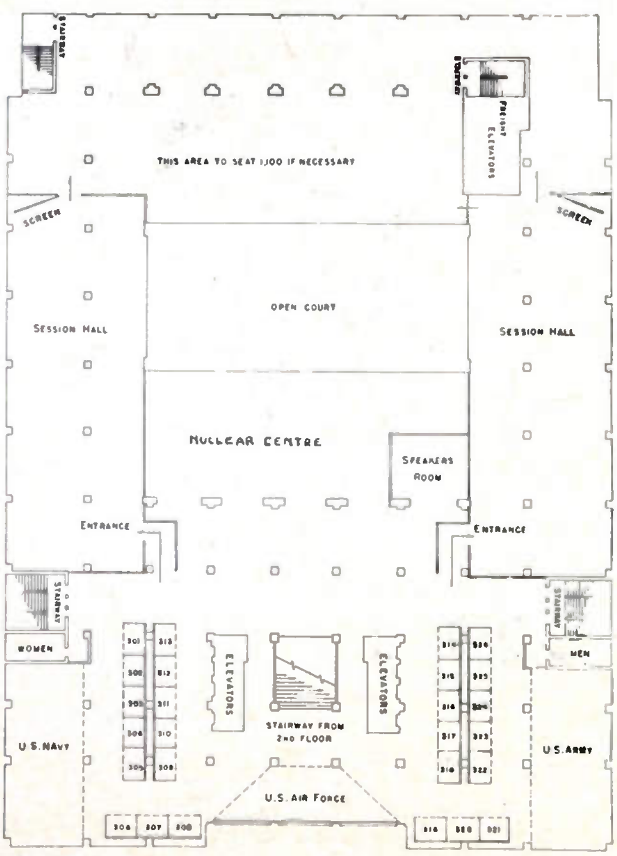
1949 Radio Engineering Show

Grand Central Palace, New York City







36,000 Square Feet of Exhibits
3 Floors—280 Booths

Here 180 exhibitors will demonstrate and present their newest developments for engineers. Some will use combination booths with sound rooms, others group with the government shows on the Third Floor or exhibit in the Nuclear Centre. Some first floor units still available at $4.00 a square foot. Second and third floor units 10′ x 8′ at $3.12½ sq. ft., or $250 a booth. 80% of all space assigned, so act today—write or wire.

J. Robert Marcett, Reservations Mgr.
THE INSTITUTE OF RADIO ENGINEERS
303 West 42nd St., New York 18, New York
Telephone Circle 6-6357

THE ARNOLD ENGINEERING COMPANY

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ALLEGHENY LUDLUM STEEL CORPORATION

147 East Ontario Street, Chicago 11, Illinois

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MYCALEX is today's improved insulation — designed to meet the exacting demands of all types of high-frequency circuits. MYCALEX is unusual in that it possesses a combination of peculiar characteristics that make it ideally suited for insulation in all types of electronic circuits. In tomorrow's designs for communications and industrial control equipment, MYCALEX 410 will be specified more than ever before because of its . . . Low dielectric loss • High dielectric strength • High arc resistance • Dimensional stability over wide humidity and temperature changes • Resistance to high temperatures • Mechanical precision • Mechanical strength • Ability to mold metal inserts in place. If you have any insulation problems, our engineers will be glad to help you in their solutions.

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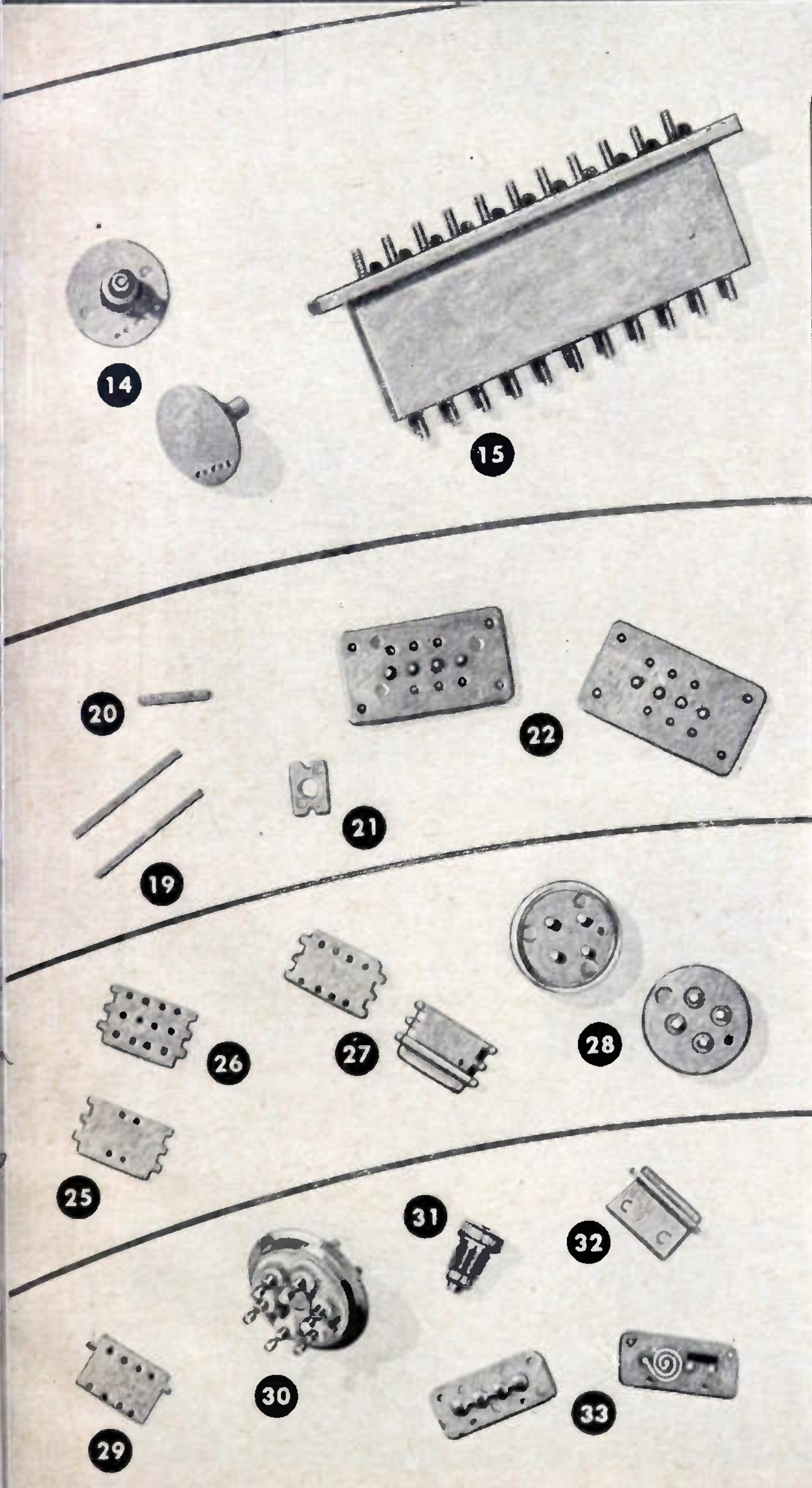
"Owners of 'MYCALEX' Patents"

Plant and General Offices: Clifton, N. J. Executive Offices: 30 Rockefeller Plaza, New York 20, N. Y.

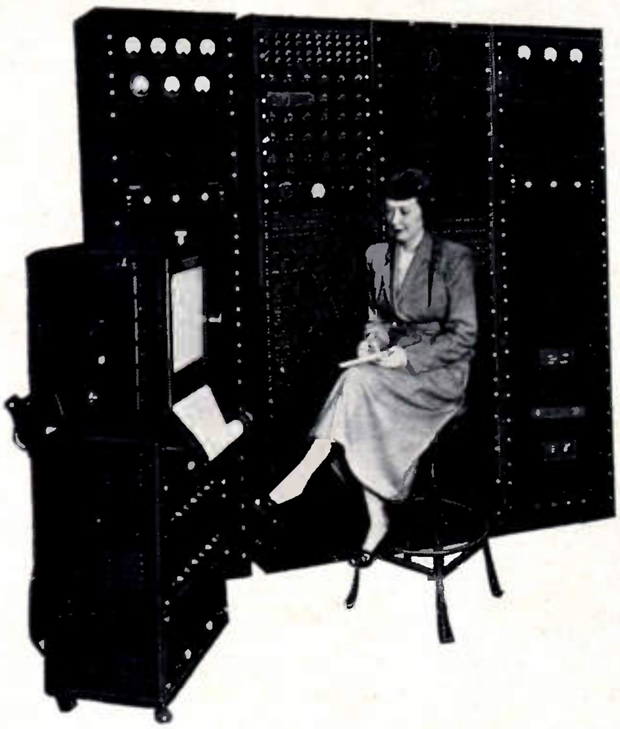
SINCE 1919 MYCALEX THE INSULATOR
TRADE MARK REG. U. S. PAT. OFF.



PART NAME	APPLICATION	INSERTS	MAX. DIMEN.
1 Bushing	Motor Generator	None	1.75"
2 Insulator	Electrical Instrument	None	3.18
3 End Seal	Thermostat Shell	Stainless Steel	3.75
4 Insulator	Electrical Instrument	None	3.00
5 Hermetic Seal	Crystal housing	Nickel and Copper	0.88
6 Hermetic Seal	Crystal housing	Copper	1.09
7 Insulator	Automobile Antenna	None	1.06
8 Bushing	Ignitron	Steel	4.50
9 Stand-Off Insulator	Electronics circuit	Brass	0.56
10 Panel	Television Selector Switch	Silver	1.38
11 Switch Wafer	Television Selector Switch	None	2.31
12 Elbow	Aircraft ignition	Steel and Brass	2.75
13 Lead	Transformer	Monel	1.75
14 Insulator	Polarizing relay	None	1.09
15 Lead through block	Oscillator	Brass	4.69
16 Insulator	Telephone Transmitter	None	0.88
17 Dual Bushing	Oil Burner Transformer	None	3.00
18 Lead	Transformer	Monel	2.50
19 Actuating Bar	Telephone relay	None	1.44
20 Actuating Bar	Telephone relay	None	0.78
21 Spacer	Radio vibrator	None	0.56
22 Panel	Television Selector Switch	None	1.75
23 Spacer	Telephone relay	None	1.00
24 Spacer	Relay	None	0.91
25 Spacer	Telephone relay	None	1.00
26 Spacer	Telephone relay	None	1.00
27 Clamping Plate	Telephone relay	None	1.00
28 Electrode Mounting	Level Indicator	Brass	1.13
29 Spacer	Telephone relay	None	1.00
30 Six Terminal Header	Transformer	Monel	1.42
31 Test jack body	High Frequency Circuits	Monel	0.75
32 Clamping Plate	Telephone relay	None	1.00
33 Printed Circuit Base	Experimental	Silver	1.38

the "electronic brain"...

both A.C. & D.C. voltages accurately held with Sorensen equipment



The Reeves Electronic Analog Computer, known in industry as REAC, is the first standard office size electronic"brain". This machine makes possible the rapid dynamic analysis of high order linear and non-linear differential equations, doing in a few hours computations formerly taking thousands of man-hours.

Nearly all major aircraft companies and many process industries are already making use of this amazing electronic "brain" to speed up computations.

voltage regulator by sorensen A.C. Voltage accuracies in REAC are held to ½ of 1% by a Model 150 Sorensen Voltage Regulator which controls all the servo follow up potentiometers. These limits must be met in order to maintain the high accuracy expected of the computer even though it is subjected to wide line variations.

nobatron by sorensen In the computing amplifiers of REAC are 80 electronic tubes which require ¼ of 1% hum free regulation of the D.C. for filament power. That is ideally done with a Model E-6-40 Nobatron.

If you calibrate meters, need quality control on test lines, work with X-ray equipment or are a research physicist or chemist, there is a *standard* Sorensen A.C. or D.C. unit to solve your voltage problems. With their use you will experience: • *Precise regulation accuracy* • *Excellent wave form* • *Fast recovery time* • *Constant output voltage* and many other advantages. Ask for our latest catalog or put your Voltage Regulation Problems up to our engineers for a specific recommendation.

THE FIRST LINE OF STANDARD ELECTRONIC VOLTAGE REGULATORS.

S

Sorensen at Stamford

SORENSEN & *Company, Inc.*

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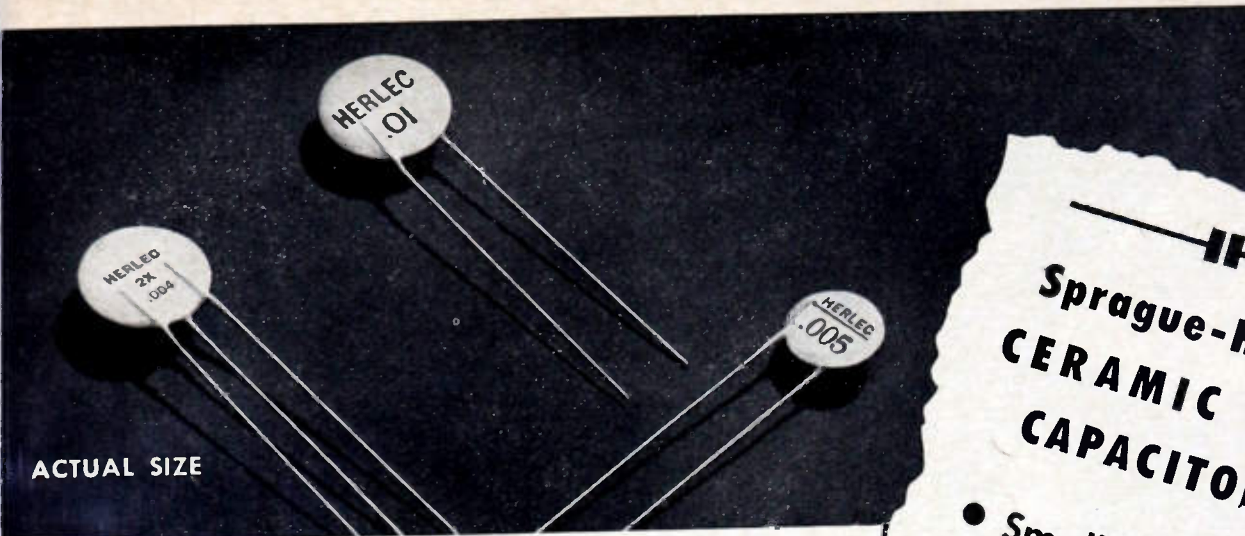
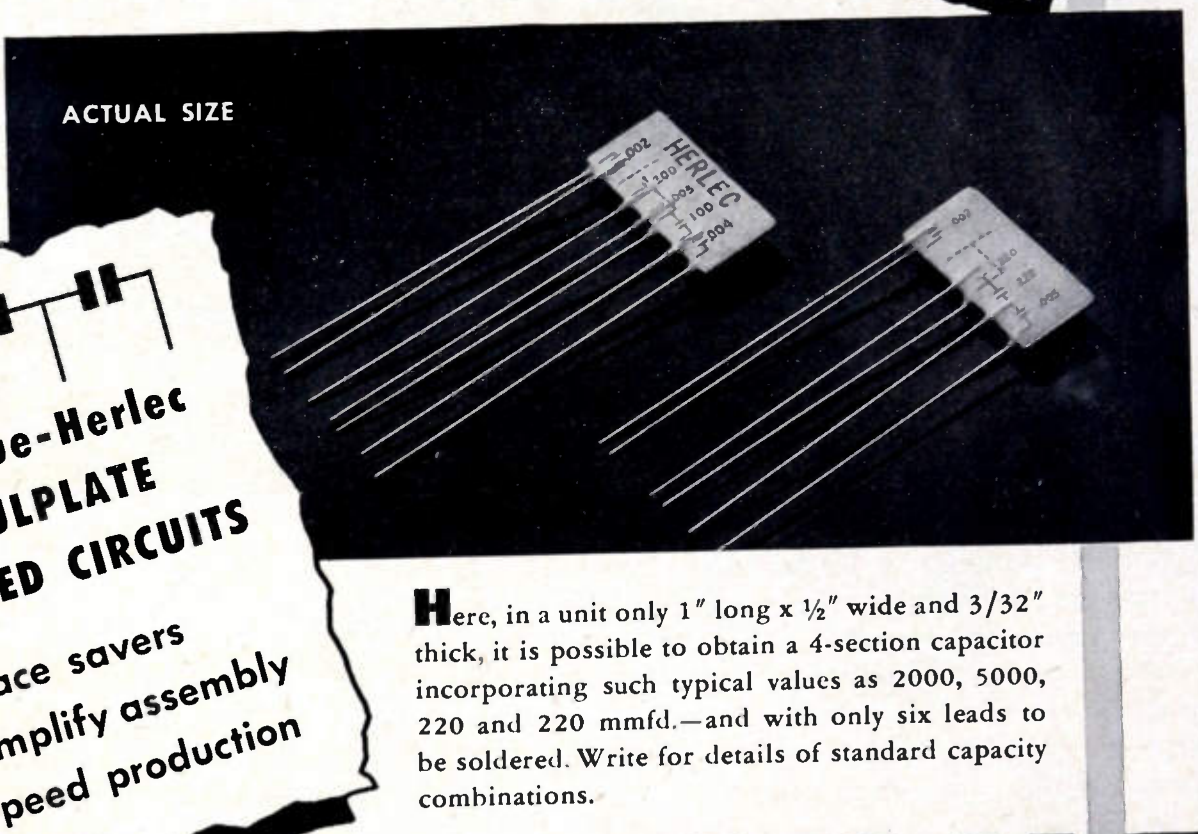
Represented in all principal cities.



Sprague-Herlec
CERAMIC DISC CAPACITORS

- Small in size
- Light in weight

These little by-pass and coupling type ceramic disc capacitors—shown here in actual size—offer distinct size advantages that warrant careful investigation. Now available in a broad range of values in single- and dual-capacity units that assure small size and a minimum of weight.

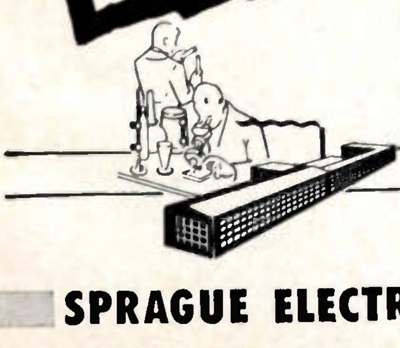


Sprague-Herlec
***BULPLATE**
PRINTED CIRCUITS

- Space savers
- Simplify assembly
- Speed production

*Trademark Applied for

Here, in a unit only 1" long x ½" wide and 3/32" thick, it is possible to obtain a 4-section capacitor incorporating such typical values as 2000, 5000, 220 and 220 mmfd.—and with only six leads to be soldered. Write for details of standard capacity combinations.



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MEASURING SPEED AND ACCURACY

FUNCTION	MODEL	FREQUENCY	CHARACTERISTICS
HARDWARE	10		Binding Post
	14		Flexible coupler, ceramic insulated; permits misalignment of 1/32'' and/or 5°
LOW FREQUENCY STANDARDS	100A	100 kc, 10 kc, 1 kc, 100 cps	Accuracy 3 cps per mc per degree Centigrade
	100B	100 kc, 10 kc, 1 kc, 100 cps	Temperature controlled; accuracy 0.001%
FREQUENCY DIVIDER	110	100 to 10 cps	Controlled by 100A or 100B. Multipliers also available up to 1 mc
RESISTANCE-TUNED OSCILLATORS	200A	35 to 35,000 cps	Output 1 watt into 500 ohms; 1% distortion
	200B	20 to 20,000 cps	Output 1 watt into 500 ohms; 1% distortion
	200C	20 to 200,000 cps	Output 10 volts into 1,000 ohms; 1% distortion
	200D	7 to 70,000 cps	Output 10 volts into 1,000 ohms; 1% distortion
	200H	60 to 600,000 cps	Output 10 mw into a 100 ohm load; 3% total distortion
	200 I	6 to 6,000 cps	Frequency setting closer than 1%; output 10 volts into 1,000 ohms; 1% distortion
	201B	20 to 20,000 cps	Output 3 watts at 1% and 1 watt at ½% distortion into 600 ohms
	202B	½ to 50,000 cps	For low frequency studies. Output 10 volts into 1,000 ohms; 1% distortion
	202D	2 to 70,000 cps	Output 10 volts into 1,000 ohms; 2% distortion
	204A	2 to 20,000 cps	Portable, battery-operated; output 5.0 volts to 10,000 ohm load; 1% distortion
AUDIO SIGNAL GENERATORS	205A	20 to 20,000 cps	Output 5 watts, 1% distortion into impedances of 50, 200, 600, 5,000 ohms. Output VTVM and 110 db attenuator, 1 db steps
	205AG	20 to 20,000 cps	Same as 205A, plus separate VTVM for complete gain measurements
	205AH	1 to 100 kc	Output 5 watts, 3% distortion into 50, 200, 500, 5,000 ohm impedances. Output VTVM and 110 db attenuator, 1 db steps
	206A	20 to 20,000 cps	Output +15 dbm with less than 0.1% distortion into 50, 150, 600 ohm impedances. Output VTVM and 111 db attenuator in 0.1 db steps
SQUARE WAVE GENERATOR	210A	20 to 10,000 cps	Output 50 volts peak to peak; 1,000 ohm internal impedance; 70 db attenuator, 5 db steps
WAVE ANALYZER	300A	30 to 16,000 cps	Variable selectivity; measurement range 1 mv to 500 volts; 5% accuracy

-hp- Model 200C
Resistance Tuned Oscillator

-hp- Model 206A
Audio Signal Generator

-hp- Model 300A
Wave Analyzer

-hp- Model 330B
Distortion Analyzer

-hp- Model 616A
UHF Signal Generator

THROUGHOUT THE ELECTRONIC FIELD

FUNCTION	MODEL	FREQUENCY	CHARACTERISTICS
DISTORTION ANALYZERS	320A	400 cps and 5 kc	Measures total distortion as low as 0.1%. 70 db attenuator, 1 db steps for comparison
	320B	50, 100, 400 cps and 1, 5 and 7.5 kc	Same as 320A
	325B	30, 50, 100, 400, 1,000 cps; 5, 7.5, 10 and 15 kc	Measures total distortion as low as 0.1%. Input amplifier and complete VTVM each usable separately
	330B	Any frequency 20 to 20,000 cps	Similar to 325B but measures at any frequency and includes AM detector
	330C	Any frequency 20 to 20,000 cps	Similar to 330B, no AM detector. Meter has VU characteristics to meet FCC requirements for FM broadcasting
FM BROADCAST MONITOR	335B	88 to 108 mc	FCC approved. Monitors carrier frequency and modulation. High fidelity output for aural monitoring
ATTENUATORS	350A	Max 100 kc	110 db, 1 db steps; 5 watts, 500 ohm level. Bridged T type. Accuracy 1 db in 50 db at 100 kc
	350B	Max 100 kc	Same as 350B but 600 ohm level
VACUUM TUBE VOLTMETERS AND ACCESSORIES	400A	10 cps to 1 mc	Nine ranges 0.03 to 300 volts full scale. Accuracy ±3% to 100 kc, ±5% to 1 mc. Average reading. Calibrated in rms.
	400B	2 cps to 100 kc	Same as 400A with response flat to 2 cps. 10 megohm input impedance
	400C	20 cps to 2 mc	Twelve ranges 0.001 to 300.0 volts full scale; accuracy ±3% to 100 kc, ±5% to 2 mc; 10 megohm input impedance; average reading; calibrated in rms volts; may be used as 54 db amplifier
	404A	2 to 50,000 cps	Portable, battery-operated; eleven ranges; 0.003 to 300 volts full scale; accuracy ±3% to 20 kc; 10 megohm input impedance
	410A	20 cps to 700 mc	AC: six ranges 1 to 300 volts. DC: seven ranges 1 to 1,000 volts. Resistance: seven ranges 0.2 ohm to 500 megohms
	415A	300 to 2,000 cps	Standing Wave Indicator for use with a bolometer or crystal rectifier; standard frequency 1000 cps, others on special order
	455A	to 1,000 mc	Connects probe of 410A across 50 ohm transmission line. Type N fittings
	458A	to 1,000 mc	Connects probe of 410A to open end of 50 ohm transmission line. Type N fittings
AMPLIFIERS	450A	10 to 1,000,000 cps	40 db and 20 db stabilized gain. Input impedance 1 megohm shunted by approximately 15 uuf.
ELECTRONIC FREQUENCY METER	500A	5 cps to 50 kc	Ten ranges, ±2% accuracy. Input 0.5 to 200 volts
ELECTRONIC TACHOMETER	505A	300 to 3,000,000 rpm	Ten ranges, ± 2% accuracy
	505B	5 to 50,000 rps	Same as 505A except calibrated in rps
SIGNAL GENERATORS	610A	500 to 1,350 mc	Calibrated output 0.1 microvolt to 0.1 volt. Internal pulse modulation. Direct calibration
	616A	1,800 to 4,000 mc	Direct reading. Pulse modulation, CW and FM. Calibrated output 0.1 microvolt to 0.2 volts
	650A	10 cps to 10 mc	Direct reading. Six bands. Output 3 volts to 600 ohm load. VTVM and output attenuator
POWER SUPPLY	710A		Any dc voltage 180 to 360 for 0 to 75 ma load; approximately 1% regulation. Also 6.3 volts, 5 amps ac.

-hp- Model 400A
Vacuum Tube Voltmeter

-hp- Model 410A
Vacuum Tube Voltmeter

-hp- Model 610A
UHF Signal Generator

-hp- Model 650A
Audio Signal Generator

-hp- Model 335B
FM Monitor

COMPANY

Export: FRAZAR & HANSEN, 301 Clay Street, San Francisco 11, California

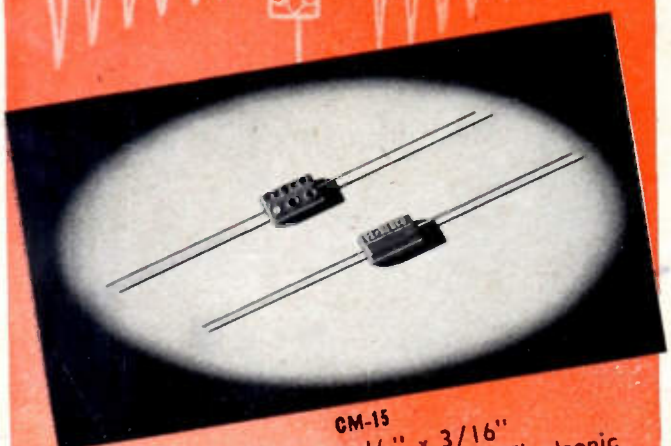
NO OSCILLATING CIRCUIT IS BETTER THAN ITS CAPACITOR(S)

NO CAPACITOR IS BETTER THAN

El-Menco

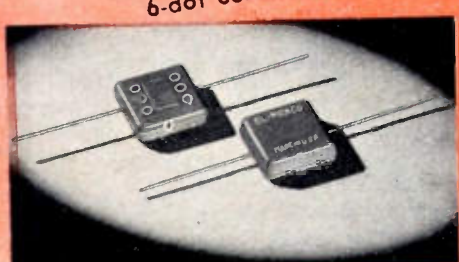
This is why more and more manufacturers of Radio, Electrical and Electronic equipment are turning to EL-MENCO. Engineers specify EL-MENCO when they want small size, high capacity and unquestioned performance in capacitors.

Each tiny EL-MENCO capacitor does the job it was designed to do, with absolute accuracy and complete dependability. When you want to protect your products' reputation for performance perfection — specify EL-MENCO for capacitors that really satisfy.



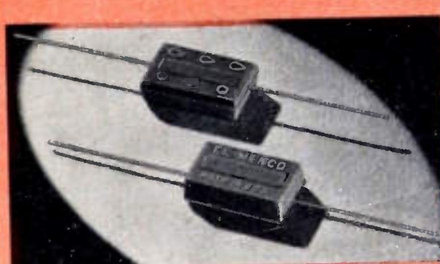
CM-15

Actual Size 9/32" x 1/2" x 3/16"
For Television, Radio and other Electronic Applications
2 — 420 mmf. cap. at 500v DCA
2 — 525 mmf. cap. at 300v DCA
Temp. Co-efficient ±50 parts per million per degree C for most capacity values.
6-dot color coded.



CM-30

Available in A, B, C, D and E Characteristics. Minimum tolerance (A and B) 5%. Minimum tolerance (C, D, E) 1%. 470 to 6,200 mmf. cap. at 500 DC working voltage*. 6-dot color coded.



CM-40

Available in A, B, C, D and E Characteristics. Minimum tolerance (A and B) 5%. Minimum tolerance (C, D, E) 1%. 3,300 to 7,500 mmf. cap. at 500 DC working voltage*. 8,200 to 10,000 mmf. cap. at 300 DC working voltage*. 6-dot color coded.

ARCO ELECTRONICS, INC., 135 Liberty Street, New York, N. Y.
Sole agent for jobbers and distributors in U. S. and Canada.

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

*Can be made in other capacity values and voltage ratings

Foreign Radio and Electronic Manufacturers communicate direct with our Export Dept. at Willimantic, Conn. for information.

Write on your firm letterhead for Catalog and Samples

El-Menco

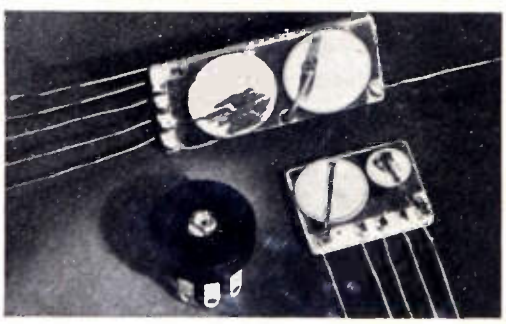
MOLDED MICA **MICA TRIMMER**
CAPACITORS

PROGRESS REPORT ON P.E.C.*

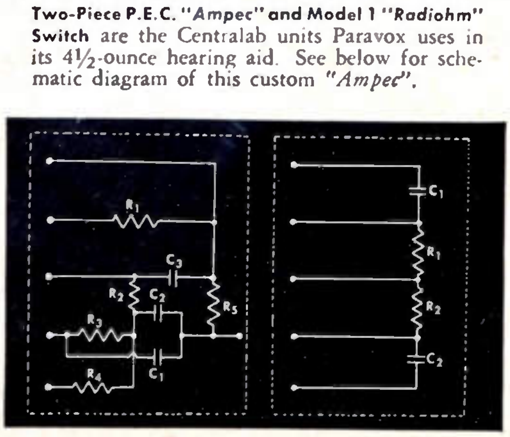
How Paravox uses Centralab's amazing *Printed Electronic Circuit* to build small, light, quickly-assembled hearing aids!

Note how CRL's special two-piece "Ampec" fits into the completely-molded Paravox unit. This plastic chassis makes it possible to dismantle or assemble a Paravox hearing aid in one minute — without the use of tools.

Models courtesy of Paravox, Inc.



Two-Piece P.E.C. "*Ampec*" and Model 1 "*Radiohm*" Switch are the Centralab units Paravox uses in its 4½-ounce hearing aid. See below for schematic diagram of this custom "*Ampec*".



*Centralab's "Printed Electronic Circuit" — Industry's newest method for improving design and manufacturing efficiency!

TIME, space and material savers! That's how Paravox, Inc., Cleveland hearing aid manufacturer, describes Centralab's revolutionary P.E.C. *"Ampecs"*. These tiny audio-amplifying units save time for Paravox by eliminating many assembling operations. They save space and material by reducing the number of components needed. What's more — like all CRL *Printed Electronic Circuits* — they are rugged, dependable, resistant to temperature and humidity.

Integral ceramic construction: Each *Printed Electronic Circuit* is an integral assembly of *"Hi-Kap"* capacitors and resistors closely bonded to a steatite ceramic plate and mutually connected by means of metallic silver paths "printed" on the base plate. All leads are always the same length, each plate is an exact duplicate of the original or "master."

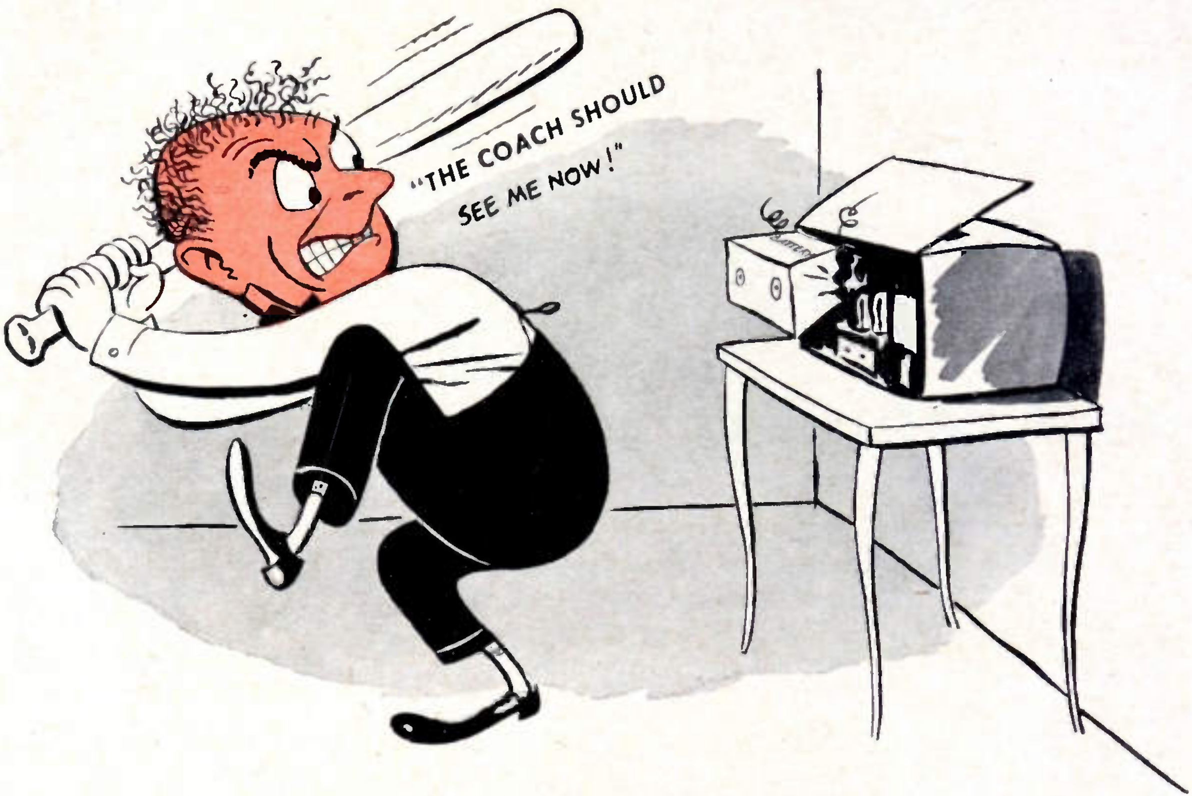
This outstanding hearing aid development, illustrated above, was the product of close cooperation between Centralab and Paravox engineers. Working with your engineers, Centralab may be able to fit its *Printed Electronic Circuit* to your specific needs. Write for complete information, or get in touch with your nearest Centralab Representative.

LOOK TO **Centralab** IN 1948!

CRL

Division of GLOBE-UNION INC., Milwaukee

You can use a ball bat to make it fit...

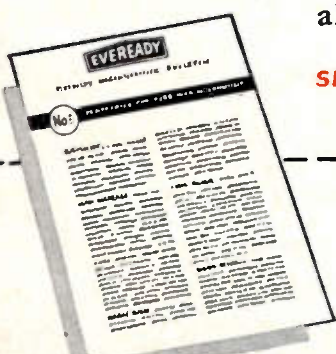


But it's simpler to design the radio around the battery!

NO MATTER what size portable receiver you are designing, you'll find a compact, long-lasting "Eveready" battery to fit it. By designing your portables around "Eveready" batteries, you automatically insure that the user will get long-playing life and—very important—he can get replacements anywhere, anytime. "Eveready" brand batteries are always top-quality, always easy for the user to obtain.

SEND COUPON BELOW FOR LATEST BATTERY DATA

The registered trade-mark "Eveready" distinguishes products of National Carbon Company, Inc.



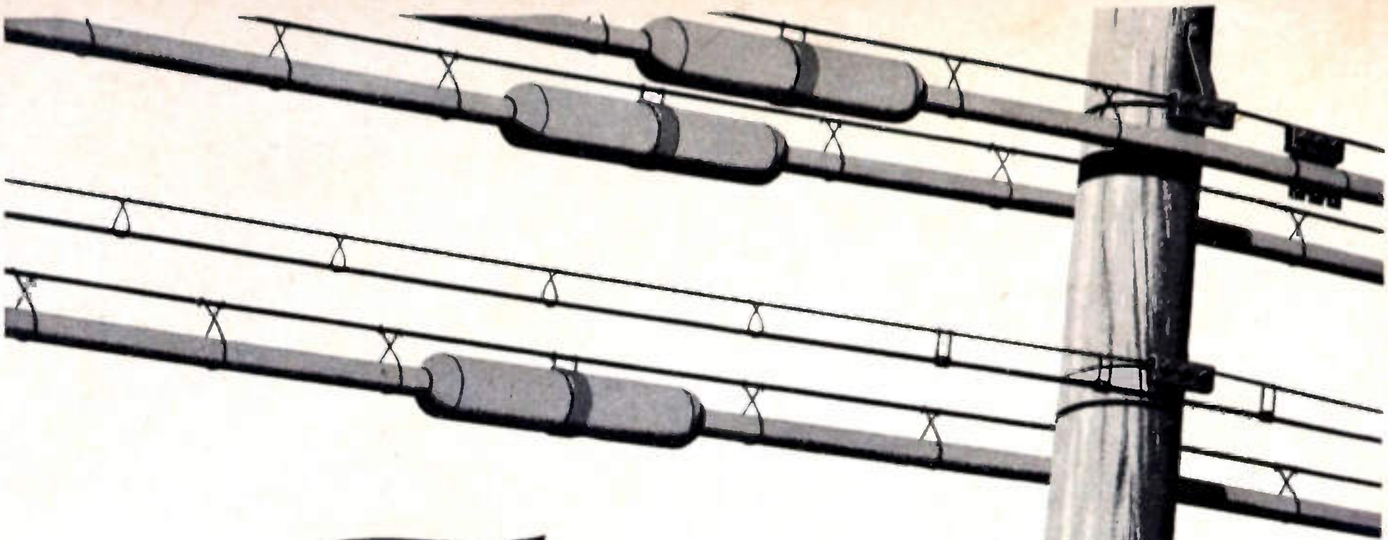
Please send me, free of charge, the 1949 Revision of "Battery Engineering Bulletin No. 1" which contains latest battery data for portable receivers and includes specifications for six completely new batteries of interest to designers.

Name ____________________

Street ____________________

City __________ *State* __________

NATIONAL CARBON COMPANY, INC.
30 East 42nd Street
New York 17, N.Y.
Unit of Union Carbide and Carbon Corporation
UCC



The case of the Creeping Sleeve

Lead sheathing on telephone cable meets many stresses — the tug of its own weight, wind pressure, contraction and expansion from cold and heat. Then, too, there's the pressure of nitrogen gas put in Long Distance cable to warn of sheath breaks and keep out moisture.

And, sometimes, lead is subject to "creep"— a permanent stretching — even when the stress is but a fraction of the normal tensile strength. Creep is especially likely at the lead sleeves used where two lengths of cable are joined. The sleeve may stretch and break open exposing telephone circuits to the elements.

So Bell Laboratories scientists have developed methods to test and control creep. In a special testing room, weights are applied to scores of samples of lead, under controlled conditions. Exact records of the amount of creep are obtained with a precision instrument.

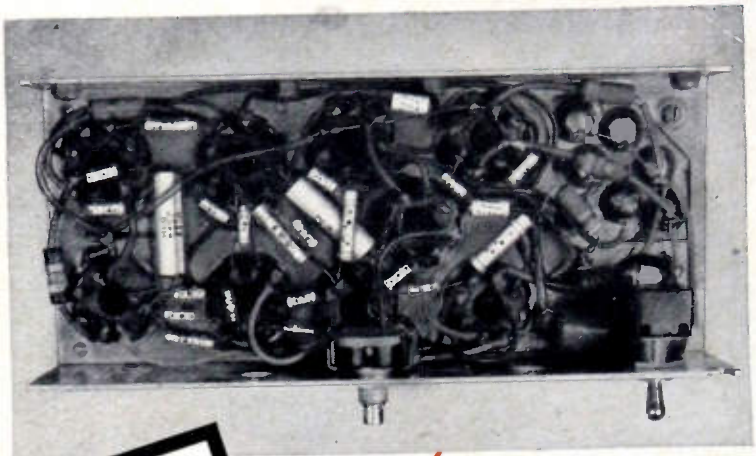
Years of careful study have produced a lead composition which resists creep and yet has all the other properties required of sleeves. This means better telephone service for you and helps give that service at lowest possible cost. It is an example of the way Bell Telephone Laboratories scientists study and improve every part of the great telephone plant.

BELL TELEPHONE LABORATORIES

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

HI-Q COMPONENTS

Specify for MINIATURIZATION



FOR superior performance in a limited space, the utilitarian compactness of **HI-Q** components is sure to meet your enthusiastic approval. Although **HI-Q** components are compact, there is no sacrifice of accuracy, dependability or uniformity. Each component meets or surpasses rigid standards for capacity, tolerance and performance. This compactness is accomplished through application of up-to-the-minute processing techniques, combined with use of highest quality materials and complete control of quality throughout all stages of manufacture. Specify **HI-Q** for *maximum* efficiency in a *minimum* amount of space.

HI-Q MINIATURE G. P. TUBULARS

G. P. Miniatures providing capacities from 5 mmf to 33,000 mmf will cover the majority of your by-passing problems. By the use of our new body 41 the above extended capacity ranges are now available. These smaller units provide closer coupling of leads thus insuring the minimum of inductance and highest self resonant frequencies.

Illustration at left is actual size.

HI-Q DISK CAPACITORS

BPD Another example of accurate and dependable miniaturization is this high dielectric by-pass, blocking or coupling Hi-Q Disk Capacitor. Many applications where physical shape is more adaptable than the tubular unit. Available in two standard capacities:

Type BPD-5: .005 mfd guar. min.
Type BPD-10: .01 mfd. guar. min.
Type BPD-1.5: .0015 mfd. guar. min.

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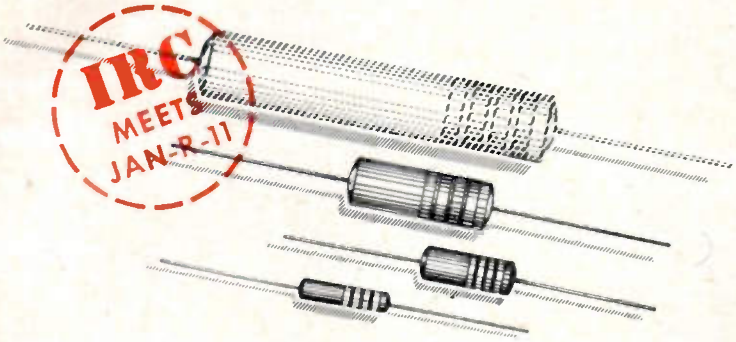

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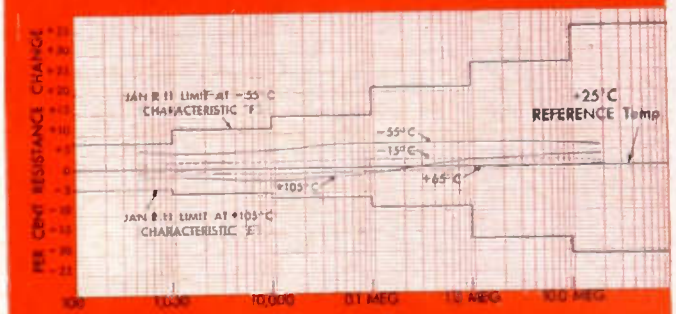

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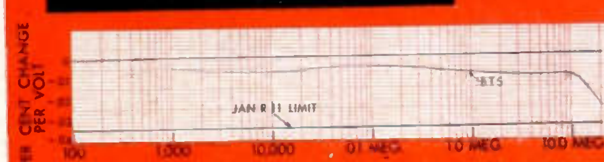

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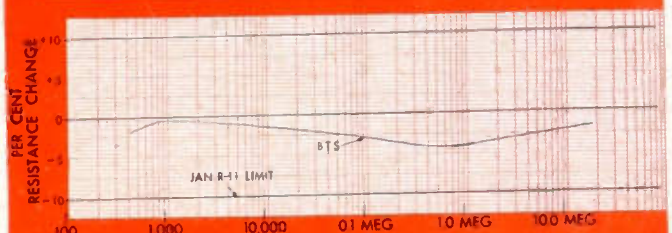

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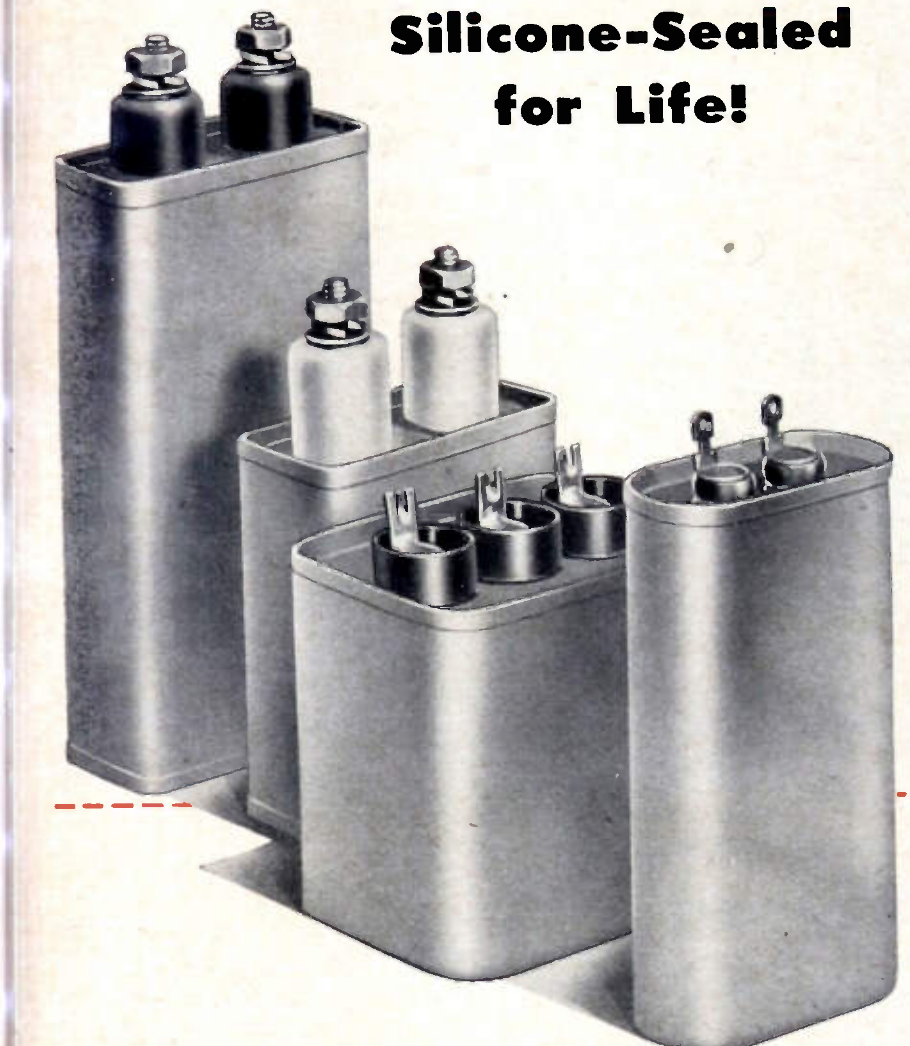

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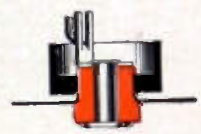

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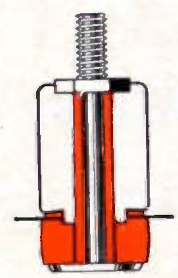

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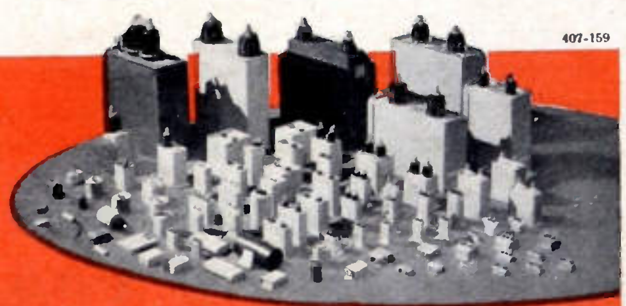

PROCEEDINGS OF THE I.R.E.

(Including the WAVES AND ELECTRONS Section)

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 37 *January, 1949* NUMBER 1

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

WAVES AND ELECTRONS SECTION

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

Stuart L. Bailey

PRESIDENT, 1949

Stuart L. Bailey was born in Minneapolis, Minnesota, on October 7, 1905. He received the degree of Bachelor of Science in Electrical Engineering from the University of Minnesota in 1927 and the Master of Science degree from the same institution in 1928. During his undergraduate years he was active on the staff of W9X1, an experimental station run by the University, and while taking his graduate work he was chief engineer of WLB, owned and operated by the University of Minnesota. His master's thesis was on the subject of radio field-intensity measurement.

In the summer of 1928, Mr. Bailey became assistant radio engineer with the airways division of the United States Department of Commerce. His experience there included work on radio aids to marine and air navigation. He initiated and supervised the early work of the Lighthouse Service on radio-controlled fog signals and participated in the development of the visual radio range for use on the airways of the United States.

In the summer of 1929, Mr. Bailey went to Panama, where he installed two automatic marine radio beacons, one at the entrance to Cristobal Harbor and the other at Cape Mala, 120 miles south of Balboa.

In September 1930 he joined with C. M. Jansky, Jr., to form the consulting engineering firm of Jansky and Bailey. Mr. Bailey's activities in the consulting field have been on both general allocation problems and specific engineering guidance for broadcast stations and commercial operating companies. He has had charge of all of the laboratory activities of the firm.

During World War II, Mr. Bailey was in charge of all government-contract work done by the firm of Jansky and Bailey, a great deal of which was done under the Office of Scientific Research and Development. This work involved a detailed study of all factors affecting mobile short-range radio communication, a study of the effect of hills and trees as obstructions to radio transmission from 4 to 116 megacycles, and detailed analysis of electronic equipment to determine certain characteristics important to its operation by the armed services. In addition, under a Signal Corps contract, he supervised the work on measurement of many existing and proposed radio antennas for use by the ground forces.

Mr. Bailey has been active in the development of frequency modulation broadcasting, supervising the technical operations of radio station W3X0, an experimental FM broadcasting station operated by Jansky and Bailey from 1938 until 1945.

He became an Associate Member of The Institute of Radio Engineers in 1928, a Member in 1936, Senior Member in 1943, and was advanced to the grade of Fellow in 1943. He has been a member of the Committee on Wave Propagation since 1937 and was a member of the Admissions Committee in 1943 and 1944. He was appointed a member of the Board of Directors of the Institute in 1943 and 1944 and was elected to the Board for a three-year term beginning in 1945. He has been a member of the Executive Committee since 1945. He was appointed Treasurer of The Institute of Radio Engineers in 1948.

He is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

Thoughtful and analytical observers of any human institution may sometimes see more of its scope and philosophy than those who are more closely and directly involved in that institution. The following analysis of the purposes and accomplishments of the PROCEEDINGS OF THE I.R.E. admirably presents some of the fundamental and underlying characteristics and achievements of that publication. It has been prepared by a Fellow of the IRE, who is a member of several of the IRE Editorial Groups, and is also Manager of Physics Laboratories of Sylvania Electric Products Inc.—*The Editor*.

The Place of the Proceedings

ROBERT M. BOWIE

Through the 47,000 pages of the PROCEEDINGS has flowed the life blood of a profession. For thirty-seven years its circulation has carried to our growing society the building blocks of scientific knowledge on which our profession, and the industry which it serves, are based. The circulation of the PROCEEDINGS has left a heritage rich in the fundamental concepts and knowledge of radio, electronics, and related sciences. These are ours to guide us in our further advances in both the science and the technology of our profession. The passing years have not only seen the recording of wisdom in succeeding issues, but have witnessed a broadening of the scope and purposes which it serves.

As our Institute has grown from the small initial group of under 50 in 1912 to the present membership of some 23,000, the publication needs have broadened. The PROCEEDINGS of earlier years was primarily a scientific source journal, the need for which continues undiminished, for the pages thus devoted serve as a medium to disseminate fundamental precepts which form the foundation of further scientific progress and the basis of industrial advances in technology. In the PROCEEDINGS of today, some five other purposes are now clearly evident. Some have developed over the years, while others are of more recent origin.

In an organization so large and so far-flung as ours, the wide circulation of its journal must carry with it reports of Institute activities and professional news and reviews, together with those life-giving human notes which lend subjective interest to an essentially objective science. These have found an important place in the PROCEEDINGS.

About the science just mentioned has grown a vast industry, many of whose transactions have technical impact, and receive notice in the pages of our PROCEEDINGS, some in the form of announcements of products, others as selected reports from the Radio Manufacturers' Association.

Stemming from this science also is a technology giving rise in turn to a technical literature, which, though of a more transitory interest, is nevertheless a vital part of the life blood of the profession. This, too, has its place in the PROCEEDINGS.

As a science and its associated technology gain in stature and age, their vital literature becomes so extensive that recording alone can no longer serve the profession which created it. Thus is born the need for systematic abstracts. Their advent marks the maturity of a profession.

There may come in the life of any society, as there has been recently in ours, the need to acquaint its membership with pre-eminent advances in a sister science which are destined to draw heavily upon the science and technology of that society. Such an end may best be served by well-planned articles of a tutorial nature. The joining of the interests of electronics and nucleonics but marks the beginning of a new era in human progress in which our profession will play an important part.

The very PROCEEDINGS which gives life to our society by its circulation is a product of that society. It can not be expected to be better than the contributions of its authors and its editors. To each of us, therefore, falls his share of responsibility for the excellence of the journal which he receives and to which he may contribute. In this interest may each of us keep constantly in mind the importance of reporting worthy scientific or technical advances, together with the importance of thoughtful and careful preparation. Certainly a scientific or technical advance deserves a presentation commensurate with its value. A lesser one may cause its import to be lost.

To some is granted the privilege of serving in an office or upon a Committee charged with responsibility for some aspect of the affairs of the PROCEEDINGS. This responsibility must not rest lightly upon such incumbents, for the content of our journal is largely in their hands. Some must sit in judgment upon the contributions of fellow members of the profession. May they exercise judgment, foresight, and courage, touched with forbearance, in upholding the professional standards of our journal!

To each of us is granted the right and duty to participate, through the democratic processes of the Institute, in the guidance of its affairs. Thus through the years to come, as in the past, those in authority who guide the destinies of our PROCEEDINGS will welcome the privilege of serving the membership's interests.

The Electron-Wave Tube—A Novel Method of Generation and Amplification of Microwave Energy*

ANDREW V. HAEFF†, FELLOW, IRE

Summary—**A novel method of generation and amplification of microwave energy is described, based on space-charge interaction of electron streams of different velocities. The electron streams are generated within a common space which, in the presence of electrons, behaves as a unidirectional transmission medium having negative attenuation. The basic theory of the new method is developed, and formulas and curves are given for propagation constant, electronic, gain, and bandwidth which can be achieved with the device based on the new method. This device is called the "electron-wave tube."**

The design and performance of the "two-beam"-type and the "single-beam"-type experimental electron-wave tubes are described. Electronic gains of the order of 80 db at a frequency of 3000 Mc and electronic bandwidths of over 30 per cent have been observed in experimental electron-wave tubes. It is pointed out that, since no passive circuit is required in the amplifying region of the electron-wave tube, the new method is important in the development of tubes for millimeter waves.

INTRODUCTION

FOR A LONG time it has been recognized by workers in the field of ultra-high-frequency tube research that a most serious obstacle to the extension of the high-frequency limit of tube operation is an apparent necessity to use resonant circuits or wave-guiding structures of smaller and smaller dimensions as the frequency is increased. This requirement arises from the fact that the tube structures have to be arranged so that kinetic energy of electrons can be efficiently converted into the energy of the high-frequency electric fields which are supported by these structures. The reduction in tube size usually leads to a reduction in useful power when maximum power dissipation on collector electrodes and maximum emission current density on cathodes are approached.

These fundamental considerations made it extremely important to search for new methods of amplification and generation of microwave energy. Such a method was found in an arrangement where the high-frequency energy is amplified or generated in the process of space-charge interaction of electron streams without the use of any field-supporting or waveguiding structure.

This paper describes the novel method in some detail, gives its basic theory, and describes some experimental arrangements and experimental results.

INTERACTION BETWEEN STREAMS OF ELECTRONS OF DIFFERENT ENERGIES

When a high-temperature body is placed close to a low-temperature body, the heat energy flows in the direction which tends to equalize the temperatures of the two bodies in contact. This is one case of a general law of statistics that the entropy tends to increase. If an electron stream of high energy is projected in the proximity of another electron stream of lower energy, then, by the mechanism of elastic collisions between fast and slow electrons, there will be a gradual interchange of energy between the two streams, with the result that, on the average, the fast electrons are slowed down and the slow ones are speeded up. The nature of this process is similar to elastic scattering of high-energy particles by other particles which are either stationary or moving slowly. In the case of electron scattering, the forces acting between the electrons obey Coulomb's law. As a consequence, the mechanism of scattering of electrons by electrons can be described in terms of the familiar concepts of electromagnetic theory.

Consider a uniform stream of electrons of charge density ρ_1 moving in the Z direction with the velocity v_1. Any disturbance of the stream will produce electron waves in the stream. By "electron waves" is meant either space-charge-density fluctuations with associated electric-field fluctuations, or velocity fluctuations of electrons in the stream. The phase velocity of these waves depends upon the space-charge density. Two waves are usually associated with any disturbance. One of these waves has its phase velocity somewhat higher than the velocity of translation of electrons in the stream, and the other wave has velocity lower than the translational velocity.

If another stream of electrons of density ρ_2 moving with velocity v_2 is injected near or into the space occupied by the first electron stream, the electron waves interact so that the phase velocities of electron waves in the two streams are modified, and an interchange of energy between the two streams takes place. It is most convenient for the purposes of analysis to consider the two streams of electrons as components of one inhomogeneous stream. The space-charge densities ρ, current densities i, and velocities v can be expressed as:

$$\left.\begin{aligned} \rho_1' &= \rho_1 + \tilde{\rho}_1 \\ \rho_2' &= \rho_2 + \tilde{\rho}_2 \end{aligned}\right\} \tag{1}$$

$$\left.\begin{aligned} v_1' &= v_1 + \tilde{v}_1 \\ v_2' &= v_2 + \tilde{v}_2 \end{aligned}\right\} \tag{2}$$

$$\left.\begin{aligned} i_1' &= i_1 + \tilde{\imath}_1 = i_1 + \tilde{\rho}_1 v_1 + \tilde{v}_1\rho_1 \\ i_2' &= i_2 + \tilde{\imath}_2 = i_2 + \tilde{\rho}_2 v_2 + \tilde{v}_2\rho_2 \end{aligned}\right\}. \tag{3}$$

In these expressions the first symbol represents average or dc quantities, and the second symbol represents first-order quantities which can be assumed to vary

* Decimal classification: R339.2. Original manuscript received by the Institute, September 29, 1948.
† Naval Research Laboratory, Washington, D. C.

with time and distance in an exponential manner similar to variations of the voltage V:

$$V = V_0 e^{-\Gamma Z + j\omega t} \tag{4}$$

where Γ is the propagation constant, and ω is the frequency of the disturbance, or signal, having an initial value V_0 at $Z=0$.

The fundamental equations which are required in the analysis are:

(a) The Poisson's equation:

$$\nabla \cdot D = \rho, \quad \text{or} \quad \epsilon \frac{\partial^2 V}{\partial Z^2} = \rho_1' + \rho_2', \tag{5}$$

or

$$\epsilon \Gamma^2 V = \tilde{\rho}_1 + \tilde{\rho}_2;$$

(b) the equation expressing conservation of electric charge:

$$\frac{\partial \rho}{\partial t} + \frac{\partial i}{\partial Z} = 0,$$

or

$$\left.\begin{aligned} j\omega_1 \tilde{\rho} &= \Gamma \tilde{i}_1 \\ j\omega \tilde{\rho}_2 &= \Gamma \tilde{i}_2 \end{aligned}\right\} ; \tag{6}$$

and (c) the force equation:

$$m \frac{dv}{dt} = e \frac{\partial V}{\partial Z} \quad \text{where} \quad \frac{dv}{dt} = \frac{\partial v}{\partial t} + \frac{\partial v}{\partial Z}\frac{\partial Z}{\partial t} = \frac{\partial v}{\partial t} + v \frac{\partial v}{\partial Z}$$

$$\left.\begin{aligned} j\omega \tilde{v}_1 - v_1 \Gamma \tilde{v}_1 &= -\frac{e}{m} \Gamma V \\ j\omega \tilde{v}_2 - v_2 \Gamma \tilde{v}_2 &= -\frac{e}{m} \Gamma V \end{aligned}\right\} . \tag{7}$$

By using the relations (3) in (6), we obtain

$$\left.\begin{aligned} j\omega \tilde{\rho}_1 &= \Gamma(\tilde{\rho}_1 v_1 + \tilde{v}_1 \rho_1) \\ j\omega \tilde{\rho}_2 &= \Gamma(\tilde{\rho}_2 v_2 + \tilde{v}_2 \rho_2) \end{aligned}\right\} . \tag{8}$$

Solving for $\tilde{v}_1$ and $\tilde{v}_2$, we get

$$\left.\begin{aligned} \tilde{v}_1 &= \frac{\tilde{\rho}_1}{\Gamma \rho_1}(j\omega - \Gamma v_1) \\ \tilde{v}_2 &= \frac{\tilde{\rho}_2}{\Gamma \rho_2}(j\omega - \Gamma v_2) \end{aligned}\right\} . \tag{9}$$

Substituting (9) into (7),

$$\left.\begin{aligned} \frac{\tilde{\rho}_1}{\Gamma \rho_1}(j\omega - \Gamma v_i)^2 &= -\frac{e}{m} \Gamma V \\ \frac{\tilde{\rho}_2}{\Gamma \rho_2}(j\omega - \Gamma v_2)^2 &= -\frac{e}{m} \Gamma V \end{aligned}\right\} . \tag{10}$$

Solving for $\tilde{\rho}_1$, and $\tilde{\rho}_2$,

$$\left.\begin{aligned} \tilde{\rho}_1 &= \frac{\left(\frac{e}{m}\rho_1\right)}{(\omega + j\Gamma v_1)^2} \Gamma^2 V \\ \tilde{\rho}_2 &= \frac{\left(\frac{e}{m}\rho_2\right)}{(\omega + j\Gamma v_2)^2} \Gamma^2 V \end{aligned}\right\} . \tag{11}$$

Finally, the substitution of (11) into (5) results in the following expression, which defines the propagation constant Γ in terms of the frequency ω, velocities v_1 and v_2 of the two electron streams, and the corresponding space-charge densities ρ_1 and ρ_2:

$$\Gamma^2 V = \left[\frac{\left(\frac{e}{m}\frac{\rho_1}{\epsilon}\right)}{(\omega + j\Gamma v_1)^2} + \frac{\left(\frac{e}{m}\frac{\rho_2}{\epsilon}\right)}{(\omega + j\Gamma v_2)^2} \right] \Gamma^2 V. \tag{12}$$

Excluding trivial solutions corresponding to conditions that either $\Gamma=0$ or $V=0$, and replacing space-charge densities by electron plasma frequencies ω_1 and ω_2 from defining expressions,

$$\frac{e}{m}\frac{\rho_1}{\epsilon} = \omega_1^2 \quad \text{and} \quad \frac{e}{m}\frac{\rho_2}{\epsilon} = \omega_2^2, \tag{13}$$

we obtain the equation from which the propagation constant of the inhomogeneous stream can be obtained:

$$1 = \frac{\omega_1^2}{(\omega + j\Gamma v_1)^2} + \frac{\omega_2^2}{(\omega + j\Gamma v_2)^2}. \tag{14}$$

It can be easily shown that, in a general case of a composite stream of electrons having many components of plasma frequencies ω_i and velocities v_i, the formula (14) takes the form:

$$1 = \sum_1^i \frac{\omega_i^2}{(\omega + j\Gamma v_i)^2}. \tag{14'}$$

Again, in the case of a continuous distribution of velocities between the limits of v_1 and v_2, (14) can be written as follows:

$$1 = \int_{v_1}^{v_2} \frac{\frac{d\omega_i^2}{dv} \cdot dv}{(\omega + j\Gamma v)^2}. \tag{14''}$$

In the case of a homogeneous electron beam, when $\omega_2=0$, (14) reduces to a familiar form:

$$(\omega + j\Gamma_1 v_1)^2 = \omega_1^2. \tag{15}$$

The solution of (15) immediately gives the familiar expression for the propagation constant:

$$\Gamma_1 = j\left(\frac{\omega}{v_1} \pm \frac{\omega_1}{v_1}\right). \tag{16}$$

The expression (16) shows that two electron waves propagate along the beam with the velocities somewhat higher and somewhat lower than the average electron velocity.

These waves produce interference effects in the beam which manifest themselves by the fact that, along the beam, periodic conversion of kinetic energy of electrons into space-charge-field energy takes place without amplification of the original disturbance.

In order to solve (14) in the case when the velocities are not equal, we can assume that

$$\left.\begin{aligned} v_1 &= v + \delta \\ v_2 &= v - \delta \end{aligned}\right\} \tag{17}$$

where v is the average velocity of electrons in the stream, and 2δ represents the difference in velocity of the two streams, i.e., $2\delta = v_1 - v_2$.

By putting

$$\Gamma = j\frac{\omega}{v} + j\gamma, \tag{18}$$

we obtain

$$\left.\begin{aligned} \omega + j\Gamma v_1 &= -\frac{\delta}{v}\omega - \gamma v \\ \omega + j\Gamma v_2 &= +\frac{\delta}{v}\omega - \gamma v \end{aligned}\right\}. \tag{19}$$

In terms of δ and γ, (14) can be written as follows:

$$1 = \frac{\omega_1^2}{\left(\frac{\delta}{v}\omega + \gamma v\right)^2} + \frac{\omega_2^2}{\left(\frac{\delta}{v}\omega - \gamma v\right)^2}. \tag{20}$$

This equation can be solved explicitly for a case when $\omega_1 = \omega_2$. In this case, the equation reduces to the following form:

$$1 = \frac{1}{\left(\frac{\delta\omega}{v\omega_1} + \frac{\gamma v}{\omega_1}\right)^2} + \frac{1}{\left(\frac{\delta\omega}{v\omega_1} - \frac{\gamma v}{\omega_1}\right)^2}. \tag{21}$$

It can be easily shown that (21) has the following solution:

$$\gamma\frac{v}{\omega_1} = \pm\sqrt{\left(\frac{\delta\omega}{v\omega_1}\right)^2 + 1 \pm \sqrt{4\left(\frac{\delta\omega}{v\omega_1}\right)^2 + 1}} \tag{22}$$

The factor $\delta\omega/v\omega_1$ is dimensionless and expresses the degree of inhomogeneity of the electron stream; the quantity $\gamma v/\omega_1$ is proportional to the looked-for component of the propagation constant and the average velocity v of the electrons, and is inversely proportional to the electron plasma frequency ω_1. The real and imaginary components of the normalized propagation constant $j(\gamma v/\omega_1)$ are shown plotted in Fig. 1 as a function of the inhomogeneity $\delta\omega/v\omega_1$ of the electron stream. The real component has finite value only over a limited

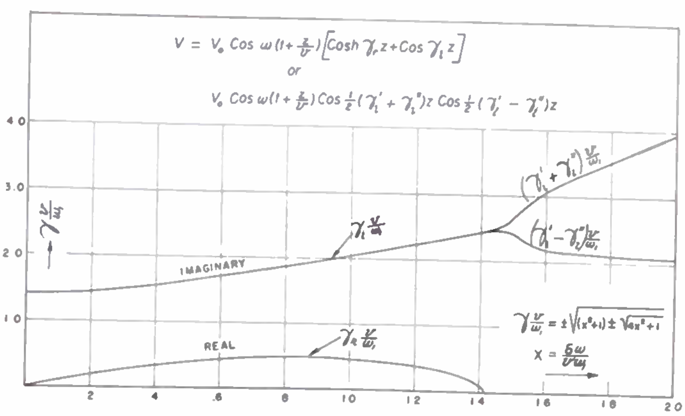

Fig. 1—Real and imaginary components of propagation constant versus inhomogeneity of the electron beam.

range of the inhomogeneity factor, from 0 to a value of $\sqrt{2}$. It is only over this range that effective amplification of the disturbances can take place. The maximum value of the real component is equal to one-half when the inhomogeneity is equal to the value of $\sqrt{3}/2$. The formula expressing the amplitude of the disturbance in this case is as follows:

$$V = V_0 \cos\omega\left(t + \frac{Z}{v}\right)\left[\cosh\gamma_r Z + \cos\gamma_i Z\right], \tag{23}$$

where γ_r is the real, and γ_i is the imaginary component of the propagation constant $j\gamma$.

The value of the imaginary component of the propagation constant increases with the degree of inhomogeneity. When the inhomogeneity exceeds the value of $\sqrt{2}$, four interfering waves are present in the beam without any amplification effects. In this case the magnitude of the disturbance along the beam can be expressed as follows:

$$V = V_0 \cos\omega\left(t + \frac{Z}{v}\right)\cos\tfrac{1}{2}(\gamma_i' + \gamma_i'')Z \cdot \cos\tfrac{1}{2}(\gamma_i' + \gamma_i'')Z. \tag{24}$$

Since the inhomogeneity of the beam is proportional to the signal frequency, the curves of Fig. 1 also represent variations of the real and imaginary components of the propagation constant with frequency. The ratio of the square of the amplitude of the disturbance at a distance Z along the beam to the magnitude of the disturbance at its origin at $Z=0$, represents a gain in energy of the disturbance. For large values of $\gamma_r Z$ the gain increases exponentially with the length Z.

Experimental Arrangements and Results

Fig. 2 represents an experimental arrangement which utilizes an inhomogeneous electron stream for amplification of high-frequency energy. The electron stream originates at two separate cathodes, No. 1 and No. 2, placed near the input end of the tube. Accelerator and focusing electrodes, placed in front of the cathodes, project the electron streams through a long, hollow metal cylinder, usually called the drift tube. At the opposite end of the tube the electrons are collected by a collector electrode. The divergence of the electron stream is prevented by the use of a magnetic field parallel to the velocity of electrons in the stream. This field is

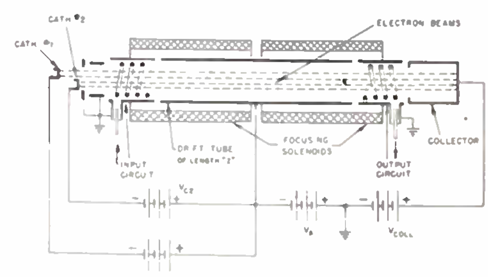

Fig. 2—Electron-wave tube (two-velocity type).

created by focusing solenoids as shown in the figure. The input signal is brought to the input end of the tube by means of a transmission line which terminates in a short helix surrounding the electron stream. In this manner the signal voltage produces initial disturbance in the beam. This disturbance is amplified in an exponential manner along the length of the drift tube. Greatly amplified fluctuations in the electron stream induce voltages in the output circuit. This circuit can be similar to the input circuit, as is shown in the figure.

In order to obtain maximum gain in the tube, the potentials between the cathode No. 1, cathode No. 2, and the drift tube are adjusted so that the inhomogeneity factor approaches the value of $\sqrt{3}/2$. Since this factor involves electron plasma frequency, the optimum adjustment depends upon the value of current density in the beam. Since the maximum value of the normalized real component of the propagation constant is equal to 0.5, the maximum gain of the electron-wave tube can be expressed as follows:

$$\text{gain (db)} = 20 \log_{10} e^{\gamma_r z} = 8.68\left(\frac{\gamma_r v}{\omega_1}\right)\frac{\omega_1}{v} Z$$

$$= 1330\, Z \sqrt{\frac{I}{V^{3/2}}}\ \text{cm}\ \frac{(\text{amp/cm}^2)^{1/2}}{\text{volts}^{3/4}}. \quad (25)$$

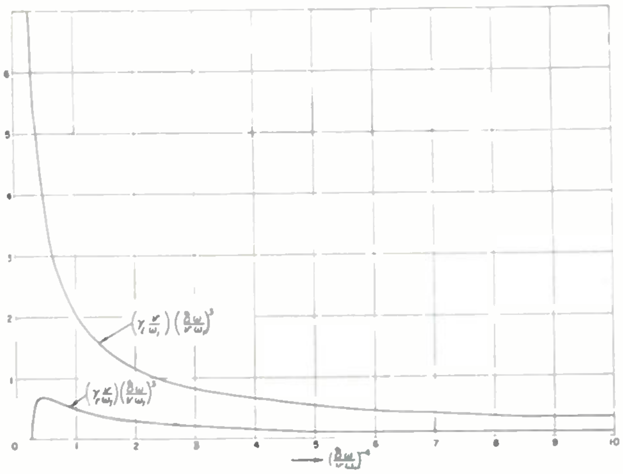

Fig. 3—Real and imaginary components of the propagation constant versus average electron energy.

Instead of adjustment of the difference of potential between the two cathodes, the drift-tube potential can be adjusted for optimum gain. For a constant current in the beam, the variation of the real and imaginary components of the propagation constant with the potential of the drift tube expressed in dimensionless units is shown in Fig. 3. The real component reaches its maximum value at the normalized drift-tube voltage of approximately 0.4. The variation of the output, or gain, of the tube with the drift-tube potential can be easily derived from the generalized curves of Fig. 1. Such a diagram is represented in Fig. 4, where a rapid increase in gain as the voltage is lowered can be readily seen. The effect of interference of the component waves is apparent in the fluctuations of the output as the drift voltage is varied.

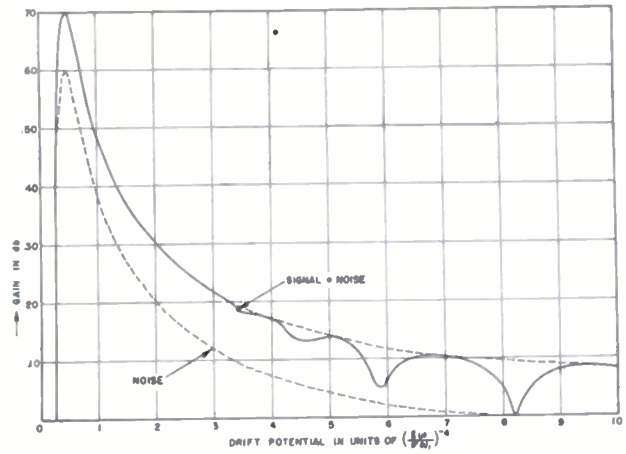

Fig. 4—Gain versus drift potential for a 70-db electron-wave tube for $S/N=10$ db.

Fig. 5 shows an electron-wave tube where only one electron beam is used. In this case the inhomogeneity of the beam is the result of the space-charge reduction of the potential along the axis of the beam, so that outer electrons travel at a velocity higher than the velocity of electrons along the axis. A more familiar form of the

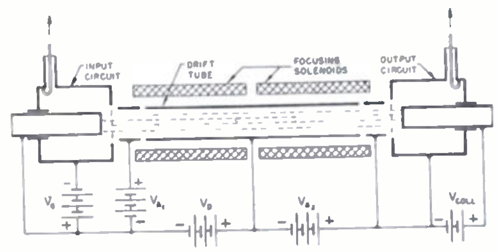

Fig. 5—Electron-wave tube (space-charge type).

resonant input and output circuits is shown in the figure. These circuits resemble those of a conventional tube, where the signal voltage, applied to the grid placed near the cathode, produces current modulation in the electron stream. The high-frequency energy is derived from the stream at the output circuit during the passage of electrons in the space between the screen-grid and the anode. The input control-grid–cathode structure is separated from the output screen-grid–anode structure by the intervening drift tube which may be of considerable length. The amplification of energy takes place entirely along the length of the drift tube. Since the signal is amplified in the drift tube before it arrives at the screen, the partition noise caused by partial interception of current by the screen grid is negligible. In this respect the signal-to-noise ratio for this tube is expected to approach that of a conventional triode. The variation of the real and imaginary components of the propagation constant with the drift-tube voltage for this "space-charge" type of electron-wave tube is shown in Fig. 6. In general, the shape of the curves in this figure very much resemble those of Fig. 3. The variation of the output with the drift-tube voltage for the space-charge type tube is shown in Fig. 7.

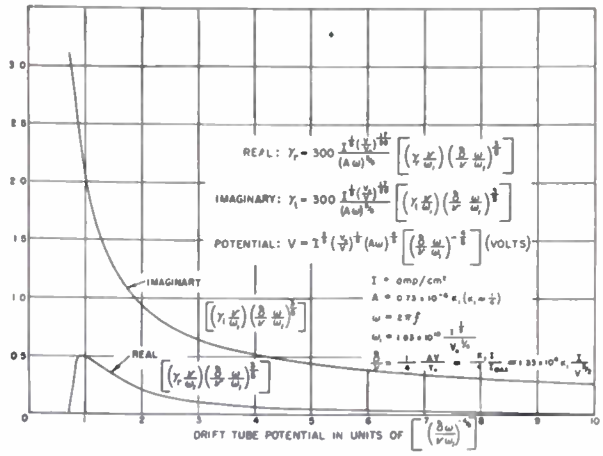

Fig. 6—Real and imaginary components of propagation constant versus drift-tube potential for the space-charge-type electron-wave tube.

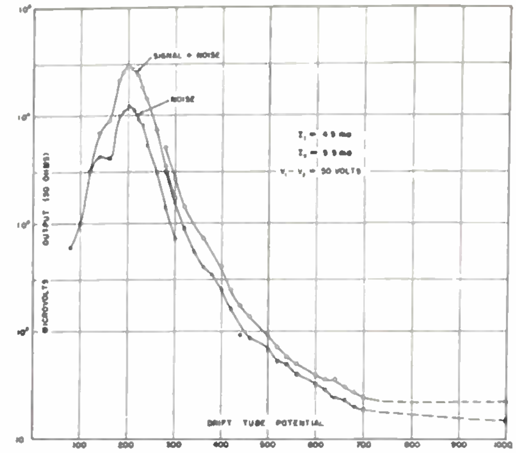

Fig. 8—Output versus drift-tube potential of the two-velocity-type electron-wave tube.

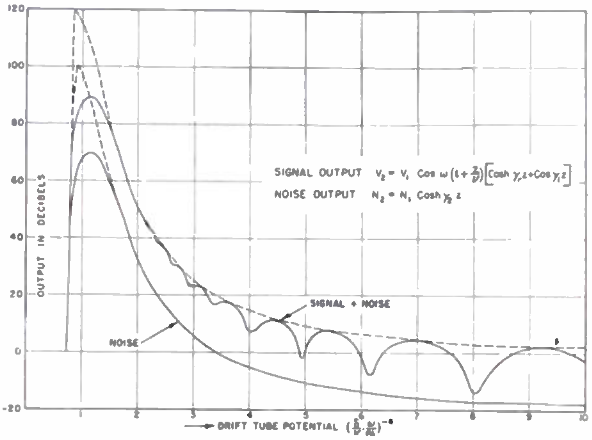

Fig. 7—Output versus drift-tube potential for a 90-db electron-wave tube (space-charge type) ($S/N=20$ db).

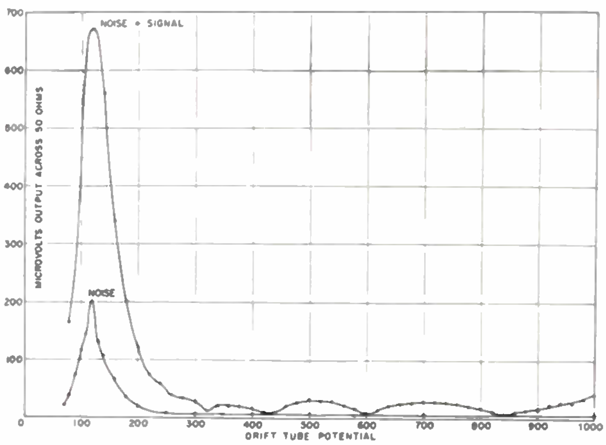

Fig. 9—Output versus drift-tube potential of the space-charge-type electron-wave tube (beam current = 5 ma).

Experimental curves of Figs. 8 and 9 show variation of output with the drift-tube potential. A close resemblance, in most respects, to the theoretical curves of Figs. 4 and 7 is apparent. These experimental data were obtained for a drift tube 20 cm in length.

The generalized curves of Fig. 1 make it possible to estimate the electronic bandwidth of the electron-wave tube. The variation of the fractional bandwidth with the gain of the tube is shown in Fig. 10. It is interesting to observe that, even at gains as high as 100 db, the electronic bandwidth is still greater than 30 per cent.

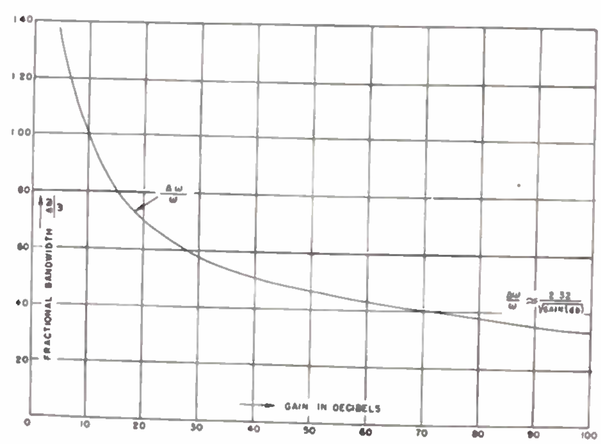

Fig. 10—Electronic bandwidth versus gain of the electron-wave tube.

Fig. 11 shows a cross section of a demountable tube of the two-velocity type. Two flat spiral cathodes are arranged along the axis of the tube in such a way that electrons emitted by the first cathode can penetrate between the turns of the spiral of the second cathode and enter into the input-circuit region. By the use of such semitransparent cathodes, a very thorough mixing of the two electron streams is achieved. After passing through the input circuit, consisting of a helix surrounded by a metal cylindrical shield, the electron streams enter the region of the drift tube, which can be maintained at a suitable potential independent of the potential of the helix. The output circuit is similar to the input circuit. Both input and output helices are maintained at a high potential, so that the velocity of electrons in passing through the helix is approximately equal to the phase velocity of the field waves on the helix. The electrons are finally collected on the collector electrode, which can be maintained at a somewhat

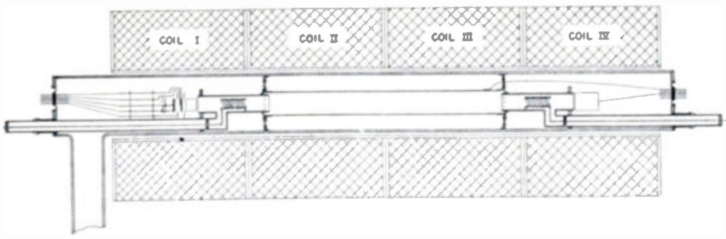

Fig. 11—Cross section of an experimental demountable electron-wave tube of the two-velocity type.

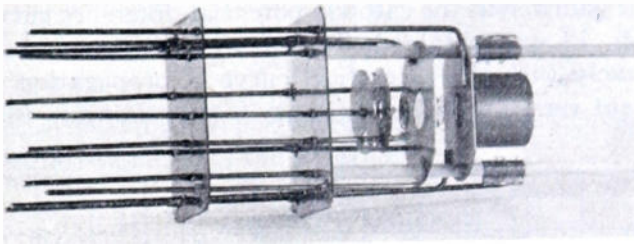

Fig. 13—Detail of the cathode assembly.

higher potential in order to suppress secondaries which, if permitted to re-enter the drift tube in the backwards direction, may produce undesirable regenerative effects. The input and output helices are tuned by means of external tuners attached to the coaxial transmission lines connected to the helices, one end of which is shorted to the surrounding shield. When energy is fed to the input helix, the interaction of the field waves on the helix produces modulations in the electron streams. This modulation is amplified as the electrons drift towards the output circuit. The density-modulated electron stream entering the output circuit induces voltages in the output helix which are then fed to the output circuit. Fig. 12 shows a photograph of a demountable tube constructed in accordance with the sketch of the previous figure. The input and output structures and the drift tube can be readily seen in the photograph. Fig. 13 shows detail of the construction of the spiral cathodes. Figs. 14 and 15 show the details of the input circuit and the cathode assembly, and the output circuit and the collector assembly, respectively.

Fig. 12—Photograph of an experimental demountable electron-wave tube of the two-velocity type.

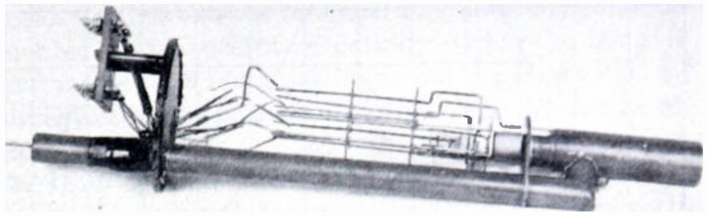

Fig. 14—Detail of the input circuit and cathode assembly.

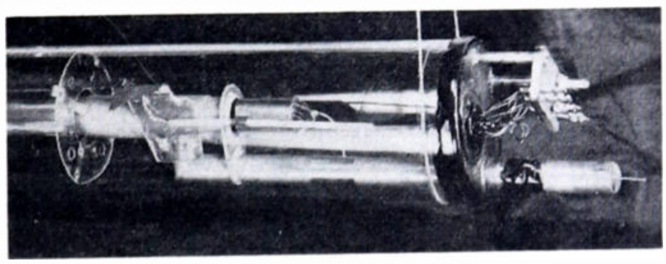

Fig. 15—Detail of the output circuit and collector assembly.

Fig. 16 shows the experimental arrangement which was used to measure the performance of the electron-wave tube. The demountable electron tube is mounted on the pumping table, and is surrounded by the focusing solenoids, at the extreme ends of which the input and output tuners can be seen. On the left are the power supplies, and on the right, the signal generator, the receiver, and the rf voltmeter. The input signal is fed from the signal generator to the input circuit. The output of the tube is fed from the output circuit through a series of attenuators to the input of the wide-range microwave receiver. The attenuators are used in order to prevent overload of the receiver because of the high-level output from the electron-wave tube. The rf voltmeter is used to indicate the output voltage.

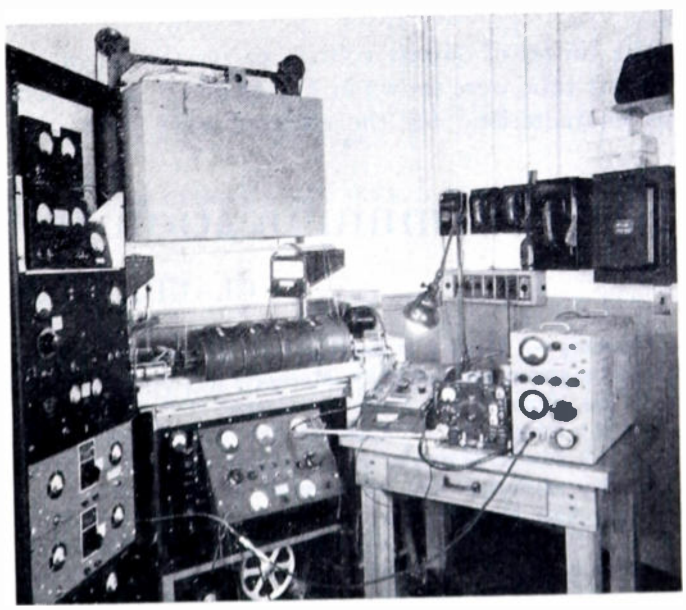

Fig. 16—Experimental arrangement used to measure the performance of electron-wave tubes.

The gain measurements were made by the direct-substitution method. For a given output voltmeter reading, the signal-generator voltage level is recorded for the case when the electron-wave-tube amplifier is in the circuit, and for the case when the signal is fed directly through the attenuator to the receiver. In this manner, the gain versus the cathode-potential-difference curve of Fig. 17 was obtained. This figure corresponds rather closely with the theoretical curve of propagation constant versus the inhomogeneity factor, shown in Fig. 1.

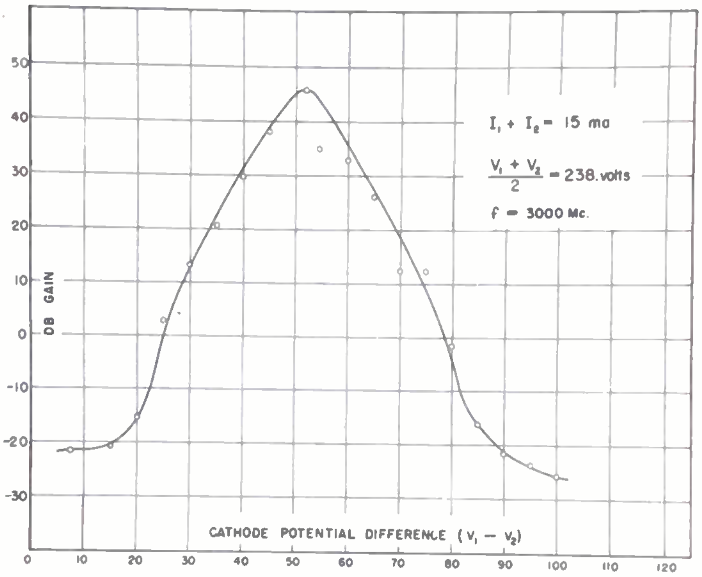

Fig. 17—Gain versus cathode-potential-difference characteristics of the two-velocity-type electron-wave tube.

At a frequency of 3000 Mc and a total current of 1.5 ma, a net gain of 46 db was obtained, even though no attempt was made to match either the input or output circuits. The lack of appropriate match is responsible for the fact that the gain curve assumes negative values when the electronic gain is not sufficient to overcome the losses due to mismatch. At the peak of the curve, it is estimated that the electronic gain is of the order of 80 db.

The curves of output voltage versus the potential of the drift tube were shown in Figs. 8 and 9. Fig. 9 shows this characteristic for the electron-wave tube of the space-charge type illustrated in Fig. 5. The shape of this curve corresponds rather closely with the shape of the theoretical curve given in Fig. 7. Fig. 8 shows the output voltage versus drift-potential characteristic for the two-velocity-type electron-wave tube. When the drift-tube voltage is high, the tube behaves like the two-cavity klystron amplifier. As the drift voltage is lowered the gain gradually increases, due to the space-charge interaction effect, and achieves a maximum which is approximately 60 db higher than the output achieved with klystron operation. With further reduction of the drift-tube potential the output drops rather rapidly, because the space-charge conditions become unfavorable; that is, the inhomogeneity factor becomes too large.

The electronic bandwidth was measured by measuring the gain of the tube over a frequency range from 2000 to 3000 Mc and retuning the input and output circuits for each frequency. It was observed that the gain of the tube was essentially constant over this frequency range, thus confirming the theoretical prediction of electronic bandwidth of over 30 per cent at the gain of 80 db.

The electron-wave tube, because of its remarkable property of achieving energy amplification without the use of any resonant or waveguiding structures in the amplifying region of the tube, promises to offer a satisfactory solution to the problem of generation and amplification of energy at millimeter wavelengths, and thus will aid in expediting the exploitation of that portion of the electromagnetic spectrum.

Acknowledgment

The author wishes to express his appreciation of the enthusiastic support of all his co-workers at the Naval Research Laboratory who helped to carry out this project from the stage of conception to the production and tests of experimental electron-wave tubes. The untiring efforts of two of the author's assistants, C. B. Smith and R. S. Ware, are particularly appreciated.

Communication in the Presence of Noise*

CLAUDE E. SHANNON†, MEMBER, IRE

Summary—A method is developed for representing any communication system geometrically. Messages and the corresponding signals are points in two "function spaces," and the modulation process is a mapping of one space into the other. Using this representation, a number of results in communication theory are deduced concerning expansion and compression of bandwidth and the threshold effect. Formulas are found for the maximum rate of transmission of binary digits over a system when the signal is perturbed by various types of noise. Some of the properties of "ideal" systems which transmit at this maximum rate are discussed. The equivalent number of binary digits per second for certain information sources is calculated.

* Decimal classification: 621.38. Original manuscript received by the Institute, July 23, 1940. Presented, 1948 IRE National Convention, New York, N. Y., March 24, 1948; and IRE New York Section, New York, N. Y., November 12, 1947.

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

I. Introduction

A GENERAL COMMUNICATIONS system is shown schematically in Fig. 1. It consists essentially of five elements.

1. *An information source.* The source selects one message from a set of possible messages to be transmitted to the receiving terminal. The message may be of various types; for example, a sequence of letters or numbers, as in telegraphy or teletype, or a continuous function of time $f(t)$, as in radio or telephony.

2. *The transmitter.* This operates on the message in some way and produces a signal suitable for transmission to the receiving point over the channel. In teleph-

ony, this operation consists of merely changing sound pressure into a proportional electrical current. In telegraphy, we have an encoding operation which produces a sequence of dots, dashes, and spaces corresponding to the letters of the message. To take a more complex example, in the case of multiplex PCM telephony the different speech functions must be sampled, compressed, quantized and encoded, and finally interleaved properly to construct the signal.

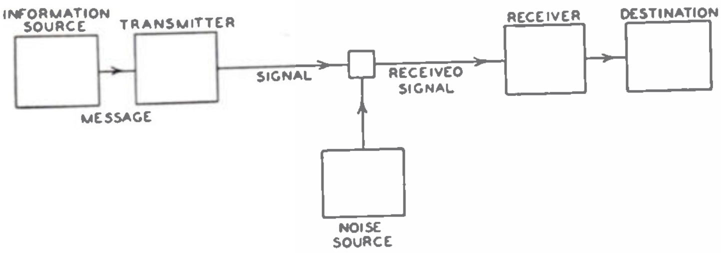

Fig. 1—General communications system.

3. *The channel.* This is merely the medium used to transmit the signal from the transmitting to the receiving point. It may be a pair of wires, a coaxial cable, a band of radio frequencies, etc. During transmission, or at the receiving terminal, the signal may be perturbed by noise or distortion. Noise and distortion may be differentiated on the basis that distortion is a fixed operation applied to the signal, while noise involves statistical and unpredictable perturbations. Distortion can, in principle, be corrected by applying the inverse operation, while a perturbation due to noise cannot always be removed, since the signal does not always undergo the same change during transmission.

4. *The receiver.* This operates on the received signal and attempts to reproduce, from it, the original message. Ordinarily it will perform approximately the mathematical inverse of the operations of the transmitter, although they may differ somewhat with best design in order to combat noise.

5. *The destination.* This is the person or thing for whom the message is intended.

Following Nyquist[1] and Hartley,[2] it is convenient to use a logarithmic measure of information. If a device has n possible positions it can, by definition, store $\log_b n$ units of information. The choice of the base b amounts to a choice of unit, since $\log_b n = \log_b c \log_c n$. We will use the base 2 and call the resulting units binary digits or bits. A group of m relays or flip-flop circuits has 2^m possible sets of positions, and can therefore store $\log_2 2^m = m$ bits.

If it is possible to distinguish reliably M different signal functions of duration T on a channel, we can say that the channel can transmit $\log_2 M$ bits in time T. The *rate* of transmission is then $\log_2 M/T$. More precisely, the *channel capacity* may be defined as

$$C = \lim_{T\to\infty} \frac{\log_2 M}{T}. \tag{1}$$

A precise meaning will be given later to the requirement of reliable resolution of the M signals.

II. The Sampling Theorem

Let us suppose that the channel has a certain bandwidth W in cps starting at zero frequency, and that we are allowed to use this channel for a certain period of time T. Without any further restrictions this would mean that we can use as signal functions any functions of time whose spectra lie entirely within the band W, and whose time functions lie within the interval T. Although it is not possible to fulfill both of these conditions exactly, it is possible to keep the spectrum within the band W, and to have the time function very small outside the interval T. Can we describe in a more useful way the functions which satisfy these conditions? One answer is the following:

Theorem 1: *If a function $f(t)$ contains no frequencies higher than W cps, it is completely determined by giving its ordinates at a series of points spaced $1/2W$ seconds apart.*

This is a fact which is common knowledge in the communication art. The intuitive justification is that, if $f(t)$ contains no frequencies higher than W, it cannot change to a substantially new value in a time less than one-half cycle of the highest frequency, that is, $1/2W$. A mathematical proof showing that this is not only approximately, but exactly, true can be given as follows. Let $F(\omega)$ be the spectrum of $f(t)$. Then

$$f(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty} F(\omega)e^{i\omega t}d\omega \tag{2}$$

$$= \frac{1}{2\pi}\int_{-2\pi W}^{+2\pi W} F(\omega)e^{i\omega t}d\omega, \tag{3}$$

since $F(\omega)$ is assumed zero outside the band W. If we let

$$t = \frac{n}{2W} \tag{4}$$

where n is any positive or negative integer, we obtain

$$f\left(\frac{n}{2W}\right) = \frac{1}{2\pi}\int_{-2\pi W}^{+2\pi W} F(\omega)e^{i\omega \frac{n}{2W}}\,d\omega. \tag{5}$$

On the left are the values of $f(t)$ at the sampling points. The integral on the right will be recognized as essentially the nth coefficient in a Fourier-series expansion of the function $F(\omega)$, taking the interval $-W$ to $+W$ as a fundamental period. This means that the values of the samples $f(n/2W)$ determine the Fourier coefficients in the series expansion of $F(\omega)$. Thus they determine $F(\omega)$, since $F(\omega)$ is zero for frequencies greater than W, and for

[1] H. Nyquist, "Certain factors affecting telegraph speed," *Bell Syst. Tech. Jour.*, vol. 3, p. 324; April, 1924.

[2] R. V. L. Hartley, "The transmission of information," *Bell Sys. Tech. Jour.*, vol. 3, p. 535–564; July, 1928.

lower frequencies $F(\omega)$ is determined if its Fourier coefficients are determined. But $F(\omega)$ determines the original function $f(t)$ completely, since a function is determined if its spectrum is known. Therefore the original samples determine the function $f(t)$ completely. There is one and only one function whose spectrum is limited to a band W, and which passes through given values at sampling points separated $1/2W$ seconds apart. The function can be simply reconstructed from the samples by using a pulse of the type

$$\frac{\sin 2\pi Wt}{2\pi Wt}. \tag{6}$$

This function is unity at $t=0$ and zero at $t=n/2W$, i.e., at all other sample points. Furthermore, its spectrum is constant in the band W and zero outside. At each sample point a pulse of this type is placed whose amplitude is adjusted to equal that of the sample. The sum of these pulses is the required function, since it satisfies the conditions on the spectrum and passes through the sampled values.

Mathematically, this process can be described as follows. Let x_n be the nth sample. Then the function $f(t)$ is represented by

$$f(t) = \sum_{n=-\infty}^{\infty} x_n \frac{\sin \pi(2Wt - n)}{\pi(2Wt - n)}. \tag{7}$$

A similar result is true if the band W does not start at zero frequency but at some higher value, and can be proved by a linear translation (corresponding physically to single-sideband modulation) of the zero-frequency case. In this case the elementary pulse is obtained from $\sin x/x$ by single-side-band modulation.

If the function is limited to the time interval T and the samples are spaced $1/2W$ seconds apart, there will be a total of $2TW$ samples in the interval. All samples outside will be substantially zero. To be more precise, we can define a function to be limited to the time interval T if, and only if, all the samples outside this interval are exactly zero. Then we can say that any function limited to the bandwidth W and the time interval T can be specified by giving $2TW$ numbers.

Theorem 1 has been given previously in other forms by mathematicians[3] but in spite of its evident importance seems not to have appeared explicitly in the literature of communication theory. Nyquist,[4,5] however, and more recently Gabor,[6] have pointed out that approximately $2TW$ numbers are sufficient, basing their arguments on a Fourier series expansion of the function over the time interval T. This gives TW sine and $(TW+1)$ cosine terms up to frequency W. The slight discrepancy is due to the fact that the functions obtained in this way will not be strictly limited to the band W but, because of the sudden starting and stopping of the sine and cosine components, contain some frequency content outside the band. Nyquist pointed out the fundamental importance of the time interval $1/2W$ seconds in connection with telegraphy, and we will call this the Nyquist interval corresponding to the band W.

The $2TW$ numbers used to specify the function need not be the equally spaced samples used above. For example, the samples can be unevenly spaced, although, if there is considerable bunching, the samples must be known very accurately to give a good reconstruction of the function. The reconstruction process is also more involved with unequal spacing. One can further show that the value of the function and its derivative at every other sample point are sufficient. The value and first and second derivatives at every third sample point give a still different set of parameters which uniquely determine the function. Generally speaking, any set of $2TW$ independent numbers associated with the function can be used to describe it.

III. Geometrical Representation of the Signals

A set of three numbers x_1, x_2, x_3, regardless of their source, can always be thought of as co-ordinates of a point in three-dimensional space. Similarly, the $2TW$ evenly spaced samples of a signal can be thought of as co-ordinates of a point in a space of $2TW$ dimensions. Each particular selection of these numbers corresponds to a particular point in this space. Thus there is exactly one point corresponding to each signal in the band W and with duration T.

The number of dimensions $2TW$ will be, in general, very high. A 5-Mc television signal lasting for an hour would be represented by a point in a space with $2\times 5 \times 10^6 \times 60^2 = 3.6\times 10^{10}$ dimensions. Needless to say, such a space cannot be visualized. It is possible, however, to study analytically the properties of n-dimensional space. To a considerable extent, these properties are a simple generalization of the properties of two- and three-dimensional space, and can often be arrived at by inductive reasoning from these cases. The advantage of this geometrical representation of the signals is that we can use the vocabulary and the results of geometry in the communication problem. Essentially, we have replaced a complex entity (say, a television signal) in a simple environment (the signal requires only a plane for its representation as $f(t)$) by a simple entity (a point) in a complex environment ($2TW$ dimensional space).

If we imagine the $2TW$ co-ordinate axes to be at right angles to each other, then distances in the space have a simple interpretation. The distance from the origin to a

[3] J. M. Whittaker, "Interpolatory Function Theory," Cambridge Tracts in Mathematics and Mathematical Physics, No. 33, Cambridge University Press, Chapt. IV; 1935.

[4] H. Nyquist, "Certain topics in telegraph transmission theory," *A.I.E.E. Transactions*, p. 617; April, 1928.

[5] W. R. Bennett, "Time division multiplex systems," *Bell Sys. Tech. Jour.*, vol. 20, p. 199; April, 1941, where a result similar to Theorem 1 is established, but on a steady-state basis.

[6] D. Gabor, "Theory of communication," *Jour. I.E.E.* (London), vol. 93; part 3, no. 26, p. 429; 1946.

point is analogous to the two- and three-dimensional cases

$$d = \sqrt{\sum_{n=1}^{2TW} x_n^2} \tag{8}$$

where x_n is the nth sample. Now, since

$$f(t) = \sum_{n=1}^{2TW} x_n \frac{\sin \pi(2Wt - n)}{\pi(2Wt - n)}, \tag{9}$$

we have

$$\int_{-\infty}^{\infty} f(t)^2 dt = \frac{1}{2W} \sum x_n^2, \tag{10}$$

using the fact that

$$\int_{-\infty}^{\infty} \frac{\sin \pi(2Wt - m)}{\pi(2Wt - m)} \frac{\sin \pi(2Wt - n)}{\pi(Wt - n)} dt = \begin{cases} 0 & m \neq n \\ \frac{1}{2W} & m = n. \end{cases} \tag{11}$$

Hence, the square of the distance to a point is $2W$ times the energy (more precisely, the energy into a unit resistance) of the corresponding signal

$$\begin{aligned} d^2 &= 2WE \\ &= 2WTP \end{aligned} \tag{12}$$

where P is the average power over the time T. Similarly, the distance between two points is $\sqrt{2WT}$ times the rms discrepancy between the two corresponding signals.

If we consider only signals whose average power is less than P, these will correspond to points within a sphere of radius

$$r = \sqrt{2WTP}. \tag{13}$$

If noise is added to the signal in transmission, it means that the point corresponding to the signal has been moved a certain distance in the space proportional to the rms value of the noise. Thus noise produces a small region of uncertainty about each point in the space. A fixed distortion in the channel corresponds to a warping of the space, so that each point is moved, but in a definite fixed way.

In ordinary three-dimensional space it is possible to set up many different co-ordinate systems. This is also possible in the signal space of $2TW$ dimensions that we are considering. A different co-ordinate system corresponds to a different way of describing the same signal function. The various ways of specifying a function given above are special cases of this. One other way of particular importance in communication is in terms of frequency components. The function $f(t)$ can be expanded as a sum of sines and cosines of frequencies $1/T$ apart, and the coefficients used as a different set of co-ordinates. It can be shown that these co-ordinates are all perpendicular to each other and are obtained by what is essentially a rotation of the original co-ordinate system.

Passing a signal through an ideal filter corresponds to projecting the corresponding point onto a certain region in the space. In fact, in the frequency-co-ordinate system those components lying in the pass band of the filter are retained and those outside are eliminated, so that the projection is on one of the co-ordinate lines, planes, or hyperplanes. Any filter performs a linear operation on the vectors of the space, producing a new vector linearly related to the old one.

IV. Geometrical Representation of Messages

We have associated a space of $2TW$ dimensions with the set of possible signals. In a similar way one can associate a space with the set of possible messages. Suppose we are considering a speech system and that the messages consist of all possible sounds which contain no frequencies over a certain limit W_1 and last for a time T_1.

Just as for the case of the signals, these messages can be represented in a one-to-one way in a space of $2T_1W_1$ dimensions. There are several points to be noted, however. In the first place, various different points may represent the same message, insofar as the final destination is concerned. For example, in the case of speech, the ear is insensitive to a certain amount of phase distortion. Messages differing only in the phases of their components (to a limited extent) sound the same. This may have the effect of reducing the number of essential dimensions in the message space. All the points which are equivalent for the destination can be grouped together and treated as one point. It may then require fewer numbers to specify one of these "equivalence classes" than to specify an arbitrary point. For example, in Fig. 2 we have a two-dimensional space, the set of points in a square. If all points on a circle are regarded as equivalent, it reduces to a one-dimensional space—a point can now be

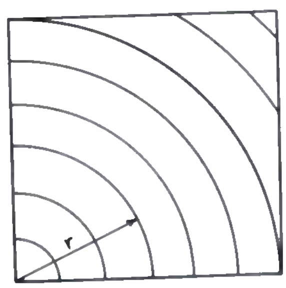

Fig. 2—Reduction of dimensionality through equivalence classes.

specified by one number, the radius of the circle. In the case of sounds, if the ear were completely insensitive to phase, then the number of dimensions would be reduced by one-half due to this cause alone. The sine and cosine components a_n and b_n for a given frequency would not need to be specified independently, but only $\sqrt{a_n^2+b_n^2}$; that is, the total amplitude for this frequency. The reduction in frequency discrimination of the ear as frequency increases indicates that a further reduction in dimensionality occurs. The vocoder makes use to a considerable extent of these equivalences among speech sounds, in the first place by eliminating, to a large degree, phase information, and in the second place by lumping groups of frequencies together, particularly at the higher frequencies.

In other types of communication there may not be any equivalence classes of this type. The final destination is sensitive to any change in the message within the full message space of $2T_1W_1$ dimensions. This appears to be the case in television transmission.

A second point to be noted is that the information source may put certain restrictions on the actual messages. The space of $2T_1W_1$ dimensions contains a point for *every* function of time $f(t)$ limited to the band W_1 and of duration T_1. The class of messages we wish to transmit may be only a small subset of these functions. For example, speech sounds must be produced by the human vocal system. If we are willing to forego the transmission of any other sounds, the effective dimensionality may be considerably decreased. A similar effect can occur through probability considerations. Certain messages may be possible, but so improbable relative to the others that we can, in a certain sense, neglect them. In a television image, for example, successive frames are likely to be very nearly identical. There is a fair probability of a particular picture element having the same light intensity in successive frames. If this is analyzed mathematically, it results in an effective reduction of dimensionality of the message space when T_1 is large.

We will not go further into these two effects at present, but let us suppose that, when they are taken into account, the resulting message space has a dimensionality D, which will, of course, be less than or equal to $2T_1W_1$. In many cases, even though the effects are present, their utilization involves too much complication in the way of equipment. The system is then designed on the basis that all functions are different and that there are no limitations on the information source. In this case, the message space is considered to have the full $2T_1W_1$ dimensions.

V. Geometrical Representation of the Transmitter and Receiver

We now consider the function of the transmitter from this geometrical standpoint. The input to the transmitter is a message; that is, one point in the message space. Its output is a signal—one point in the signal space. Whatever form of encoding or modulation is performed, the transmitter must establish some correspondence between the points in the two spaces. Every point in the message space must correspond to a point in the signal space, and no two messages can correspond to the same signal. If they did, there would be no way to determine at the receiver which of the two messages was intended. The geometrical name for such a correspondence is a mapping. The transmitter maps the message space into the signal space.

In a similar way, the receiver maps the signal space back into the message space. Here, however, it is possible to have more than one point mapped into the same point. This means that several different signals are demodulated or decoded into the same message. In AM, for example, the phase of the carrier is lost in demodulation. Different signals which differ only in the phase of the carrier are demodulated into the same message. In FM the shape of the signal wave above the limiting value of the limiter does not affect the recovered message. In PCM considerable distortion of the received pulses is possible, with no effect on the output of the receiver.

We have so far established a correspondence between a communication system and certain geometrical ideas. The correspondence is summarized in Table I.

TABLE I

Communication System	*Geometrical Entity*
The set of possible signals	A space of $2TW$ dimensions
A particular signal	A point in the space
Distortion in the channel	A warping of the space
Noise in the channel	A region of uncertainty about each point
The average power of the signal	$(2TW)^{-1}$ times the square of the distance from the origin to the point
The set of signals of power P	The set of points in a sphere of radius $\sqrt{2TW\,P}$
The set of possible messages	A space of $2T_1W_1$ dimensions
The set of actual messages distinguishable by the destination	A space of D dimensions obtained by regarding all equivalent messages as one point, and deleting messages which the source could not produce
A message	A point in this space
The transmitter	A mapping of the message space into the signal space
The receiver	A mapping of the signal space into the message space

VI. Mapping Considerations

It is possible to draw certain conclusions of a general nature regarding modulation methods from the geometrical picture alone. Mathematically, the simplest types of mappings are those in which the two spaces have the same number of dimensions. Single-sideband amplitude modulation is an example of this type and an especially simple one, since the co-ordinates in the signal space are proportional to the corresponding co-ordinates in the message space. In double-sideband transmission the signal space has twice the number of co-ordinates, but they occur in pairs with equal values. If there were only one dimension in the message space and two in the signal space, it would correspond to mapping a line onto a square so that the point x on the line is represented by (x, x) in the square. Thus no significant use is made of the extra dimensions. All the messages go into a subspace having only $2T_1W_1$ dimensions.

In frequency modulation the mapping is more involved. The signal space has a much larger dimensionality than the message space. The type of mapping can be suggested by Fig. 3, where a line is mapped into a three-dimensional space. The line starts at unit distance from

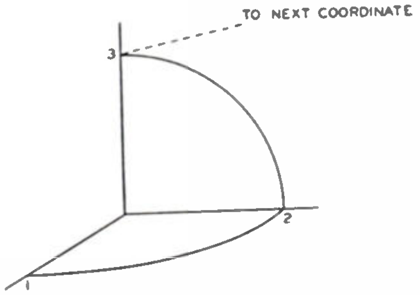

Fig. 3—Mapping similar to frequency modulation.

the origin on the first co-ordinate axis, stays at this distance from the origin on a circle to the next co-ordinate axis, and then goes to the third. It can be seen that the line is lengthened in this mapping in proportion to the total number of co-ordinates. It is not, however, nearly as long as it could be if it wound back and forth through the space, filling up the internal volume of the sphere it traverses.

This expansion of the line is related to the improved signal-to-noise ratio obtainable with increased bandwidth. Since the noise produces a small region of uncertainty about each point, the effect of this on the recovered message will be less if the map is in a large scale. To obtain as large a scale as possible requires that the line wander back and forth through the higher-dimensional region as indicated in Fig. 4, where we have mapped a line into a square. It will be noticed that when this is done the effect of noise is small relative to the length of the line, provided the noise is less than a certain critical value. At this value it becomes uncertain at the receiver as to which portion of the line contains the message. This holds generally, and it shows that *any system which attempts to use the capacities of a wider band to the full extent possible will suffer from a threshold effect when there is noise.* If the noise is small, very little distortion will occur, but at some critical noise amplitude the message will become very badly distorted. This effect is well known in PCM.

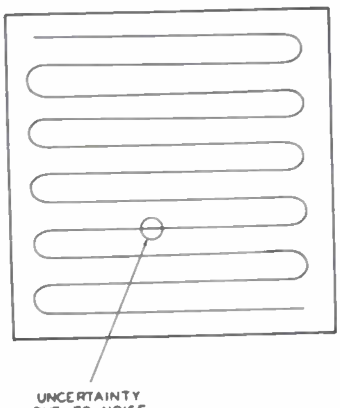

Fig. 4—Efficient mapping of a line into a square.

Suppose, on the other hand, we wish to reduce dimensionality, i.e., to compress bandwidth or time or both. That is, we wish to send messages of band W_1 and duration T_1 over a channel with $TW < T_1W_1$. It has already been indicated that the effective dimensionality D of the message space may be less than $2T_1W_1$ due to the properties of the source and of the destination. Hence we certainly need no more than D dimension in the signal space for a good mapping. To make this saving it is necessary, of course, to isolate the effective co-ordinates in the message space, and to send these only. The reduced bandwidth transmission of speech by the vocoder is a case of this kind.

The question arises, however, as to whether further reduction is possible. In our geometrical analogy, is it possible to map a space of high dimensionality onto one of lower dimensionality? The answer is that it is pos i-ble, with certain reservations. For example, the points of a square can be described by their two co-ordinates which could be written in decimal notation

$$x = .a_1a_2a_3 \cdots \qquad\qquad (14)$$
$$y = .b_1b_2b_3 \cdots .$$

From these two numbers we can construct one number by taking digits alternately from x and y:

$$z = .a_1b_1a_2b_2a_3b_3 \cdots . \qquad\qquad (15)$$

A knowledge of x and y determines z, and z determines both x and y. Thus there is a one-to-one correspondence between the points of a square and the points of a line.

This type of mapping, due to the mathematician Cantor, can easily be extended as far as we wish in the direction of reducing dimensionality. A space of n dimensions can be mapped in a one-to-one way into a space of one dimension. Physically, this means that the frequency-time product can be reduced as far as we wish when there is no noise, with exact recovery of the original messages.

In a less exact sense, a mapping of the type shown in Fig. 4 maps a square into a line, provided we are not too particular about recovering exactly the starting point, but are satisfied with a near-by one. The sensitivity we noticed before when increasing dimensionality now takes a different form. In such a mapping, to reduce TW, there will be a certain threshold effect when we perturb the message. As we change the message a small amount, the corresponding signal will change a small amount, until some critical value is reached. At this point the signal will undergo a considerable change. In topology it is shown[7] that it is not possible to map a region of higher dimension into a region of lower dimension *continuously*. It is the necessary discontinuity which produces the threshold effects we have been describing for communication systems.

[7] W. Hurewitz and H. Wallman, "Dimension Theory," Princeton University Press, Princeton, N. J.; 1941.

This discussion is relevant to the well-known "Hartley Law," which states that ". . . an upper limit to the amount of information which may be transmitted is set by the sum for the various available lines of the product of the line-frequency range of each by the time during which it is available for use."[2] There is a sense in which this statement is true, and another sense in which it is false. It is not possible to map the message space into the signal space in a one-to-one, continuous manner (this is known mathematically as a *topological* mapping) unless the two spaces have the same dimensionality; i.e., unless $D=2TW$. Hence, if we limit the transmitter and receiver to continuous one-to-one operations, there is a lower bound to the product TW in the channel. This lower bound is determined, not by the product W_1T_1 of message bandwidth and time, but by the number of *essential* dimension D, as indicated in Section IV. There is, however, no good reason for limiting the transmitter and receiver to topological mappings. In fact, PCM and similar modulation systems are highly discontinuous and come very close to the type of mapping given by (14) and (15). It is desirable, then, to find limits for what can be done with no restrictions on the type of transmitter and receiver operations. These limits, which will be derived in the following sections, depend on the amount and nature of the noise in the channel, and on the transmitter power, as well as on the bandwidth-time product.

It is evident that any system, either to compress TW, or to expand it and make full use of the additional volume, must be highly nonlinear in character and fairly complex because of the peculiar nature of the mappings involved.

VII. The Capacity of a Channel in the Presence of White Thermal Noise

It is not difficult to set up certain quantitative relations that must hold when we change the product TW. Let us assume, for the present, that the noise in the system is a white thermal-noise band limited to the band W, and that it is added to the transmitted signal to produce the received signal. A white thermal noise has the property that each sample is perturbed independently of all the others, and the distribution of each amplitude is Gaussian with standard deviation $\sigma=\sqrt{N}$ where N is the average noise power. How many different signals can be distinguished at the receiving point in spite of the perturbations due to noise? A crude estimate can be obtained as follows. If the signal has a power P, then the perturbed signal will have a power $P+N$. The number of amplitudes that can be reasonably well distinguished is

$$K\sqrt{\frac{P+N}{N}} \tag{16}$$

where K is a small constant in the neighborhood of unity depending on how the phrase "reasonably well" is interpreted. If we require very good separation, K will be small, while toleration of occasional errors allows K to be larger. Since in time T there are $2TW$ independent amplitudes, the total number of reasonably distinct signals is

$$M=\left[K\sqrt{\frac{P+N}{N}}\right]^{2TW}. \tag{17}$$

The number of bits that can be sent in this time is $\log_2 M$, and the rate of transmission is

$$\frac{\log_2 M}{T}=W\log_2 K^2\frac{P+N}{N} \text{ (bits per second).} \tag{18}$$

The difficulty with this argument, apart from its general approximate character, lies in the tacit assumption that for two signals to be distinguishable they must differ at some sampling point by more than the expected noise. The argument presupposes that PCM, or something very similar to PCM, is the best method of encoding binary digits into signals. Actually, two signals can be reliably distinguished if they differ by only a small amount, provided this difference is sustained over a long period of time. Each sample of the received signal then gives a small amount of statistical information concerning the transmitted signal; in combination, these statistical indications result in near certainty. This possibility allows an improvement of about 8 db in power over (18) with a reasonable definition of reliable resolution of signals, as will appear later. We will now make use of the geometrical representation to determine the exact capacity of a noisy channel.

Theorem 2: *Let P be the average transmitter power, and suppose the noise is white thermal noise of power N in the band W. By sufficiently complicated encoding systems it is possible to transmit binary digits at a rate*

$$C=W\log_2\frac{P+N}{N} \tag{19}$$

with as small a frequency of errors as desired. It is not possible by any encoding method to send at a higher rate and have an arbitrarily low frequency of errors.

This shows that the rate $W\log(P+N)/N$ measures in a sharply defined way the capacity of the channel for transmitting information. It is a rather surprising result, since one would expect that reducing the frequency of errors would require reducing the rate of transmission, and that the rate must approach zero as the error frequency does. Actually, we can send at the rate C but reduce errors by using more involved encoding and longer delays at the transmitter and receiver. The transmitter will take long sequences of binary digits and represent this entire sequence by a particular signal function of long duration. The delay is required because the transmitter must wait for the full sequence before the signal is determined. Similarly, the receiver must wait for the full signal function before decoding into binary digits.

We now prove Theorem 2. In the geometrical representation each signal point is surrounded by a small region of uncertainty due to noise. With white thermal noise, the perturbations of the different samples (or co-ordinates) are all Gaussian and independent. Thus the probability of a perturbation having co-ordinates $x_1, x_2, \cdots, x_n$ (these are the differences between the original and received signal co-ordinates) is the product of the individual probabilities for the different co-ordinates:

$$\prod_{n=1}^{2TW} \frac{1}{\sqrt{2\pi 2TWN}} \exp - \frac{x_n^2}{2TWN} = \frac{1}{(2\pi 2TWN)^{TW}} \exp \frac{-1}{2TW} \sum_1^{2TW} x_n^2.$$

Since this depends only on

$$\sum_1^{2TW} x_n^2,$$

the probability of a given perturbation depends only on the *distance* from the original signal and not on the direction. In other words, the region of uncertainty is spherical in nature. Although the limits of this region are not sharply defined for a small number of dimensions ($2TW$), the limits become more and more definite as the dimensionality increases. This is because the square of the distance a signal is perturbed is equal to $2TW$ times the average noise power during the time T. As T increases, this average noise power must approach N. Thus, for large T, the perturbation will almost certainly be to some point near the surface of a sphere of radius $\sqrt{2TWN}$ centered at the original signal point. More precisely, by taking T sufficiently large we can insure (with probability as near to 1 as we wish) that the perturbation will lie within a sphere of radius $\sqrt{2TW(N+\epsilon)}$ where ϵ is arbitrarily small. The noise regions can therefore be thought of roughly as sharply defined billiard balls, when $2TW$ is very large. The received signals have an average power $P+N$, and in the same sense must almost all lie on the surface of a sphere of radius $\sqrt{2TW(P+N)}$. How many different transmitted signals can be found which will be distinguishable? Certainly not more than the volume of the sphere of radius $\sqrt{2TW(P+N)}$ divided by the volume of a sphere of radius $\sqrt{2TWN}$, since overlap of the noise spheres results in confusion as to the message at the receiving point. The volume of an n-dimensional sphere[8] of radius r is

$$V = \frac{\pi^{n/2}}{\Gamma\left(\frac{n}{2}+1\right)} r^n . \tag{20}$$

Hence, an upper limit for the number M of distinguishable signals is

$$M \leqq \left(\sqrt{\frac{P+N}{N}}\right)^{2TW} . \tag{21}$$

Consequently, the channel capacity is bounded by:

$$C = \frac{\log_2 M}{T} \leqq W \log_2 \frac{P+N}{N} . \tag{22}$$

[8] D. M. Y. Sommerville, "An Introduction to the Geometry of N Dimensions," E. P. Dutton, Inc., New York, N. Y., 1929; p. 135.

This proves the last statement in the theorem.

To prove the first part of the theorem, we must show that there exists a system of encoding which transmits $W \log_2 (P+N)/N$ binary digits per second with a frequency of errors less than ϵ when ϵ is arbitrarily small. The system to be considered operates as follows. A long sequence of, say, m binary digits is taken in at the transmitter. There are 2^m such sequences, and each corresponds to a particular signal function of duration T. Thus there are $M=2^m$ different signal functions. When the sequence of m is completed, the transmitter starts sending the corresponding signal. At the receiver a perturbed signal is received. The receiver compares this signal with each of the M possible transmitted signals and selects the one which is nearest the perturbed signal (in the sense of rms error) as the one actually sent. The receiver then constructs, as its output, the corresponding sequence of binary digits. There will be, therefore, an over-all delay of $2T$ seconds.

To insure a frequency of errors less than ϵ, the M signal functions must be reasonably well separated from each other. In fact, we must choose them in such a way that, when a perturbed signal is received, the nearest signal point (in the geometrical representation) is, with probability greater than $1-\epsilon$, the actual original signal.

It turns out, rather surprisingly, that it is possible to choose our M signal functions at random from the points inside the sphere of radius $\sqrt{2TWP}$, and achieve the most that is possible. Physically, this corresponds very nearly to using M different samples of band-limited white noise with power P as signal functions.

A particular selection of M points in the sphere corresponds to a particular encoding system. The general scheme of the proof is to consider all such selections, and to show that the frequency of errors averaged over all the particular selections is less than ϵ. This will show that there are particular selections in the set with frequency of errors less than ϵ. Of course, there will be other particular selections with a high frequency of errors.

The geometry is shown in Fig. 5. This is a plane cross section through the high-dimensional sphere defined by a typical transmitted signal B, received signal A, and the origin 0. The transmitted signal will lie very

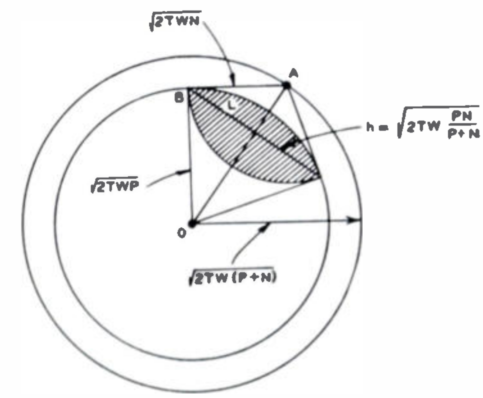

Fig. 5—The geometry involved in Theorem 2.

close to the surface of the sphere of radius $\sqrt{2TWP}$, since in a high-dimensional sphere nearly all the volume is very close to the surface. The received signal similarly will lie on the surface of the sphere of radius $\sqrt{2TW(P+N)}$. The high-dimensional lens-shaped region L is the region of possible signals that might have caused A, since the distance between the transmitted and received signal is almost certainly very close to $\sqrt{2TWN}$. L is of smaller volume than a sphere of radius h. We can determine h by equating the area of the triangle OAB, calculated two different ways:

$$\tfrac{1}{2}h\sqrt{2TW(P+N)} = \tfrac{1}{2}\sqrt{2TWP}\sqrt{2TWN}$$

$$h = \sqrt{2TW\frac{PN}{P+N}}\,.$$

The probability of any particular signal point (other than the actual cause of A) lying in L is, therefore, less than the ratio of the volumes of spheres of radii $\sqrt{2TW\,PN/P+N}$ and $\sqrt{2TWP}$, since in our ensemble of coding systems we chose the signal points at random from the points in the sphere of radius $\sqrt{2TWP}$. This ratio is

$$\left(\frac{\sqrt{2TW\dfrac{PN}{P+N}}}{\sqrt{2TWP}}\right)^{2TW} = \left(\frac{N}{P+N}\right)^{TW}. \tag{23}$$

We have M signal points. Hence the probability p that all except the actual cause of A are *outside* L is greater than

$$\left[1-\left(\frac{N}{P+N}\right)^{TW}\right]^{M-1}. \tag{24}$$

When these points are outside L, the signal is interpreted correctly. Therefore, if we make P greater than $1-\epsilon$, the frequency of errors will be less than ϵ. This will be true if

$$\left[1-\left(\frac{N}{P+N}\right)^{TW}\right]^{(M-1)} > 1-\epsilon. \tag{25}$$

Now $(1-x)^n$ is always greater than $1-nx$ when n is positive. Consequently, (25) will be true if

$$1-(M-1)\left(\frac{N}{P+N}\right)^{TW} > 1-\epsilon \tag{26}$$

or if

$$(M-1) < \epsilon\left(\frac{P+N}{N}\right)^{TW} \tag{27}$$

or

$$\frac{\log(M-1)}{T} < W\log\frac{P+N}{N} + \frac{\log\epsilon}{T}\,. \tag{28}$$

For any fixed ϵ, we can satisfy this by taking T sufficiently large, and also have $\log(M-1)/T$ or $\log M/T$ as close as desired to $W \log P+N/N$. This shows that, with a random selection of points for signals, we can obtain an arbitrarily small frequency of errors and transmit at a rate arbitrarily close to the rate C. We can also send *at* the rate C with arbitrarily small ϵ, since the extra binary digits need not be sent at all, but can be filled in at random at the receiver. This only adds another arbitrarily small quantity to ϵ. This completes the proof.

VIII. Discussion

We will call a system that transmits without errors at the rate C an ideal system. Such a system cannot be achieved with any finite encoding process but can be approximated as closely as desired. As we approximate more closely to the ideal, the following effects occur: (1) The rate of transmission of binary digits approaches $C = W\log_2(1+P/N)$. (2) The frequency of errors approaches zero. (3) The transmitted signal approaches a white noise in statistical properties. This is true, roughly speaking, because the various signal functions used must be distributed at random in the sphere of radius $\sqrt{2TWP}$. (4) The threshold effect becomes very sharp. If the noise is increased over the value for which the system was designed, the frequency of errors increases very rapidly. (5) The required delays at transmitter and receiver increase indefinitely. Of course, in a wide-band system a millisecond may be substantially an infinite delay.

In Fig. 6 the function $C/W = \log(1+P/N)$ is plotted with P/N in db horizontal and C/W the number of bits per cycle of band vertical. The circles represent PCM systems of the binary, ternary, etc., types, using positive and negative pulses and adjusted to give one error in about 10^5 binary digits. The dots are for a PPM system with two, three, etc., discrete positions for the pulse.[9]

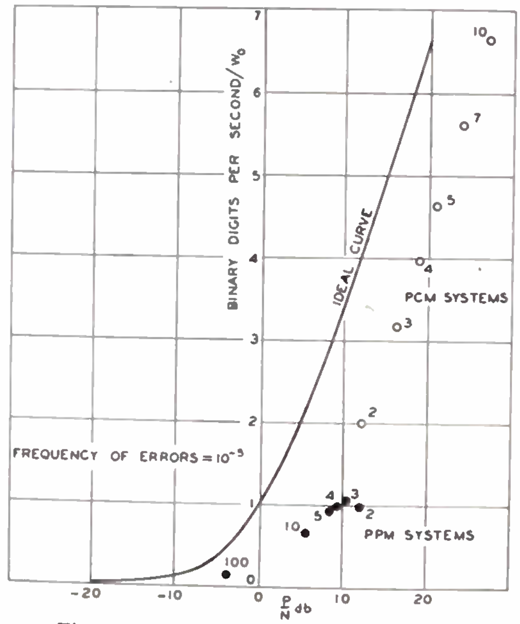

Fig. 6—Comparison of PCM and PPM with ideal performance.

[9] The PCM points are calculated from formulas given in "The philosophy of PCM," by B. M. Oliver, J. R. Pierce, and C. E. Shannon, Proc. I.R.E., vol. 36, pp. 1324–1332; November, 1948. The PPM points are from unpublished calculations of B. McMillan, who points out that, for very small P/N, the points approach to within 3 db of the ideal curve.

The difference between the series of points and the ideal curve corresponds to the gain that could be obtained by more involved coding systems. It amounts to about 8 db in power over most of the practical range. The series of points and circles is about the best that can be done without delay. Whether it is worth while to use more complex types of modulation to obtain some of this possible saving is, of course, a question of relative costs and valuations.

The quantity $TW \log(1+P/N)$ is, for large T, the number of bits that can be transmitted in time T. It can be regarded as an exchange relation between the different parameters. The individual quantities T, W, P, and N can be altered at will without changing the amount of information we can transmit, provided $TW \log(1+P/N)$ is held constant. If TW is reduced, P/N must be increased, etc.

Ordinarily, as we increase W, the noise power N in the band will increase proportionally; $N=N_0W$ where N_0 is the noise power per cycle. In this case, we have

$$C = W \log\left(1+\frac{P}{N_0W}\right). \tag{29}$$

If we let $W_0 = P/N_0$, i.e., W_0 is the band for which the noise power is equal to the signal power, this can be written

$$\frac{C}{W_0} = \frac{W}{W_0}\log\left(1+\frac{W_0}{W}\right). \tag{30}$$

In Fig. 7, C/W_0 is plotted as a function of W/W_0. As we increase the band, the capacity increases rapidly until the total noise power accepted is about equal to the signal power; after this, the increase is slow, and it approaches an asymptotic value $\log_2 e$ times the capacity for $W=W_0$.

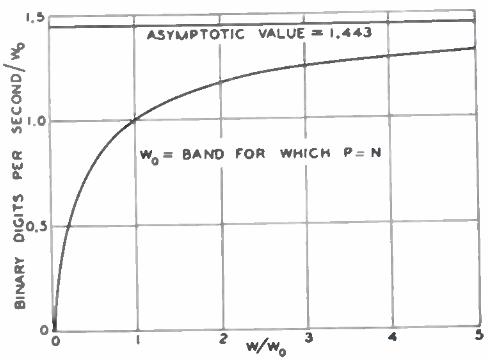

Fig. 7—Channel capacity as a function of bandwidth.

IX. Arbitrary Gaussian Noise

If a white thermal noise is passed through a filter whose transfer function is $Y(f)$, the resulting noise has a power spectrum $N(f) = K|Y(f)|^2$ and is known as Gaussian noise. We can calculate the capacity of a channel perturbed by any Gaussian noise from the white-noise result. Suppose our total transmitter power is P and it is distributed among the various frequencies according to $P(f)$. Then

$$\int_0^W P(f)df = P. \tag{31}$$

We can divide the band into a large number of small bands, with $N(f)$ approximately constant in each. The total capacity for a given distribution $P(f)$ will then be given by

$$C_1 = \int_0^W \log\left(1+\frac{P(f)}{N(f)}\right)df, \tag{32}$$

since, for each elementary band, the white-noise result applies. The maximum rate of transmission will be found by maximizing C_1 subject to condition (31). This requires that we maximize

$$\int_0^W \left[\log\left(1+\frac{P(f)}{N(f)}\right) + \lambda P(f)\right]df. \tag{33}$$

The condition for this is, by the calculus of variations, or merely from the convex nature of the curve $\log(1+x)$,

$$\frac{1}{N(f)+P(f)} + \lambda = 0, \tag{34}$$

or $N(f)+P(f)$ must be constant. The constant is adjusted to make the total signal power equal to P. For frequencies where the noise power is low, the signal power should be high, and vice versa, as we would expect.

The situation is shown graphically in Fig. 8. The curve is the assumed noise spectrum, and the three lines correspond to different choices of P. If P is small, we cannot make $P(f)+N(f)$ constant, since this would require negative power at some frequencies. It is easily shown, however, that in this case the best $P(f)$ is obtained by making $P(f)+N(f)$ constant whenever possible, and making $P(f)$ zero at other frequencies. With low values of P, some of the frequencies will not be used at all.

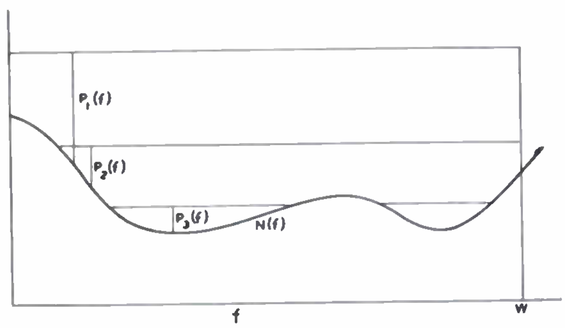

Fig. 8—Best distribution of transmitter power.

If we now vary the noise spectrum $N(f)$, keeping the total noise power constant and always adjusting the signal spectrum $P(f)$ to give the maximum transmission, we can determine the worst spectrum for the noise. This turns out to be the white-noise case. Although this only shows it to be worst among the Gaussian noises, it will be shown later to be the worst among all possible noises with the given power N in the band.

X. The Channel Capacity with an Arbitrary Type of Noise

Of course, there are many kinds of noise which are not Gaussian; for example, impulse noise, or white noise that has passed through a nonlinear device. If the signal is perturbed by one of these types of noise, there will still be a definite channel capacity C, the maximum rate of transmission of binary digits. We will merely outline the general theory here.[10]

Let $x_1, x_2, \cdots, x_n$ be the amplitudes of the noise at successive sample points, and let

$$p(x_1, x_2, \cdots, x_n)dx_1 \cdots dx_n \tag{35}$$

be the probability that these amplitudes lie between x_1 and x_1+dx_1, x_2 and x_2+dx_2, etc. Then the function p describes the statistical structure of the noise, insofar as n successive samples are concerned. The *entropy*, H, of the noise is defined as follows. Let

$$H_n = -\frac{1}{n}\int \cdots \int p(x_1, \cdots, x_n) \cdot \log_e p(x_1, \cdots, x_n)dx_1, \cdots, dx_n. \tag{36}$$

Then

$$H = \lim_{n\to\infty} H_n. \tag{37}$$

This limit exists in all cases of practical interest, and can be determined in many of them. H is a measure of the randomness of the noise. In the case of white Gaussian noise of power N, the entropy is

$$H = \log_e \sqrt{2\pi e N}. \tag{38}$$

It is convenient to measure the randomness of an arbitrary type of noise not directly by its entropy, but by comparison with white Gaussian noise. We can calculate the power in a white noise having the same entropy as the given noise. This power, namely,

$$\overline{N} = \frac{1}{2\pi e} \exp 2H \tag{39}$$

where H is the entropy of the given noise, will be called the *entropy power* of the noise.

A noise of entropy power $\overline{N}$ acts very much like a white noise of power $\overline{N}$, insofar as perturbing the message is concerned. It can be shown that the region of uncertainty about each signal point will have the same volume as the region associated with the white noise. Of course, it will no longer be a spherical region. In proving Theorem 1 this volume of uncertainty was the chief property of the noise used. Essentially the same argument may be applied for any kind of noise with minor modifications. The result is summarized in the following:

Theorem 3: *Let a noise limited to the band W have power N and entropy power N_1. The capacity C is then bounded by*

$$W \log_2 \frac{P+N_1}{N_1} \leq C \leq W \log_2 \frac{P+N}{N_1} \tag{40}$$

where P is the average signal power and W the bandwidth.

If the noise is a white Gaussian noise, $N_1 = N$, and the two limits are equal. The result then reduces to the theorem in Section VII.

For any noise, $N_1 < N$. This is why white Gaussian noise is the worst among all possible noises. If the noise is Gaussian with spectrum $N(f)$, then

$$N_1 = W \exp \frac{1}{W}\int_0^W \log N(f)df. \tag{41}$$

The upper limit in Theorem 3 is then reached when we are above the highest noise power in Fig. 8. This is easily verified by substitution.

In the cases of most interest, P/N is fairly large. The two limits are then nearly the same, and we can use $W \log (P+N)/N_1$ as the capacity. The upper limit is the best choice, since it can be shown that as P/N increases, C approaches the upper limit.

XI. Discrete Sources of Information

Up to now we have been chiefly concerned with the channel. The capacity C measures the maximum rate at which a random series of binary digits can be transmitted when they are encoded in the best possible way. In general, the information to be transmitted will not be in this form. It may, for example, be a sequence of letters as in telegraphy, a speech wave, or a television signal. Can we find an equivalent number of bits per second for information sources of this type? Consider first the discrete case; i.e., the message consists of a sequence of discrete symbols. In general, there may be correlation of various sorts between the different symbols. If the message is English text, the letter E is the most frequent, T is often followed by H, etc. These correlations allow a certain compression of the text by proper encoding. We may define the entropy of a discrete source in a way analogous to that for a noise; namely, let

$$H_n = -\frac{1}{n}\sum_{i,j,\cdots,s} p(i,j,\cdots,s)\log_2 p(i,j,\cdots,s) \tag{42}$$

where $p(i, j, \cdots, s)$ is the probability of the sequence of symbols $i, j, \cdots, s$, and the sum is over all sequences of n symbols. Then the entropy is

$$H = \lim_{n\to\infty} H_n. \tag{43}$$

[10] C. E. Shannon, "A mathematical theory of communication," *Bell Sys. Tech. Jour.*, vol. 27, pp. 379–424 and 623–657; July and October, 1948.

It turns out that H is the number of bits produced by the source for each symbol of message. In fact, the following result is proved in the Appendix.

THEOREM 4. *It is possible to encode all sequences of n message symbols into sequences of binary digits in such a way that the average number of binary digits per message symbol is approximately H, the approximation approaching equality as n increases.*

It follows that, if we have a channel of capacity C and a discrete source of entropy H, it is possible to encode the messages via binary digits into signals and transmit at the rate C/H of the original message symbols per second.

For example, if the source produces a sequence of letters A, B, or C with probabilities $p_A=0.6$, $p_B=0.3$, $p_C=0.1$, and successive letters are chosen independently, then $H_n=H_1=-[0.6 \log_2 0.6+0.3 \log_2 0.3 +0.1 \log_2 0.1]=1.294$ and the information produced is equivalent to 1.294 bits for each letter of the message. A channel with a capacity of 100 bits per second could transmit with best encoding $100/1.294=77.3$ message letters per second.

XII. CONTINUOUS SOURCES

If the source is producing a continuous function of time, then without further data we must ascribe it an infinite rate of generating information. In fact, merely to specify exactly one quantity which has a continuous range of possibilities requires an infinite number of binary digits. We cannot send continuous information *exactly* over a channel of finite capacity.

Fortunately, we do not need to send continuous messages exactly. A certain amount of discrepancy between the original and the recovered messages can always be tolerated. If a certain tolerance is allowed, then a definite finite rate in binary digits per second can be assigned to a continuous source. It must be remembered that this rate depends on the nature and magnitude of the allowed error between original and final messages. The rate may be described as the rate of generating information *relative to the criterion of fidelity*.

Suppose the criterion of fidelity is the rms discrepancy between the original and recovered signals, and that we can tolerate a value $\sqrt{N_1}$. Then each point in the message space is surrounded by a small sphere of radius $\sqrt{2T_1W_1N_1}$. If the system is such that the recovered message lies within this sphere, the transmission will be satisfactory. Hence, the number of different messages which must be capable of distinct transmission is of the order of the volume V_1 of the region of possible messages divided by the volume of the small spheres. Carrying out this argument in detail along lines similar to those used in Sections VII and IX leads to the following result:

THEOREM 5: *If the message source has power Q, entropy power $\overline{Q}$, and bandwidth W_1, the rate R of generating information in bits per second is bounded by*

$$W_1 \log_2 \frac{\overline{Q}}{N_1} \leqq R \leqq W_1 \log_2 \frac{Q}{N_1} \qquad (44)$$

where N_1 is the maximum tolerable mean square error in reproduction. If we have a channel with capacity C and a source whose rate of generating information R is less than or equal to C, it is possible to encode the source in such a way as to transmit over this channel with the fidelity measured by N_1. If $R>C$, this is impossible.

In the case where the message source is producing white thermal noise, $\overline{Q}=Q$. Hence the two bounds are equal and $R=W_1 \log Q/N_1$. We can, therefore, transmit white noise of power Q and band W_1 over a channel of band W perturbed by a white noise of power N and recover the original message with mean square error N_1 if, and only if,

$$W_1 \log \frac{Q}{N_1} \leqq W \log \frac{P+N}{N}\,. \qquad (45)$$

APPENDIX

Consider the possible sequences of n symbols. Let them be arranged in order of decreasing probability, $p_1 \geqq p_2 \geqq p_3 \cdots \geqq p_s$. Let $P_i=\sum_1^{i-1} p_j$. The ith message is encoded by expanding P_i as a binary fraction and using only the first t_i places where t_i is determined from

$$\log_2 \frac{1}{p_i} \leqq t_i < 1+\log_2 \frac{1}{p_i}\,. \qquad (46)$$

Probable sequences have short codes and improbable ones long codes. We have

$$\frac{1}{2^{t_i}} \leqq p_i \leqq \frac{1}{2^{t_i-1}}\,. \qquad (47)$$

The codes for different sequences will all be different. P_{i+1}, for example, differs by p_i from P_i, and therefore its binary expansion will differ in one or more of the first t_i places, and similarly for all others. The average length of the encoded message will be $\sum p_i t_i$. Using (46),

$$-\sum p_i \log p_i \leqq \sum p_i t_i < \sum p_i(1-\log p_i) \qquad (48)$$

or

$$nH_n \leqq \sum p_i t_i < 1+nH_n. \qquad (49)$$

The average number of binary digits used per message symbol is $1/n \sum p_i t_i$ and

$$H_n \leqq \frac{1}{n}\sum p_i t_i < \frac{1}{n}+H_n. \qquad (50)$$

As $n\rightarrow\infty$, $H_n\rightarrow H$ and $1/n\rightarrow 0$, so the average number of bits per message symbol approaches H.

Multiplex Employing Pulse-Time and Pulsed-Frequency Modulation*

HAROLD GOLDBERG†, SENIOR MEMBER, IRE, AND CARL C. BATH ‡, ASSOCIATE, IRE

Summary—**A microwave communication system giving two-channel voice transmission over a single carrier is described. Diplexing is accomplished by the use of two types of modulation, rather than by time or frequency division. One channel is transmitted by a form of pulse-time modulation using 1-microsecond pulses at a mean repetition rate of 10 kc, and the other channel by frequency modulation of the pulsed microwave carrier (FM burst). The addition of the pulsed-FM channel requires no additional carrier power or change of duty cycle. Cross talk is negligible. The double system is applicable to most of the common forms of pulse-time modulation. Time-division techniques are as applicable as in systems employing pulse-time modulation alone. If a time-division index of *n* is employed, i.e., that necessary to give *n* channels in a straight pulse-time time-division multiplex system, the employment of the diplex system described gives 2*n* channels.[1] Compared to a straight pulse-time time-division system giving 2*n* channels, the double-modulation system results in economies in synchronizing equipment, and provides a better signal-to-noise ratio on the pulse-time portion of the system, but requires some additional bandwidth.**

Introduction

THIS PAPER describes a proposal, and the experimental study, of a method of diplexing a pulse-time-modulation channel by the superposition of an additional channel on the same channel pulses by the frequency modulation of the pulsed carrier. The motivation for such a system is the desire to increase the number of channels that may be carried in a pulse-time-division system having a particular time-division index *n*. (By index *n* is meant the number of channels provided in a single modulation system by a system of time division.) The experimental study was carried out on a system which provided only a single pulse-time channel, i.e., index of unity. The use of a single pulse-time channel in this study, which is eventually aimed at time-division systems, was justified by the preliminary conclusion that cross talk between pulse-time and pulsed-FM channels would be the most difficult problem, and that this cross talk would be greatest in a system of unit time-division index. The validity of these preliminary conclusions should become clear later in this paper.

The pulse-time system used in this study has not been previously reported in the literature. Its characteristics are not particularly germane to the study, except that it makes the cross-talk problem as severe as can be envisaged. The pulse-time modulating system used transmits information in terms of the periods between consecutive pulses, these periods being single-valued functions of the modulating voltage. The relationship is not a linear one, the over-all system linearity being provided by a complementary nonlinearity in the pulse-time demodulator. The ratio of maximum to minimum length of period in this system is greater than four to one for heavy modulation. Space transmission is by pulse modulation of a microwave carrier; i.e., the information is carried by the variation in time interval between consecutive carrier pulses. It is proposed to transmit an additional channel by modulating the frequency of the pulsed carrier (not the repetition rate of the pulses). It should be noted that requirements on the pulse-repetition rate imposed by the modulating frequencies impressed on the pulse-time channel are sufficient to satisfy the demands for the pulsed-FM channel as well, provided both channels have the same upper limit for the modulating frequency. A mean pulse-repetition rate greater than twice the highest modulating frequency is required. A discussion of these requirements for certain pulse-modulation systems has been given in the literature.[2]

Cross-Talk Problem

Cross talk was judged to be the most severe problem, since many successful systems may be proposed for accomplishing pulse-time or pulsed-FM transmission, but not all of these alternative methods will keep cross talk within satisfactory limits. In fact, the object of this study is to provide a system which does keep the cross talk within bounds. Two kinds of cross talk were considered; one, that arising from interactions between the two modulating processes up to but not including the final demodulator; secondly, that inherent in the demodulation process. Consider the first type in the transmitter-modulator portion of the system. Any change in the shape of the transmitter keying pulse due to the time-modulation process might conceivably cause a frequency modulation of the transmitter, and thus impose some of the pulse-time-channel information on the pulsed-FM channel. Additionally, the process of frequency-modulating the transmitter might alter the envelope of the outgoing carrier pulse even though the keying pulses were not affected, and in this way impose some pulsed-FM-channel information on the pulse-

* Decimal classification: R460×R148.6. Original manuscript received by the Institute, July 8, 1947; revised manuscript received, May 26, 1948.

† National Bureau of Standards, Washington 25, D. C.
‡ Bendix Radio Division, Bendix Aviation Corporation, Baltimore, Md.

[1] In pulse-time systems which employ a synchronizing pulse that does not carry channel information, an additional pulsed-FM channel can be sent via the synchronizing pulse, thus resulting in $2n+1$ channels.

[2] G. L. Fredendall, K. Schlesinger, and A. C. Schroeder, "Transmission of television sound on the picture carrier," PROC. I.R.E., vol. 34, pp. 49P–61P; February, 1946.

time channel. A consideration of the properties of the space transmission path and the portions of the receiver up to the demodulator shows that additional envelope distortion may occur as a result of the fact that the carrier is frequency-modulated. This can result in additional cross talk in the pulse-time channel.

The cross-talk effects in the transmitter-modulator may be minimized by proper circuit design and choice of transmitter operating parameters. Envelope delay and distortion variations with frequency over the space propagation path are generally minimized by making the frequency deviation a small percentage of the mean carrier frequency. The effect of the phase versus frequency and amplitude versus frequency characteristics of the receiver on the envelope delay and distortion as the frequency of the carrier shifts may be minimized if the amplitude response is flat, and the phase response linear, over a frequency band equal to the maximum total frequency deviation plus the width of the significant portion of the pulse spectrum. The diplex system here proposed, therefore, requires more bandwidth than the pulse-time system alone, to the extent that additional bandwidth equal to the total possible frequency deviation is necessary.

The greatest problem with respect to cross talk arises in the demodulation process. The method proposed in this report cannot, even theoretically, eliminate cross-talk effects in demodulation. It does provide a satisfactorily low level of cross talk, at least for speech circuits, even under the drastic conditions previously outlined. Once the necessary steps have been taken to minimize envelope distortion and delay, the pulse-time channel may be demodulated without further cross-talk problems. This channel is not affected as to cross-talk by the demodulation process. It is the pulsed-FM channel that gives evidence of cross talk from the pulse-time channel as a result of the demodulation process.

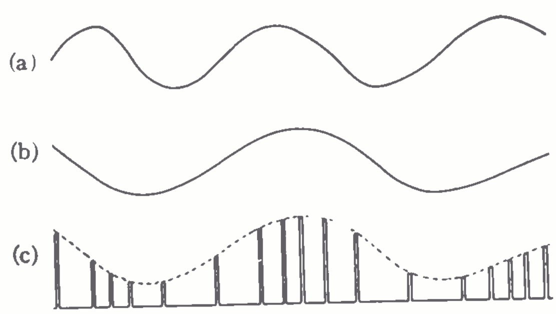

Fig. 1—(a) Pulse-time modulating voltage. (b) Pulsed-FM modulating voltage. (c) Pulse-time modulation+pulsed FM with FM converted to pulsed AM.

One might propose to demodulate the pulsed-FM channel by passing the signal through a conventional frequency discriminator, and then passing the resulting amplitude-modulated pulse train through an appropriate low-pass filter to recover the modulation. This would be satisfactory as long as no modulation was impressed on the pulse-time channel. The pulse train is modulated both in amplitude and time, however, if modulation is present on both channels. In such a case the output of the low-pass filter would then be a hopeless mixture of both channel modulations. One should note, however, that the envelope of the pulse train obtained from the discriminator is the modulation on the pulsed-FM channel only, even though pulse-time modulation is also present (see Fig. 1c). The proposed demodulation scheme, therefore, is to recover the envelope of this train as closely as possible, as the means of demodulating the pulsed-FM channel.

Fig. 2 illustrates the method for recovering a reasonable facsimile of the envelope of the discriminator output. It is proposed that the pulse train will be operated on by a circuit which will produce the stepped envelope shown in the figure. This will be termed the step-function envelope, and the process will be called step-function demodulation. The envelope is produced by pro-

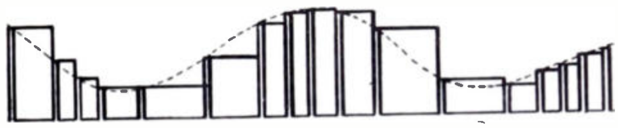

Fig. 2—Conversion of amplitude-modulated train to step function.

longing each peak pulse amplitude until the occurrence of the next pulse. The step-function envelope is passed through a low-pass filter as the final step in the demodulation process. Simple diode peak detection is not satisfactory as a means of recovering the envelope because of the low ratio of pulse-repetition frequency to modulating frequency.

The following discussion indicates the difficulties involved in using simple discriminators; outlines a partial solution requiring broad-band discriminators and additional filters; and, finally, treats the successful solution to the demodulation problem.

The production of the amplitude-modulated pulse train by means of a discriminator is not simply done by the use of a conventional discriminator as proposed. Due to the fact that a frequency-modulated pulsed carrier may be considered to be a pulse spectrum whose center frequency is varied, it is obvious that the output of a typical balanced discriminator, such as the Foster-Seeley,[8] will not be an undistorted amplitude-modulated pulse train whose envelope represents the frequency modulation. This may easily be seen by considering the output of such a discriminator when the input pulse spectrum is centered at the discriminator cross over. The output of the discriminator is not zero because a pulse spectrum is applied, not a single frequency. The output in this case is a differentiated pulse, and has the form of a pair of pulses, opposite in polarity, representing the leading and trailing edges of the applied

[8] D. E. Foster and S. W. Seeley, "Automatic tuning, simplified circuits and design practice," Proc. I.R.E., vol. 25, pp. 289–314; March, 1937.

pulse. To use such a discriminator in the proposed system, it is necessary that the distortions take place only at the leading and trailing edges of the pulse, i.e., the transient response of the discriminator must be short compared to the pulse length. If that is the case, the effects of the distortion may be suppressed by passing the discriminator output through a low-pass filter of cutoff frequency of the order of the reciprocal of the original pulse length. To make the transient response adequate requires discriminator bandwidths of the order of several times the reciprocal of the pulse width. This bandwidth is undesirable if the frequency deviations employed are of the order of the reciprocal of the pulse width. The dual problem of discrimination and step-function demodulation has been met by means of a circuit which accomplishes both functions simultaneously and which may have a bandwidth only slightly greater than the maximum total frequency deviation, even when this deviation is of the order of the reciprocal of the pulse width.

The properties of the step-function demodulation process should be carefully considered. In the absence of modulation on the pulsed-FM channel, i.e., during silence, there can be no cross talk from the pulse-time channel, since the output of the step-function demodulator is a constant voltage, independent of the degree of time modulation. Cross talk, therefore, can only occur when information is transmitted, i.e,, when the channel is not silent. The way in which this cross talk occurs is by means of the variation of the step lengths. It cannot be greater than the modulation on the pulsed-FM channel. It appears as distortion, essentially by a cross-modulation process, and is not recognizable as the interfering modulation. Psychologically, therefore, the absence of cross talk during silence, and the masking effect of the desired modulation during periods of transmission, give the effect of negligible cross talk. The absence of cross talk during silence has enormous psychological value. Quite aside from the cross-talk suppression feature, step-function demodulation is valuable as a means of pulse-communication demodulation. The repetition frequency disappears during silence and the amount of repetition-frequency amplitude in the step function envelope is generally less than the amplitude of the desired signal. The resulting masking of the repetition rate reduces the requirements on attenuation in the low-pass filters generally employed in pulse demodulators. Its value is such that the pulse-time channel used in these experiments also uses step-function demodulation. An additional gain from the use of this system of demodulation is that the signal recovery from the step-function demodulator is roughly the reciprocal of the duty cycle greater than the signal recovery from an identical pulse train by means of a low-pass filter. (The duty cycle is the product of pulse length and repetition frequency.)

Since the cross-talk distortion appearing in the pulsed-FM channel output is a function of step-length variation, it will be reduced if the variation in step length is reduced. This may be done by a decrease in depth of modulation on the pulse-time channel and by the employment of other pulse-time systems which do not result in step-length variations which follow the modulation. (There are certain systems, essentially push-pull systems, in which the sum of any two adjacent pulse-to-pulse time intervals is almost a constant.) The conditions of this study were severe because the step-length variations were deliberately large. Regardless of the type of pulse-time modulation used, the employment of time division to increase the number of channels reduces this cross-talk distortion, provided all other factors remain constant. To illustrate, a single-channel system is first assumed. The channel pulse-repetition rate is determined by the highest modulating frequency to be employed, and is $1/\tau$ where τ is the pulse-repetition period. Ignoring the reference pulse, the upper and lower limits of the interval between adjacent-channel pulses are approximately 2τ and 0. The step length in such a system can therefore range between 0 and 2τ in length. If time division is now employed to allow the transmission of n channels, and the highest modulating frequencies employed are the same as for the single-channel system, the channel repetition rate must still be $1/\tau$, independent of n. The total number of pulses transmitted per second, excluding the reference or synchronizing pulse, is n/τ. Since $(n-1)$ pulses will occur between any two consecutive channel pulses, the time swing available for any channel is reduced compared to the single-channel case. The upper and lower limits for the time interval between any two consecutive channel pulses becomes $\tau(1+1/n)$ and $\tau(1-1/n)$. The pulsed-FM channels are carried by their complementary pulse-time channel pulses. The time interval between any two consecutive pulses for any FM channel is the same as for the pulse-time channels. It is obvious, therefore, that the time interval between any two consecutive channel pulses tends to a constant τ, as n increases without limit for both pulse-time and pulsed-FM channels. An increase in n, therefore, reduces the maximum signal-to-noise ratio that may be achieved, per channel, in the pulse-time channels, assuming that the peak power and pulse length has not been changed in the time-division process. If it is assumed that time division is not carried to the point of interference between adjacent pulses (this is a necessary limitation for both pulse-time and pulsed-FM), the carrier frequency deviation employed to transmit information over pulsed-FM channels is independent of the time-division process. The peak power and pulse length have been assumed to have been kept constant. The result is a signal-to-noise ratio in the pulsed-FM channels independent of the degree of time division employed. Since it has already been shown that the interval between consecutive channel pulses tends to a constant as n increases, the step lengths in the FM demodulation process will tend toward a constant, and the cross talk from pulse time into pulsed FM will tend towards zero.

Signal-to-Noise-Ratio Advantages

An m-channel system may be provided by a single-modulation time-division multiplex of index m, or by a diplex system of index $m/2$. If all factors are kept constant in a pulse-time system with the exception of the number of channels, a lower index allows a correspondingly higher maximum signal-to-noise ratio. Diplexing of the sort proposed in this report increases the possible signal-to-noise ratio of the pulse-time channels by reducing the time-division index. It was found experimentally that the pulsed-FM channels could be given signal-to-noise performance equivalent to pulse-time systems of index one. In a time-division system, while the maximum possible performance of the pulse-time channels is reduced, the pulsed-FM channels, which do not carry information by means of pulse position, are not affected as to signal-to-noise performance (assuming that time division is not carried to the point of interference between adjacent pulses).

Additional Systems

Pulsed FM may be diplexed to pulsed-AM systems, pulse-width modulated systems, and pulse-code systems.

Experimental Equipment

The test gear used in this study consisted of two complete transmitter-receiver units, each diplexed, providing two-way two-channel operation. Carrier frequencies in the neighborhood of 3,000 Mc were employed. The gear was designed for voice transmission; emitted a peak pulse power of approximately 1.5 watts; used a pulse length of 1 microsecond, a mean repetition rate of 10 kc, and an average power output of about 15 milliwatts. Tests were conducted with a loop around one transmitter-receiver unit, and with space paths of up to 8 miles. Horn radiators were used.

The transmitter block diagram is shown in Fig. 3. The input audio amplifiers have the frequency char-

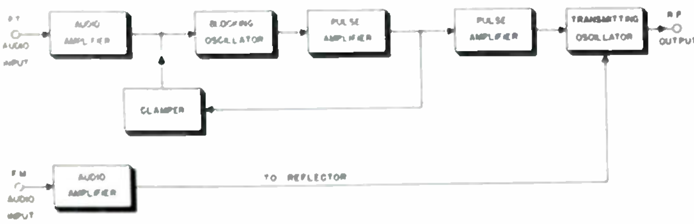

Fig. 3—Transmitting system; block diagram.

acteristics shown in Fig. 4, and are designed to take care of the requirements imposed by voice transmission and pulse modulation. No additional filters are employed to increase the rate of high-frequency cutoff. The pulse-time modulator, Fig. 5, is a blocking oscillator whose "off" time is varied by injecting the modulating voltage into the grid circuit. A two-stage pulse amplifier and clipper steps up the pulse power, shapes the pulse, and removes any amplitude modulation introduced by the

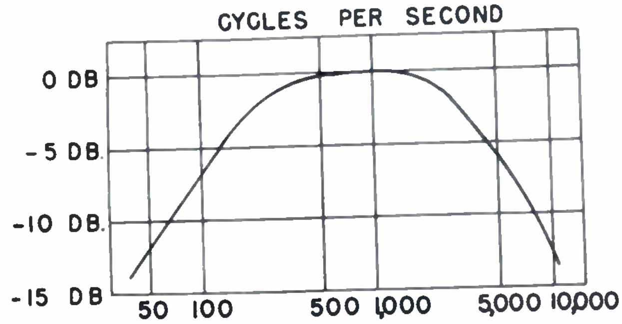

Fig. 4—Input audio-amplifier frequency response.

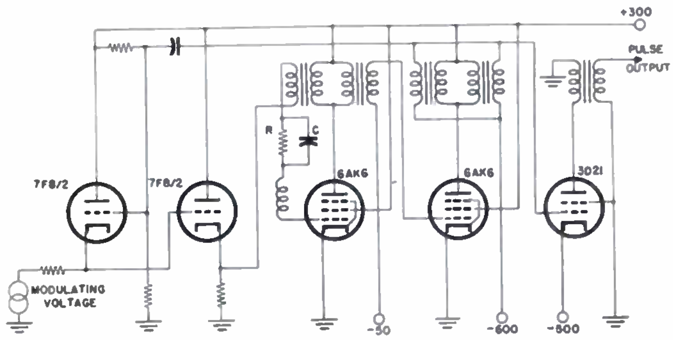

Fig. 5—Pulse-time modulator.

modulator. The output pulse power is in the neighborhood of 1 kilowatt peak and may be adjusted over a narrow range in that neighborhood. Pulse is fed back to a clamper in the modulating circuit. The feed back makes the voltage to which the blocking-oscillator capacitor charges independent of the modulating voltage. Typical wave forms are shown in Figs. 6 through 10.

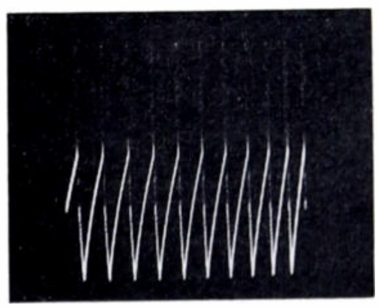

Fig. 6—Blocking-oscillator grid, no modulation.

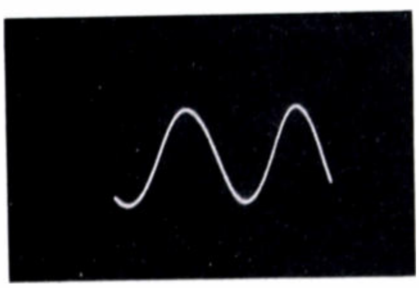

Fig. 7—Modulating voltage.

The carrier oscillator, a 707B reflex klystron, is anode and grid pulsed. Since the impedance of the klystron is

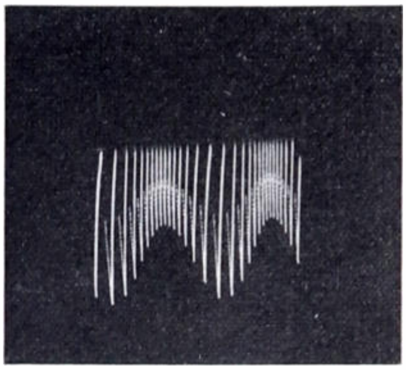

Fig. 8—Blocking-oscillator grid, modulated.

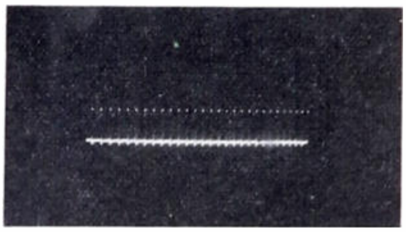

Fig. 9—Pulse output, no modulation.

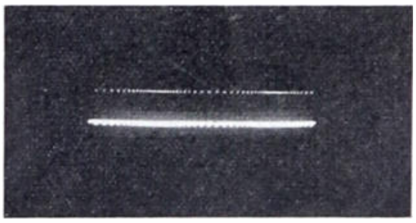

Fig. 10—Pulse output, modulated.

relatively high, it is shunted by a resistor to allow fast rise and fall of the applied pulse. Thus, only about one-fourth of the 1-kilowatt keying pulse is actually applied to the klystron. The klystron supplies between 1 and 2 watts of peak power output under these conditions. The transmitter is frequency-modulated by applying a modulating voltage to the klystron reflector. The peak deviation employed was ±1 Mc. It was found that the system was noisy unless a good carrier pulse spectrum was obtained. Operating conditions for good output spectrum can be found, and these stay substantially constant for any given klystron. Interaction between frequency modulation and output pulse envelope is kept to a minimum by the use of deviations small compared to the electrical tuning range of the klystron. A satisfactory carrier pulse spectrum is shown in Fig. 11. The spectrum shows the typical energy distribution expected of a rectangular pulse. It is quite free of the aberrations introduced by frequency modulation during the pulse interval and multiple moding of the oscillator. Noisy conditions can be correlated with deviations from the theoretical rectangular-pulse spectrum.

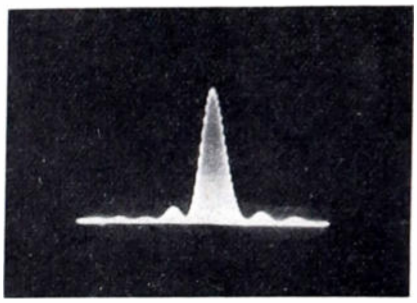

Fig. 11—Pulse output spectrum.

The receiver block diagram is shown in Fig. 12. The receiver is a superheterodyne employing a 2K41 reflex klystron as the local oscillator, a 1N21B silicon-crystal-mixer, and an if amplifier system centered at 60 Mc. A five-stage if preamplifier, 10 Mc wide between half-power points, amplifies the intermediate-frequency sig-

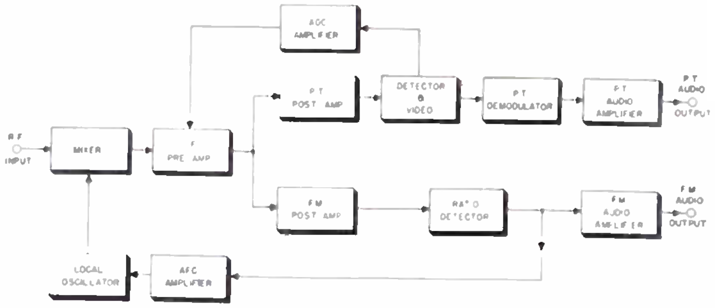

Fig. 12—Receiving system; block diagram.

nal coming from the mixer and drives a 75-ohm coaxial line. This line terminates in the inputs of two if post-amplifiers. One of these, a three-stage amplifier 10-Mc wide between 3-db points, is associated with the pulse-time demodulator and the receiver agc circuit. Amplified and delayed agc is employed.

The other if post-amplifier is associated with the pulsed-FM demodulator and the receiver afc control. It has a half-power bandwidth of 6 Mc, employs two stages, and drives a step-function pulsed-FM demodulator which has a linear range of about 2.5 Mc. The step-function pulsed-FM demodulator is shown schematically in Fig. 13. Fig. 14 is a typical waveform

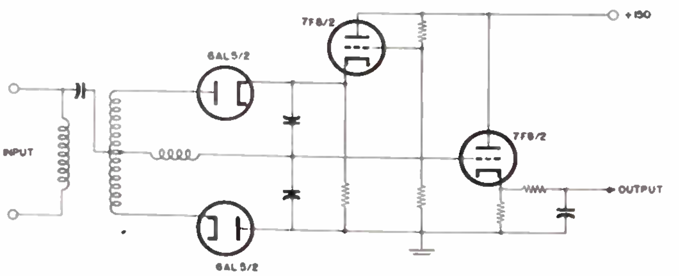

Fig. 13—Modified ratio detector.

of the output of this demodulator when the system is modulated with a sine wave. The demodulator is a modification of the Seeley ratio detector.[4] The modification consists of replacing the "battery" in the original Seeley circuit by the output terminals of a cathode follower and providing a very high-impedance take-off for the output of the demodulator (a cathode follower). The operation is explained in the following way: Conduction

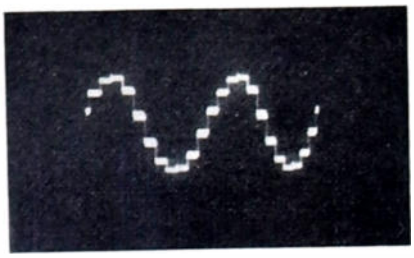

Fig. 14—Step-function output of detector.

of the diodes during a pulse causes the output terminal to assume an equilibrium voltage which is a function of the pulse carrier frequency. This equilibrium is reached in less than 1 microsecond, the pulse length employed in the study. Upon removal of the pulse, the diodes cease to conduct because of the threshold imposed by the fixed "battery" voltage. Since no discharge path is provided for the output point, the equilibrium voltage established by any one pulse must be maintained until the next consecutive pulse establishes a new equilibrium voltage corresponding to its frequency. Thus, the step-function demodulation is achieved automatically. It has been found experimentally that proper operation is achieved even when the frequency separation between discriminator peaks is not large with respect to the reciprocal of the pulse length. (A width of 2.5 Mc was used with ± 1 Mc deviation and 1 microsecond pulse.) The output of the cathode-follower take-off is passed through a simple *RC* low-pass filter to give the result typified by the waveform in Fig. 15. An audio amplifier having the response curve shown in Fig. 16 completes the pulsed-FM channel. No other low-pass filters are employed. Nevertheless, the 10-kc mean repetition rate is below audibility in the output.

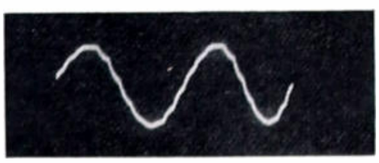

Fig. 15—Integrated step function.

The step-function demodulator removes noise between pulses, but does not adequately suppress noise appearing on the pulse. As a result, the quieting threshold is not adequately sharp. A limiter preceding the demodulator should improve performance in this respect. Output signal-to-noise ratios better than 50 db have been achieved with the pulsed-FM channel. Cross talk from the pulse-time channel, during silence on the pulsed-FM channel, is better than 60 db down with full modulation on the pulse-time channel. The demodulator output, after passing through a low-pass filter, is amplified by a dc amplifier and applied to the reflector of the local-oscillator klystron in proper phase to provide automatic frequency control.

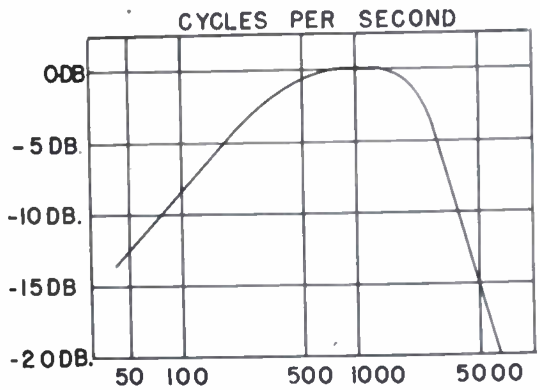

Fig. 16—Output audio-amplifier frequency response.

The pulse-time if post-amplifier drives a video pulse detector and agc rectifier. The output of the agc rectifier passes to a dc amplifier whenever the rectifier voltage exceeds a fixed threshold. The output of the dc amplifier controls the gain of the if preamplifier. The output pulses generated by the video pulse detector are passed through a pulse amplifier and "slicer." The "slicer" removes a slice from the pulse in the region of steepest pulse rise and fall. This gates the output pulse train with respect to noise between pulses and noise on pulse tops. Any input signal-to-noise ratio greater than 6 db allows removal of a slice that is noise free, except for noise components in the leading and trailing edges of the slice. Thus, a very sharp quieting threshold is achieved. The "slicer" also permits discrimination against interfering pulse signals which are less than the signal pulses in amplitude. Radar interference, occasionally encountered during tests, was often eliminated by proper adjustment of the "slicer" level.

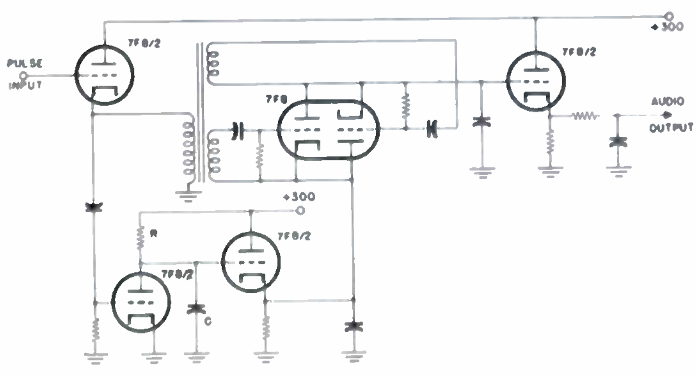

Fig. 17—Pulse-time demodulator.

[4] J. Avins, "The Ratio Detector," RCA Licensee Bulletin LB-712, May 12, 1947.

The "slice" is amplified and passed on to the pulse-time step-function demodulator (see Fig. 17). The amplified slice is passed to a pulse transformer by means of a cathode follower. The three outputs of this transformer operate a sawtooth generator and a double clamper. The sawtooth generator is a series *RC* charging circuit which is discharged by a triode when its grid receives a positive pulse. The *RC* time constant is identical to that in the transmitter pulse-time modulator. Variation in the period between pulses causes variation in the peak amplitude of the sawtooth. The result of the process is an amplitude-modulated sawtooth typified by the waveform of Fig. 18. Note that the return time of the sawtooth must of necessity be less than the pulse length of the discharging pulse.

Fig. 18—Sawtooth-generator output, modulated.

Step-function demodulation is to be accomplished in the manner illustrated in Fig. 19. Two cathode followers and the double clamper accomplish this, as shown in Fig. 21. A small capacitor is allowed to "view" the peaks of the sawtooth for a fraction of a microsecond in such a manner that the capacitor voltage is made equal to the peak voltage during the viewing interval, and is constant at that voltage until the next viewing interval. The "viewing" is done through the double clamper, a bidirectional pulse-operated triode switch. The voltage on the capacitor is transferred by means of a cathode follower. To properly "view" the peak the sawtooth output must be at low impedance, and some delay must be provided, since the clamper is operated at the time the sawtooth generator is discharged. Both requirements are met by passing the sawtooth through a second cathode follower which has an *RC* cathode load. The capacitor in the cathode circuit is insufficient to delay the rise time of the follower appreciably, but causes the decay time of the follower to be appreciably greater than the microsecond return time of the sawtooth, thus giving the necessary delay. The mechanism involves cutoff of the follower during the return time, and has been discussed in the literature.[5] The combination of cathode follower and low-impedance load give the requisite low impedance for the "viewing" operation. The output of the second follower is typified by the waveform of Fig. 20. The output of the demodulator is passed through an *RC* low-pass filter to provide the result of Fig. 22, and is then amplified in an audio amplifier having the characteristics shown in Fig. 16.

Fig. 19—Conversion of sawtooth to step function.

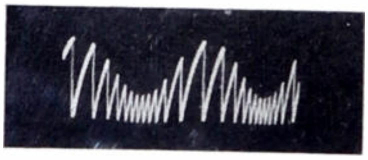

Fig. 20—Follower sawtooth output.

Fig. 21—Step-function output.

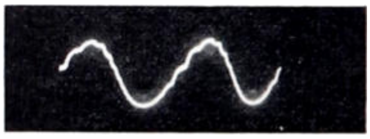

Fig. 22—Integrated step function.

Conclusions

Field and laboratory tests have shown the described diplex system to be very satisfactory for voice communication. It may be used, with appropriate repetition rates, for program service where the pulsed-FM channel is used as an order wire. Cross talk, measured during silence on the desired channel with full modulation on the interfering channel, can be kept 60 db below maximum output on the desired channel. This applies to either channel.

[5] Harold Goldberg, "Some considerations concerning the internal impedance of the cathode follower," Proc. I.R.E., vol. 33, pp. 778–783; November, 1945.

Theory of the Superregenerative Amplifier*

LEON RIEBMAN†, ASSOCIATE, IRE

Summary—A very general solution of a superregenerative circuit from which reception of almost any type of modulation can be analyzed is derived. Familiarity with this impedance method of analysis should result in a clear physical picture of superregenerator operation.

Examples are worked out on the basis of an ideal control wave form, revealing the fundamental limitations inherent in a superregenerative amplifier. It is shown that the minimum realizable static bandwidth of a superregenerator is 0.89 f_q. Although the ideal wave form can only be approached in practice, it serves as a useful analytical tool in clarifying how the various factors affect the operation of a superregenerator.

Introduction

Superregeneration was first introduced by E. H. Armstrong in 1922. Until recently, the principle has not come into common use because of the difficulty of understanding the underlying theory.

The analysis appearing in this paper is based on an approximate solution of the differential equation for the response of a tuned circuit containing a varying resistance to an applied signal.[1] To a first approximation, this varying resistor is assumed to be independent of the tuned-circuit voltage. The approximate solution of the resultant linear differential equation is found to hold very well for most superregenerative applications. This solution permits obtaining the output spectrum of a superregenerative amplifier and reveals the factors affecting amplification, shape and width of frequency response, maximum control frequency, and, in general, gives a "physical" picture of superregenerator operation.

Superregeneration is a form of regenerative amplification in which the circuit oscillations are made to increase and decay at a rate that is usually small compared to the resonant frequency of the tuned circuit. The control action may be obtained either by varying the plate voltage or grid bias of the amplifier tube, or by varying an auxiliary resistance. An analysis of the control action is extremely simplified by substituting for the control voltage the equivalent variation of the effective resistance. The frequency of the variation of the effective resistance is called the control frequency.

Then the effective resistance of the tuned circuit of the superregenerative amplifier may be taken to be composed of the following three terms:

1. The ohmic conductance G_0.
2. The conductance $G(t, e)$ due to the control action and also a function of the tuned-circuit oscillations when the superregenerator operates in the logarithmic mode.
3. The negative conductance $\gamma(e)$ due to regeneration, which is a function of the tuned-circuit voltage e.

The negative conductance $\gamma(e)$ can be expressed as a power series in e. Thus this term can cause cross-modulation terms when the signal is impressed. In most practical cases this term is much less than $G(t, e)$, and can be neglected.

Assuming a linear mode of operation for the superregenerator, $G(t, e)$ becomes $G(t)$.

Neglecting $\gamma(e)$ and setting $g(t) = G_0 + G(t)$, the equivalent circuit of Fig. 1 reduces to Fig. 2.

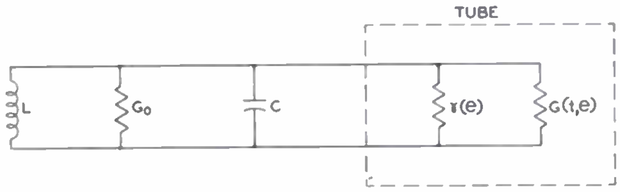

Fig. 1—Equivalent circuit of a superregenerative amplifier.

The superregenerative cycle may be described in terms of two phases:

1. Degenerative phase.
2. Regenerative phase.

The degenerative phase may be further subdivided into two subphases:

a) Quench phases—the circuit oscillations from the previous regenerative phase decay to a minimum value determined by the input signal or the thermal noise, whichever is greater.

b) Sensitive phase—the input signal current establishes an amplitude of oscillation on the tuned circuit.

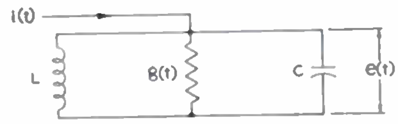

Fig. 2—Reduced equivalent circuit.

During the quench phase, $g(t)$ should be as large a positive value as possible in order to cause very rapid decay of oscillations from the previous regenerative phase. The value of $g(t)$ during the sensitive phase is determined by the prescribed bandwidth requirements and the shape of the variation of $g(t)$ during this phase.

At the instant between phases, $g(t)$ passes through zero, and the circuit oscillations, as determined by the previous sensitive phase, establish an initial amplitude for the regenerative phase.

The regenerative phase can also be subdivided into two subphases:

(a) Amplification phase—the circuit oscillations grow under compulsion of the negative $g(t)$.

(b) Power phase—when the logarithmic mode is considered, the circuit oscillations grow until an amplitude

* Decimal classification: R363.23×R132. Original manuscript received by the Institute, March 16, 1948; revised manuscript received, August 10, 1948. This paper is from a master's thesis for the Moore School of Electrical Engineering, University of Pennsylvania.

† Philco Corporation, Philadelphia, Pa.

[1] W. E. Bradley, "Theory of the superregenerative receiver," presented, 1948, National IRE Convention, New York, N. Y., March 23, 1948.

is reached that causes overload of the electron tube responsible for the negative component of the conductance. As long as overload persists, the conductance is zero.[2] During this phase, maximum power is being fed to the load. As was mentioned previously, only the linear mode will be considered, and for this mode of operation the power phase does not exist.

Theory

The details of the behavior during the complete cycle are far from obvious, and it is necessary to study the equivalent circuit in Fig. 2 in order quantitatively to evaluate the effects of the various circuit parameters. The differential equation for the tuned circuit, consisting of a constant capacitance C, constant inductance L, and the variable conductance $g(t)$ in parallel, can be written in the form

$$\frac{d^2e}{dt^2} + P\frac{de}{dt} + (\omega_0^2 + P')e = i'/c \qquad (1)$$

where

$$P = \frac{g(t)}{C}, \quad P' = \frac{dP}{dt}, \quad \omega_0 = \sqrt{\frac{1}{LC}}, \quad i' = \frac{di}{dt}.$$

For the case

$$\left|\frac{P^1}{2} - \frac{P^2}{4}\right| \ll \omega_0^2,$$

a good approximation to the solution of the above equation is

$$e(t) = \text{real}^3 \left[E(t_0)\epsilon j^{\omega_0(t-t_0)}\epsilon^{1/2\int_t^{t_0} P dt} + \frac{\epsilon j^{\omega_0 t}}{j\omega_0 C}\int_{t_0}^{t} i'(\tau)\epsilon^{1/2\int_t^{\tau} P dt}\epsilon^{-j\omega_0\tau}d\tau \right] \qquad (2)$$

where real[3] $[E(t_0)] = e(t_0)$ contains the two constants of integration (i.e., amplitude and phase), assuming that $e(t)$ is known at some arbitrary time t_0.

At this point, assume a signal of varying amplitude $I = I_{(t)}$ and phase $\theta = \theta_{(t)}$.

$$i = I \sin(\omega_c t + \theta), \qquad I(t) = I\epsilon^{j\theta}. \qquad (3)$$

If the signal is assumed to be varying at a rate that is slow compared to the carrier frequency ω_C, and further that $|\omega_C - \omega_0| \ll \omega_C + \omega_0$, $\omega_C/\omega_0 \cong 1$, (2) reduces to

$$e(t) = E(t_0)\epsilon j^{\omega_0(t-t_0)}\epsilon^{1/2\int_t^{\tau} P dt} + \frac{\epsilon j^{\omega_0 t}}{2jC}\int_{t_0}^{t} I(t)\epsilon^{1/2\int_t^{\tau} P dt}\epsilon^{j(\omega_C-\omega_0)\tau}d\tau. \qquad (4)$$

[2] Strictly speaking, $g(t)$ remains slightly positive due to the small driving force that is constantly supplying energy to the system. As the average energy of the system remains constant during the power phase, $g(t)$ takes on a positive value sufficient to dissipate the impressed energy.

[3] Hereinafter, in all equations for $e(t)$ the real part of the right-hand member is implied.

Equation (4) is very similar to the solution obtained by Bradley.[1]

Limiting the maximum interval of the integral to T_Q (control period), and assuming that the input signal varies slowly as compared to the control frequency, so that $I(t)$ may be assumed constant within the interval, (4) becomes

$$e(t) = G(t, t_0)E(t_0)\epsilon j^{\omega_0(t-t_0)} + I(t_0)Z(t_0, t)\epsilon j^{\omega_0 t} \qquad (5)$$

where

$$G(t, t_0) = \epsilon^{1/2\int_t^{t_0} P dt}$$

$$Z(t_0, t) = \frac{1}{2jC}\int_{t_0}^{t} G(t, \tau)\epsilon j^{(\omega_C-\omega_0)\tau}d\tau.$$

The above method of analysis can now be applied to a superregenerative wave form. (See Fig. 3.)

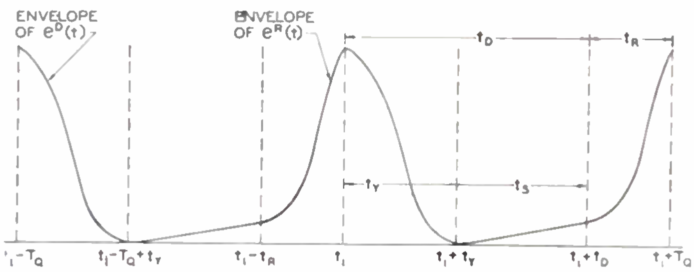

Fig. 3—Typical superregenerative wave form (not to scale).

The symbols to be used in the remainder of this paper are defined as follows:

$e^D(t)$ = voltage across tuned circuit during degenerative phase

$e^R(t)$ = voltage across tuned circuit during regenerative phase

t_D, t_R = duration of degenerative and regenerative phases respectively. $t_D + t_R = T_Q$, the control period

t_r, t_S = duration of decay and sensitive phases respectively. $t_r + t_S = t_0$

t_i = value of time at the beginning of any arbitrary degenerative phase sufficiently long after the turning on of the superregenerator that the starting transients may be neglected

P_D, P_R = values of P during degenerative and regenerative phases, respectively

$G^D(t) = G(t, t_i)$ for $t_i \leqq t \leqq t_i + t_D$ = decay factor

$G^R(t) = G(t_i t_i + t_D)$ for $t_i + t_D \leqq t \leqq t_i + T_Q$ = growth factor

$Z^D(t) = Z(t_i, t)$ for $t_i \leqq t \leqq t_i + t_D$ = active degenerative impedance

$Z^R(t) = Z(t_i + t_D, t)$ for $t_i + t_D$, $t \leqq t_i + T_Q$ = active regenerative impedance.

Limiting the interval to $t_i \leqq t \leqq t_i + T_Q$ and considering (5) for the case $e(t_0) \neq 0$ and $t_0 = t_i$,

$$e^D(t) = G^D(t)E(t_i)\epsilon j^{\omega_0(t-t_i)} + I(t_i)Z^D(t)\epsilon j^{\omega_0 t}, \qquad t_i \leqq t \leqq t_i + t_D \qquad (6)$$

$$e^R(t) = G^R(t)E(t_i + t_D)\epsilon j^{\omega_0(t-t_i+t_D)} + I(t_i + t_D)Z^R(t)\epsilon j^{\omega_0 t}, \quad t_i + t_D \leqq t \leqq t_i + T_Q. \tag{7}$$

As the interval t_D between two consecutive regenerative phases is made shorter, a point will be reached where a considerable residual voltage will exist at the time when the regeneration starts again. The condition under which the residual voltage is much less than the input signal voltage is called incoherent. This is by far the most important mode of operation for a superregenerator, and will be considered in the remainder of this paper.

For incoherence, it is obvious that the term containing $E(t_i)$ must vanish in (7); therefore,

$$e^R(t) = [G^R(t)I(t_i)Z^D(t_i+t_D) + I(t_i+t_D)Z^R(t)]\epsilon j^{\omega_0 t}. \tag{8}$$

Noting that P_D and P_R are cyclic at the control period, and further that $\epsilon^{\omega_0 t}$ is cyclic at the control period due to the imposed condition of incoherence, $E(t_i)$ can be evaluated. Neglecting the variation of the input signal in one quench period, $I(t_i - T_Q) \cong I(t_i - T_Q + t_D)$. Equation (6) becomes

$$e^D(t) = [G^D(t)Z_P I(t_i - T_Q) + I(t_i)Z^D(t)]\epsilon j^{\omega_0 t} \tag{9}$$

where

$$Z_P = G^R(t_i + T_Q)Z^D(t_i + t_D) + Z^R(t_i + T_Q).$$

Equations (8) and (9) completely determine the voltage variation during a superregenerative cycle.

Mathematically, the condition for incoherence becomes

$$G^D(t_i + t_D)\,|Z_P| \ll Z^D(t_i + t_D), \tag{10}$$

or, approximately,

$$G^D(t_i + t_D)G^R(t_i + T_Q) \ll 1. \tag{11}$$

Equation (11) simply states that the decay of circuit oscillations must be greater than the growth in order to obtain incoherence.

If the regenerative amplification A_R is defined as the increase in amplitude of circuit oscillations due to regenerative buildup, then

$$A_R = \frac{e^R(t_i + T_Q)}{e^R(t_i + t_D)} \cong G^R(t_i + T_Q) + \frac{Z^R(t_i + T_Q)}{Z^D(t_i + t_D)}. \tag{12}$$

In a practical circuit it is, therefore, desirable to use a tube with a high transconductance-to-capacitance ratio, and to keep stray circuit capacitance to a minimum in order to obtain maximum amplification.

If it is assumed that $\omega_Q = a\omega_S$ and $\omega_C = b\omega_Q$, the superregenerative wave form can be expanded into a Fourier series over one period of the input signal variation T_S.

Noting that the second term in (9) contributes negligibly to the Fourier series (although this term is extremely important to the operation of the superregenerator), and further that

$$Z^R(t_i + T_Q) \ll G^R(t_i + T_Q)Z^D(t_i + t_D)$$

under most conditions, the above equations can be rewritten for the $(n+1)$th cycle:

$$e^R(t + nT_Q) = G^R(t)Z^D(t_D)I(nT_Q)\epsilon j^{(\omega_0 t + n\omega_c T_Q)}, \quad t_D \leqq t \leqq T_Q \tag{13}$$

$$e^D[t + (n+1)T_Q] = G^D(t)G^R(T_Q)Z^D(t_D)I(nT_Q)\epsilon^{j[\omega_0 t + (n+1)\omega_c T_Q]}, \quad 0 \leqq t \leqq t_D. \tag{14}$$

Equations (13) and (14) are in a form that can be easily expanded into a Fourier series. To simplify notation, the case of a frequency-modulated input signal and the case of an amplitude-modulated input signal will be treated separately.

For FM, the sampling function $I(nT_Q)$ is

$$I(nT_Q) = I_0\epsilon j^{m_f \sin n\omega_s T_Q},$$

and the expansion becomes

$$e^F(t) = \sum_{\mu=-\infty}^{\infty} C_u{}^Z \epsilon^{ju\omega_Q t} I_0 \epsilon^{j\sum_{s=0}^{\infty} B_s^\theta \sin s\omega_s t} \tag{15}$$

where

$$C_u^Z = \frac{Z^D(t_D)}{T_Q}\left[G^R(T_Q)\int_0^{t_D} G^D(t)\epsilon^{j(\omega_0-\mu\omega_Q)t}dt + \int_{t_D}^{T_Q} G^R(t)\epsilon^{j(\omega_0-u\omega_Q)t}dt\right] \tag{16}$$

$$B_S{}^\theta = m_f \frac{\sin\frac{\pi S}{a}}{\pi S}\left[\frac{\sin 2\pi(S-1)\cos\frac{\pi}{a}(S-1)}{\sin\frac{\pi}{a}(S-1)} - \frac{\sin 2\pi(S+1)\cos\frac{\pi}{a}(S+1)}{\sin\frac{\pi}{a}(S+1)}\right]. \tag{17}$$

See Appendix I for derivation of $B_S{}^\theta$.

For AM,

$$I(nT_Q) = I_0(1 + m \sin n\omega_S T_Q),$$

and the expansion becomes

$$e^A(t) = \sum_{u=-\infty}^{\infty}\sum_{S=0}^{\infty} C_u{}^Z \epsilon^{ju\omega_Q t} I_0 B_S^A \sin S\omega_S t \tag{18}$$

where

$$B_S^A = 1 + \frac{m}{m_f} B_S{}^\theta.$$

To proceed further, it is necessary to know the form of the control voltage. The following observations are useful in determining an ideal control wave form:

1. The output spectrum of the superregenerator consists of frequencies spaced by multiples of the control frequency ω_Q, and containing the carrier frequency

$\omega_C = b\omega_Q$. Each of these frequencies has double sidebands produced by the signal variation.

2. The signal-frequency harmonic distortion is a function of the ratio of the control frequency to the signal frequency a.

3. The decay factor $G^D(t)$ determines the length of the decay phase. This decay time, t_Y, should be made as short as is consistent with the requirements for incoherence.

4. The growth factor $G^R(t)$ determines to a large degree the length of the amplification phase; the amplification time t_R should be made as short as is consistent with the required amplification.

5. It is desirable to make the sensitive time t_S as long as is consistent with the control period in order to make the second term in (9) as large as possible.

An ideal control wave form can now be formulated.[1,4] From the above discussion, it is clear that it is desirable to have steep-sided wave forms. The ideal control wave form is illustrated in Fig. 4.

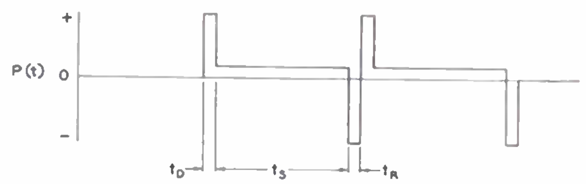

Fig. 4—Ideal control wave form (not to scale).

Static Bandwidth of the Superregenerator

Considering the case of an ideal control wave form,[5] the static bandwidth (i.e., bandwidth measured by using an unmodulated carrier) is completely determined by Z_P from (9). Performing the proper integration (see Appendix II), the static bandwidth $\Delta\omega_S$ becomes

$$\Delta\omega_S = \frac{P_S}{\sinh \frac{1}{2}P_S t_S}\left(\cosh^2 \tfrac{1}{4}P_S t_S - \cos\frac{\Delta\omega_S t_S}{2}\right)^{1/2}. \quad (19)$$

The minimum value of the above expression occurs when $P_S = 0$. For this case,

$$\Delta\omega_S = \frac{2^{5/2}}{t_S}\left|\sin\frac{\Delta\omega_S t_S}{4}\right|. \quad (20)$$

Solving this equation graphically,

$$\Delta f_S = 0.89/t_S \cong 0.89 f_Q. \quad (21)$$

This is an extremely important relationship, and shows that the minimum realizable static bandwidth is limited by the control frequency. In order to achieve narrow bandwidth, it is necessary not only to have the Q infinite during the sensitive phase, but also to have a low quench frequency.

Sideband Gain

Assuming that the incoming carrier ω_C is tuned to the center frequency of the superregenerator $\omega_0 = \omega_C = b\omega_Q$, the amplitude coefficient becomes

$$C_u{}^Z = \frac{1}{2jT_QC}\left\{\frac{\epsilon^{-1/2P_S t_S}(1-\epsilon^{-1/2P_Y t_Y})}{\frac{1}{2}P_Y} + \frac{1-\epsilon^{-1/2P_S t_S}}{\frac{1}{2}P_S}\right\}$$
$$\cdot\left\{\epsilon^{-1/2P_R t_R}\left[\frac{\epsilon^{-1/2P_S t_S}(1-\epsilon^{-1/2P_Y t_Y - ja\omega_Q t_Y})}{\frac{1}{2}P_Y + ja\omega_Q} + \frac{\epsilon^{-ja\omega_Q t_Y} - \epsilon^{-1/2P_S t_S - ja\omega_Q t_S}}{\frac{1}{2}P_S + ja\omega_Q}\right] + \frac{\epsilon^{-ja\omega_Q t_D} - \epsilon^{-1/2P_R t_R}}{\frac{1}{2}P_R + ja\omega_Q}\right\} \quad (22)$$

where $a = b - u$; a is, therefore, the order of the sideband. Subject to the approximations used in Appendix II,

$$C_u{}^Z = \frac{A}{2jCT_Q}\left\{\frac{1-\epsilon^{-1/2P_S t_S}}{\frac{1}{2}P_S}\right\}\left\{\frac{1-\epsilon^{-1/2P_S t_S + ja\omega_Q t_S}}{\frac{1}{2}P_S + ja\omega_Q}\right\}. \quad (23)$$

Taking the absolute value,

$$\left|C_u{}^Z\right| = \frac{A}{2CT_Q}\left(\frac{1-\epsilon^{-1/2P_S t_S}}{\frac{1}{2}P_S}\right)\sqrt{\frac{1+\epsilon^{-P_S t_S} - 2\epsilon^{-1/2P_S t_S}\cos a\omega_Q t_S}{\frac{1}{4}P_S{}^2 + a^2\omega_Q{}^2}}. \quad (24)$$

The ratio R of the gain at center frequency to the gain at any sideband is

$$R = (1-\epsilon^{-1/2P_S t_S})\sqrt{\frac{1+\dfrac{4a^2\omega_Q{}^2}{P_S{}^2}}{1+\epsilon^{-P_S t_S} - 2\epsilon^{-1/2P_S t_S}\cos a\omega_Q t_S}}. \quad (25)$$

There are two special cases of interest:

1. P_S is finite and positive, and $t_S \cong T_Q$. For this case,

$$R \cong \sqrt{1 + \frac{4a^2\omega_Q{}^2}{P_S{}^2}}. \quad (26)$$

For minimum sideband gain, it is seen that the ratio ω_Q/P_S should be as high as possible.

2. The minimum sideband gain occurs when $P_S = 0$. For this case,

$$R = \left|\frac{\frac{1}{2}a\omega_Q t_S}{\sin \frac{1}{2}a\omega_Q t_S}\right| = \left|\frac{x}{\sin x}\right|, \quad x = \tfrac{1}{2}a\omega_Q t_S. \quad (27)$$

It is clear that, as $t_S \to T_Q$, the sideband gain approaches zero. In applications where sideband response causes interference, it is very important to make the decay and amplification phases as short as possible.

Acknowledgments

The author wishes to acknowledge the assistance of William E. Bradley, Director of Research, Philco

[4] United States Patent No. 2,171,148, W. S. Percival.

[5] It might seem that the original assumption of $|P'/2 - P^2/4| \ll \omega_0{}^2$ has been violated by using control waveforms of infinite slope. However, in any practical case the above assumption holds very well, and the ideal control wave form can be considered as a limiting case of a practical control wave form. The added effects produced by the very steep slope are either negligible or can be minimized in practice and need not be considered.

Corporation, and J. G. Brainerd, Professor in the Moore School of Electrical Engineering, University of Pennsylvania, in the preparation of this material, and of Miss Eve Freyer of the Technical Report Group, Philco Corporation, for her assistance in the final preparation of this manuscript.

Appendix I

Derivation of Fourier Series for Sampling Function

The series of steps in Fig. 5 illustrate very closely the manner in which the sampling function, $I(nT_Q)$, varies.

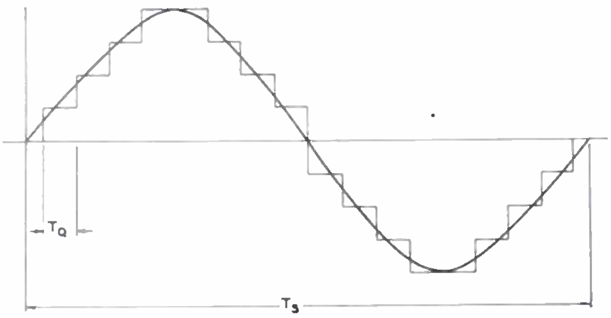

Fig. 5—Superregenerative sampling function.

The Fourier series for the sampling function can be derived as follows:

Let

$$\frac{T_S}{T_Q} = a \qquad (a \text{ is an integer})$$

$$B_S = \frac{m_f}{T_S}\int_{(n-1/2)T_Q}^{(n+1/2)T_Q} \sum_{n=0}^{a-1} \sin\frac{2\pi n}{a} \sin\left(S\frac{2\pi t}{T_S}\right)dt.$$

By simple trigonometric substitution and summation,

$$B_S = \frac{m_f}{\pi S}\sin\frac{\pi S}{a}\left[\frac{\sin 2\pi(S-1)\cos\frac{\pi}{a}(S-1)}{\sin\frac{\pi}{a}(S-1)} - \frac{\sin 2\pi(S+1)\cos\frac{\pi}{a}(S+1)}{\sin\frac{\pi}{a}(S+1)}\right].$$

The complete Fourier series becomes

$$I(t) = \sum_{S=1}^{\infty} B_S \sin S\omega_S t.$$

It is interesting to note that all the harmonics are zero between the fundamental and the $(S-1)$th harmonic. The Fourier coefficients can be easily computed in the following form:

for $m=1$

$$b_1 = m_f\frac{\sin\frac{\pi}{a}}{\frac{\pi}{a}};$$

for $an+1$, where n is any integer greater than zero,

$$b_{an+1} = (-1)^n\frac{\sin\frac{\pi}{a}(an+1)}{\frac{\pi}{a}(an+1)};$$

for $m=an-1$

$$b_{an-1} = (-1)^{2n-1}\frac{\sin\frac{\pi}{a}(an-1)}{\frac{\pi}{a}(an-1)}.$$

Appendix II

Derivation of Superregenerator Bandwidth Using Ideal Control Wave Form

Integrating (9),

$$Z_P = \frac{\epsilon^{-1/2P_R t_R}}{2jC}\left[\frac{\epsilon^{-1/2P_S t_S}\epsilon^{j\sigma t_Y}}{\frac{1}{2}P_Y + j\sigma} + \frac{\epsilon^{j\sigma t_Y}(\epsilon^{j\sigma t_S} - \epsilon^{-1/2P_S t_S})}{\frac{1}{2}P_S + j\sigma} - \frac{\epsilon^{j\sigma(t_Y+t_S)}}{\frac{1}{2}P_R + j\sigma}\right]$$

where

$$\sigma = \omega_C - \omega_0.$$

Assuming $P_Y \to \infty$, $t_Y \to 0$, $P_R \gg 2\sigma$,

$$Z_P \cong \frac{\epsilon^{-1/2P_R t_R}}{jC}\left[\frac{\left(1-\frac{P_S}{P_R}\right)\epsilon^{j\sigma t_S} - \epsilon^{-1/2P_S t_S}}{P_S + j2\sigma}\right].$$

For reasonably narrow bandwidths $P_S \ll P_R$,

$$Z_P = \frac{\epsilon^{-1/2P_R t_R}}{jC}\left[\frac{\epsilon^{j\sigma t_S} - \epsilon^{-1/2P_S t_S}}{P_S + j2\sigma}\right].$$

Considering the bandwidth at the 3-db points, the expression for the static bandwidth $\Delta\omega_S$ becomes

$$\Delta\omega_S = \frac{P_S}{\sinh\frac{1}{2}P_S t_S}\left[\cosh^2\tfrac{1}{2}P_S t_S - \cos\frac{\Delta\omega_S t_S}{2}\right]^{1/2}.$$

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Some Slow-Wave Structures for Traveling-Wave Tubes*

LESTER M. FIELD†

Summary—A comparison of four slow-wave propagating structures for traveling-wave tubes is made. Recent experimental performance of several types of tubes using two of these circuits is given. Amplifier performance of helix tubes at 10,000 Mc, and wide-tuned oscillator performance (1.5 to 1) with second-harmonic output around 20,000 Mc are described. The analysis, construction, and performance of a 10,000-Mc disk-loaded-rod tube are presented. Apertured-disk and spiraled-waveguide structures are discussed briefly and compared with the helix and disk-loaded rod.

Introduction

This paper considers four representative types of structures for propagating slow waves in traveling-wave amplifier tubes.[1,2] Some recent applications of two of the structures in operating vacuum tubes are also described. A comparison of the gain properties of all of the structures described is generalized to some extent, and may serve as a guide to a search for other useful forms of structures.

The properties of these structures considered here are the variation of wave velocity with frequency and the ability to provide a high gain per unit length, allowing reasonable dimensions for the electron stream. The structures are compared to each other on these grounds, with additional consideration being given their power-dissipating properties.

The ability to provide a high gain per unit length may be expressed as a structure property determinable in the absence of an electron stream by[3]

$$\left[\frac{E^2}{\beta^2 P}\right]^{1/3}$$

where

E = the longitudinal field in volts per meter in the region where the electron stream would flow

$\beta = \omega/v_p$ where v_p is the wave phase velocity in meters per second and ω is 2π times the frequency

P = the power in watts being transmitted by the helix to support the field E.

$[E^2/\beta^2 P]^{1/3}$, it will be observed, has the dimensions of the one-third power of impedance, and enters into the gain of a traveling-wave tube very directly, since the amplification at synchronous velocity for a tube with no attenuation is

$$\text{voltage amplification} = 1/3e^{0.866\beta CL} \tag{1}$$

$$\text{or gain in db} = -9.54 + 47.3CN \tag{2}$$

where

$$C = \left[\frac{E^2}{\beta^2 P}\right]^{1/3}\left[\frac{I_0}{8V_0}\right]^{1/3}$$

L = the length of structure in meters

N = the number of cycles to traverse the structure at average electron velocity

I_0 = the dc beam current in amperes interacting usefully

V_0 = the voltage in volts specifying the beam velocity.

A wave velocity varying rapidly as a function of frequency will limit the bandwidth over which the tube can provide amplification at a fixed beam velocity, inasmuch as gain depends upon relative beam and wave velocities. The amount of gain variation resulting from a given velocity variation depends on the gain parameter C as described by J. R. Pierce[3] or by Chu and Jackson.[4]

I. The Helix

Analyses of the helix have been published in the literature[3,4] and hence need not be repeated here. Only the results significant for comparison with other circuits are repeated. The helix may be considered a standard of comparison, since it is simply constructed, has relatively high gain, and has such wide-band characteristics as compared with other circuits that it is probable that other circuits will be used only if they excel the helix in some significant property.

For the helix,[3]

$$\left[\frac{E^2}{\beta^2 P}\right]^{1/3} = \left(\frac{\beta}{\beta_0}\right)^{1/3}\left(\frac{\gamma}{\beta}\right)^{4/3} F(\gamma a) \tag{3}$$

where

$\beta_0 = \omega/c$

a = the turn radius of the helix in meters

c = the velocity of light in free space (3×10^8 meters per second)

$$\gamma^2 = \beta^2 - \beta_0^2 = \left(\frac{\omega}{v_p}\right)^2 - \left(\frac{\omega}{c}\right)^2$$

$F(\gamma a)$ = the gain function whose value for a helix is given versus γa by one of the curves of Fig. 1.

It is found that the velocity versus frequency characteristic for the helix is constant except at relatively

* Decimal classification: R339.2. Original manuscript received by the Institute, February 9, 1948; revised manuscript received, July 21, 1948.

† Stanford University, Stanford, Calif.

[1] R. Kompfner, "The traveling-wave tube as amplifier at microwaves," Proc. I.R.E., vol. 35, pp. 124–127; February, 1947.

[2] J. R. Pierce and L. M. Field, "Traveling-wave tubes," Proc. I.R.E., vol. 35, pp. 108–111; February, 1947.

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

[4] L. J. Chu and D. Jackson, "Field theory of traveling-wave tubes," Technical Report no. 38, Research Laboratory of Electronics, M.I.T., April, 1947. Also, Proc. I.R.E., vol. 36, pp. 853–863; July, 1948.

ow frequencies. This is in marked contrast to the other structures to be described.

That the helix is capable of excellent performance has now been demonstrated experimentally in several laboratories. Typical results include 23 db gain over a band 800 Mc wide centered at 4,000 Mc, reported from Bell

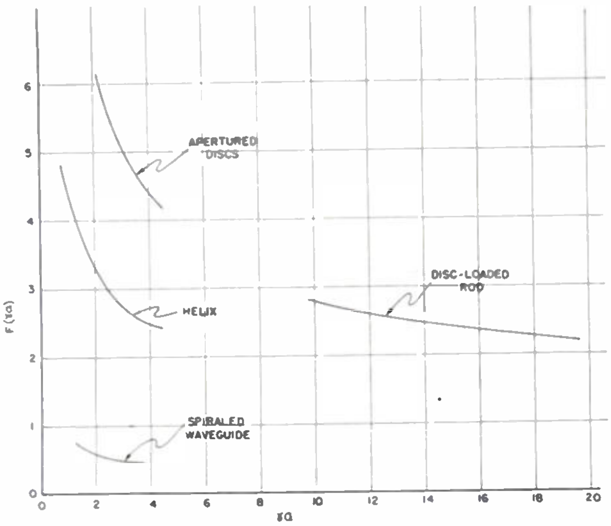


Fig. 1—Gain function F (proportional to gain in db per unit length) for various structures based on field at fin *edges* or at helix *surface*.

Telephone Laboratories, and similar results obtained at Stanford University, 13-db gain over a band 2,000 Mc wide at 9,000 Mc at Stanford, and production of oscillations of the order of 10 mw power output at 20,000 Mc at Stanford. Initial 20,000-Mc oscillator tubes were made without the very lossy attenuation region near the helix center that had been used to suppress oscillation in useful amplifier tubes at 9,000 and 4,000 Mc. Such attenuating regions are visible in the tubes of Fig. 3.

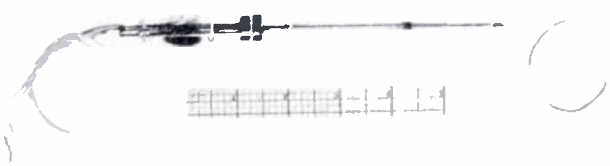


Fig. 2—20,000-Mc harmonic-output helix traveling-wave tube.

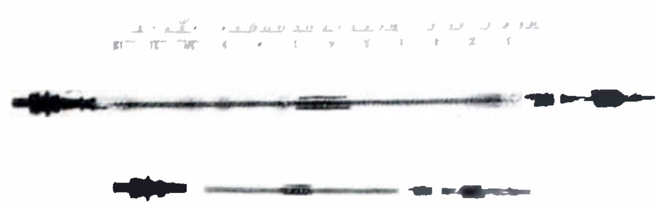


Fig. 3—4,000- and 9,000-Mc coaxially fed grid-helix tubes.

A tube for 20,000 Mc is shown in Fig. 2. It differs from earlier tubes for lower frequencies in its use of a relatively large helix, having a circumference for one turn approximately equal to one-half wavelength at 1.5 cm. Tubes of this type have been made for both 2,500 volts and 1,200 volts with 3 to 20 ma beam current. Oscillation frequency has been tuned continuously from 15,000 to 23,500 Mc by tuning a feedback circuit from 7,500 to 11,750 Mc and taking off several milliwatts of second-harmonic power through a waveguide cut off for the fundamental.

The helix may also be used as an amplifying section following a coaxially fed grid over the cathode surface, as illustrated by the tubes shown in Fig. 3. This idea, first suggested at Bell Telephone Laboratories, appears to be useful for obtaining a low-noise-figure amplifier, and with a high-transconductance grid structure can produce more gain than a helix-fed traveling-wave amplifier. With the signal applied to a grid, the bandwidth of the amplifier is usually limited by the bandwidth of the tuned grid circuit of the input, since the output circuit is a helix matched over at least a 20 per cent band. However, the bandwidth of this grid circuit may be several hundred megacycles at 3,000 Mc rather than the tens of megacycles available in an equivalent triode, because the Q of the transit-time-loaded grounded-grid input circuit will be the order of 10, rather than of the order of several hundred required for an output load for a triode at this frequency.

An analysis carried out at Stanford indicates that the gain may be many times that of the triode or helix portions separately, and also indicates that the noise figure of the combination tube will be the same as that of a triode of the same input impedance, beam current, and transconductance.

This analysis was carried out assuming that a signal applied to a grid over the cathode surface would produce ac beam current, but negligible velocity modulation and initial signal voltage on the helix in so far as excitation of a traveling-wave tube is concerned. These initial beam conditions, applied to the three forward waves of the Pierce traveling-wave-tube analysis,[3] give rise to the following relations for noise figure and gain:

$$\text{noise figure} = 1 + \frac{20\Gamma^2 I_0}{g_m{}^2 Rg} \tag{4}$$

where

Γ^2 = the space-charge noise-reduction factor
I_0 = the beam current in amperes
g_m = the transconductance of the grid portion of the tubes in mhos
Rg = the input resistance of the tube at resonance, including transit-time loss, measured in ohms;

$$\text{gain in db} = 10 \log_{10}\left[\frac{Rg}{\dfrac{E^2}{\beta^2 P}}\right] + 20 \log_{10}\left[2/3\,\frac{V_0 C^2 g_m}{I_0}\right] + 47.3CN. \tag{5}$$

This may be compared with (2) for the helix tube, and for practical values of the other parameters predicts equal gain for the two types of tubes at 4,000 Mc with a g_m of approximately 8,500 micromhos. However, the noise figure of the combination should be that of the triode. With a g_m of 17,000 micromhos, R_g of 50 ohms, and C of 0.034, the combination theoretically gives a 29-db higher gain than the helix alone. Since some measured noise figures for very high g_m (order of 50,000 μmhos) close-spaced triodes at 4,000 Mc are of the order of 12 to 15 db, as compared with approximately 25- to 30-db measured noise figures for present helix tubes, the combination tube may prove useful.

Experimental tubes of this type have been built at Stanford, and with a transconductance of 10,000 micromhos have given equal gains of approximately 30 db at 3,000 Mc, when fed either on the grid or on the helix, essentially as predicted by the analysis. Fig. 3, previously discussed in connection with center-loss oscillation suppression, shows two combination grid-helix tubes.

II. The Disk-Loaded Rod

This structure, shown in Figs. 4 and 6, is analyzed here in some detail because its analysis is typical of the

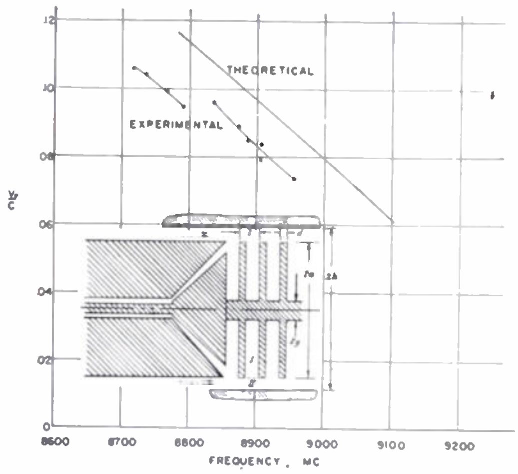


Fig. 4—Disk-loaded-rod structure and velocity versus frequency plot for dimensions given in Section II.

type of analysis applied to the other configurations described in this paper, and because it was selected as among the most promising of the circuits other than the helix for the following reasons:

1. The disk-loaded-rod structure requires the electron stream to be small in only one dimension (radial width), rather than requiring a small beam cross-sectional area as does the apertured-disk or helix structure. This permits the use of a much higher total current at comparable beam density and gain relative to the apertured disk or helix. This high current at a low density is useful in providing a tube which can be driven to high power levels before nonlinear amplification sets in. The high current would be bad for noise figure, but this is of relatively little importance in a high-power-level stage of amplification.

2. Although the disk-loaded rod is a relatively narrow-band structure (like the apertured-disk structure of Section III), it is wide enough in bandwidth to be of some value and, unlike the apertured disk, has an additional parameter ζ, capable of simple adjustment, which can considerably improve the bandwidth. (See Fig. 5.)

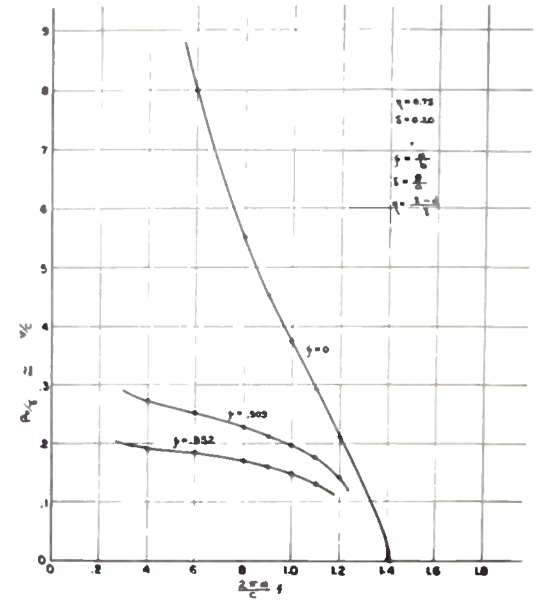


Fig. 5—Velocity versus frequency for disk-loaded-rod (or finned-center-conductor) structure. $\eta=0.75$, $\delta=0.20$, $\zeta=(a/b)$, $\delta=(g/a)$, $\eta=(l-d/l)$.

3. The large beam current, in addition to being a step in the direction necessary for higher limiting-power output, is permitted in view of the relatively high power capable of being dissipated by the disk-on-rod structure, especially as compared with a helix. In addition, the structure may be relatively sturdy and easy to make in small sizes, as compared with the helix or apertured-disk structure.

4. The calculated gain parameter is sufficiently high that a favorable comparison can be made with a helix tube for the same frequency, as shown in Table I. No

TABLE I

	Helix Tube	Disk on Rod
Frequency	9,000 Mc	9,000 Mc
Helix or disk diameter	0.090 inch	0.686 inch
Wire or disk thickness	0.010 inch	0.0113 inch
Voltage	1700 volts	2180 volts
Beam current	10 ma	100 ma
Helix or disk length	6.3 inches	6.3 inches
Theoretical gain	56 db	88 db
Limiting power output (theoretical)	0.5 watt	8.7 watts
Beam density	244 ma/cm²	135 ma/cm²

corrections for possible helix or disk-on-rod attenuation are included. The helix-tube gain figure is based on an averaged longitudinal field over the beam area. The gain figure for the disk-loaded rod assumes that the beam flows in a ring between 0.010 and 0.035 inch from the disk edges, as is described following the analysis of this structure. The structure being discussed is shown in Fig. 6.

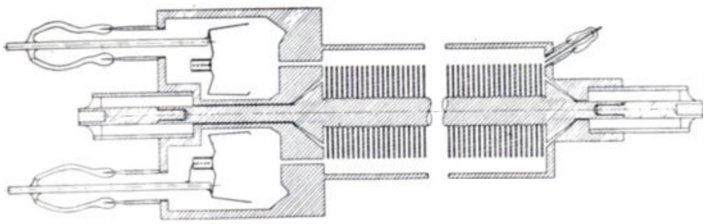


Fig. 6—Cross section showing disk-loaded-rod tube construction.

Analysis of the Disk-Loaded-Rod Structure

Solution for the velocity and gain properties of the disk-loaded rod in terms of dimensions may be carried out by setting up fields which are known solutions of the wave equation in appropriate regions of the structure. Solutions are chosen which fit the boundary conditions imposed by the structure. Then tangential electric and magnetic fields from each solution are set equal at the boundary between regions. This operation gives rise to a transcendental relation to be solved for velocity as a function of dimensions and frequency.

In region I of Fig. 4, $(g<r<a)$, assume a TEM wave propagating radially (a standing wave) which gives E normal to the side walls and zero at the bottom of the slots.

$$E_z = E_1[J_0(\beta_0 r)N_0(\beta_0 g) - N_0(\beta_0 r)J_0(\beta_0 g)]e^{j\omega t} \tag{6}$$

$$E_r = 0 \tag{7}$$

$$B_\phi = \frac{1}{j\omega}\frac{\partial E_z}{\partial r}$$
$$= j\frac{E_1}{c}[J_1(\beta_0 r)\cdot N_0(\beta_0 g) - N_1(\beta_0 r)\cdot J_0(\beta_0 g)]e^{j\omega t}. \tag{8}$$

In region II, $(a<r<b)$, assume a TM wave with no ϕ variation propagating in the Z direction, with velocity v_p where $v_p=\omega/\beta$.

$$E_z = E_0[I_0(\gamma r)K_0(\gamma b) - K_0(\gamma r)I_0(\gamma b)]e^{j(\omega t-\beta z)} \tag{9}$$

$$E_r = \frac{1}{\gamma^2}\left[j\beta\frac{\partial E_z}{\partial r}\right] = \frac{j\beta E_0}{\gamma}[I_1(\gamma r)K_0(\gamma b)$$
$$+ K_1(\gamma r)I_0(\gamma b)]e^{j(\omega t-\beta z)} \tag{10}$$

$$B_\phi = \frac{1}{\gamma^2}\left[j\omega\mu\epsilon\frac{\partial E_z}{\partial r}\right] = j\frac{\beta_0}{\gamma c}E_0[I_1(\gamma r)K_0(\gamma b)$$
$$+ K_1(\gamma r)I_0(\gamma b)]e^{j(\omega t-\beta z)}. \tag{11}$$

E_r and B_ϕ are found from E_z from the general relations between field components in cylindrical co-ordinates.[5] Since, in region II, the phase velocity is less than the velocity of light, the solutions are Bessel functions of imaginary argument written here as I_0 and K_0, the appropriate modified Bessel functions of real arguments. Again, $\gamma^2=\beta^2-\beta_0^2$.

Now, at $r=a$ we require continuous electric and magnetic fields, so that

$$\frac{E_{zI}}{B_{\phi I}} = \frac{E_{zII}}{B_{\phi II}}. \tag{12}$$

Since Maxwell's equations are linear, this ratio alone is sufficient to insure that

$$E_{zI} = E_{zII} \quad \text{and} \quad B_{\phi I} = B_{\phi II}$$

separately, which gives, on substituting (6), (8), (9), and (11) in (12), the following transcendental equation to be solved to give velocity as a function of frequency by giving γ for any prescribed β_0 and dimensions.

$$\eta\frac{J_0(\beta_0 a)N_0(\beta_0 g) - N_0(\beta_0 a)J_0(\beta_0 g)}{J_1(\beta_0 a)N_0(\beta_0 g) - N_1(\beta_0 a)J_0(\beta_0 g)}$$
$$= \frac{\gamma}{\beta_0}\frac{I_0(\gamma a)K_0(\gamma b) - K_0(\gamma a)I_0(\gamma b)}{I_1(\gamma a)K_0(\gamma b) + K_1(\gamma a)I_0(\gamma b)} \tag{13}$$

where $\eta=(l-d)/l$ is the space factor required to give equal line integrals of electric field in the two regions.

The solution of this relation has been carried out for several values of the parameters, and, from γ as a function of β_0, phase velocity versus frequency is plotted on Fig. 4 for the dimensions of Table I, and is plotted on Fig. 5 for generalized parameters.

It will be observed that no attempt has been made to match E_r on the two sides of the boundary. This should be done for a more accurate solution than the one given here, and will require the use of higher-order modes in both regions. The correction will be relatively small, as may be seen from the comparison of measured and theoretical velocity versus frequency (see Fig. 4). This measured velocity was taken as the electron velocity which gave maximum gain at each frequency in the operating tube.

To determine the gain parameter, we need to know $[E_z^2/\beta^2 P]^{1/3}$. For a given value of E, the power transmitted is found from the relation

$$P = 1/2\, Re\int(E_r \times H_\phi^*)\cdot ds.$$

Since E_r and B_ϕ are known from (10) and (11), it is possible to carry out this integration directly. The result is

$$[E_a^2/\beta^2 P]^{1/3} = (\beta/\beta_0)^{1/3}(\gamma/\beta)^{4/3}F(\gamma a, \gamma b) \tag{14}$$

where E_a is the value of E_z at $r=a$, and $F(\gamma a, \gamma b)$ is the gain function whose value for b infinite is plotted versus γa in Fig. 1, marked disk-loaded rod.

For b infinite (outer conductor removed),

$$F(\gamma a) = \left[\frac{(\gamma a)^2}{240}\{K_0(\gamma a)K_2(\gamma a) - K_1^2(\gamma a)\}\right]^{-1/3}[K_0(\gamma a)]^{2/3}. \tag{15}$$

[5] S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, Inc., New York, N. Y., 1944; p. 327.

For the tube design shown in Fig. 6, with the dimensions and operating conditions of Table I, γa is 17.8, which gives an $F(\gamma a)$ of 2.28. Since the beam does not flow at the edge of the disk but at some distance away from it, as previously described, if limited to a hollow cylinder 0.025 inch thick and starting 0.010 inch from the edge, the average value of $(E^2/\beta^2 P)^{1/3}$ (obtained by integration under a K_0 curve) is 2.22, rather than 5.04 as given by $F(\gamma a)\cdot(\beta/\beta_0)^{1/3}(\gamma/\beta)^{4/3}$.

Tube gain, following Pierce's analysis,[3] is, then,

$$\text{gain} = -9.5 + 47.3(I_0/8V_0)^{1/3}(2.22)N \text{ db}$$

which, at a beam current of 100 ma, beam voltage of 2,180 volts, and $N=52$, predicts 88.3 db gain, assuming no cold attenuation in the disk structure.

The limiting power output expected under these conditions is approximately CI_0V_0, where $C=[(E^2/\beta^2P)(I_0/8V_0)]^{1/3}=0.0397$, which is 8.65 watts for this structure.

The 3-db-down bandwidth is determined by velocity variation such that the band limits are reached when the velocity falls or rises 2.5 per cent. From Fig. 4, this will occur at frequencies 25 Mc apart.

The low current density for relatively high power-output level required for the disk-loaded-rod tube is of importance for two reasons. First, it is a reasonable density to draw from an oxide-coated cathode, whereas the density required for a 100-ma beam in the corresponding helix tube would not be. Second, for a hollow cylindrical beam whose radial thickness is small compared to the radius of the cylinder, space-charge repulsion effects depend upon density, rather than total current, as is the case for cylindrical beams.[6] Hence, the low density is useful and gives a much smaller beam spread resulting from space-charge repulsion than would occur in the solid cylindrical beam of a comparable helix tube at the same current and velocity.

Measured Performance of a Disk-Loaded Rod Tube

A tube of the design shown in Fig. 6 has been constructed, and photographs of it are shown in Figs. 7 and 8. Tube performance confirmed the velocity versus frequency predictions to the extent shown on Fig. 4.

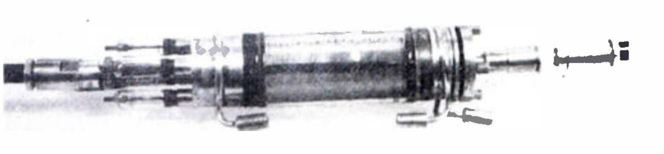


Fig. 7—Disk-loaded-rod tube with coaxial input and output cables.

The tube consists of a disk-loaded rod connected to coaxial lines at input and output by conical quarter-wave transformers. At the input end, the coaxial line passes through the center of the ring-shaped cathode. The cathode, designed with Pierce electrodes,[7] produces a hollow cylindrical electron beam which passes through a circular gap in the shaped anodes which is 0.150 inch long axially and 0.030 inch wide radially. The beam emerges from this gap just surrounding the edges of the disks. Successful operation of this gun is indicated by typical performance at 2,500 volts, as follows:

Emitted current from cathode	100 ma
Current to anode and tungsten screen shielding glass	25 ma
Current to disks	20 ma
Current to collector at output end of tube	55 ma.

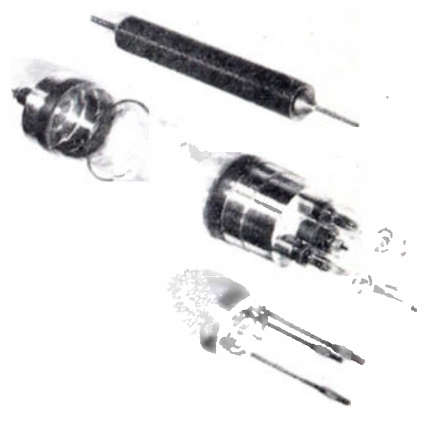


Fig. 8—Cathode structure, envelope, and disk-loaded rod.

No magnetic field was required to obtain these results, but any beam deflection by a magnetic field greatly reduced collector current and increased disk current.

Signal transmission by the disk structure in the absence of an electron beam was of the order of 40 db loss when only a slow mode was excited. This attenuation was far higher than had been desired, and persisted despite reasonably nonreflective matching from the coaxial feed to the disk structure. The effect of this attenuation is to produce complete absence of internal oscillation, appreciable sensitivity of gain to disk temperature, and lower gain than expected. The order of 5 db net gain or 45 db change in output signal level at 10,000 Mc has been observed. Linear amplification up to 1.5 watts output power was measured with no sign of compression. No noise-figure measurements were attempted in view of the high beam current and consequent power-amplifier application of this type of tube.

It is anticipated that much lower-attenuation disk

[6] D. P. R. Petrie, "The effect of space charge on electron beams," *Elec. Commun.*, vol. 20, pp. 100–111; 1941.

[7] J. R. Pierce, "Rectilinear electron flow in beams," *Jour. Appl. Phys.*, vol. 11, pp. 548–554; August, 1940.

structures can be made, since the observed loss resulted from a low-Q structure rather than from radiation, which is negligible on this structure. With lowered attenuation, subject to the limit imposed by self-oscillation, tube performance should improve correspondingly. In addition, similar tubes built for lower frequencies may have appreciably scaled-up power output.

III. The Apertured-Disk Structure

The analysis of this structure, which is an internally finned, loaded, cylindrical waveguide as shown in Fig. 9, has been carried out previously in the range of velocities useful for linear accelerators,[8] but is repeated here for completeness of presentation, using the same notation as for other traveling-wave-tube circuits.

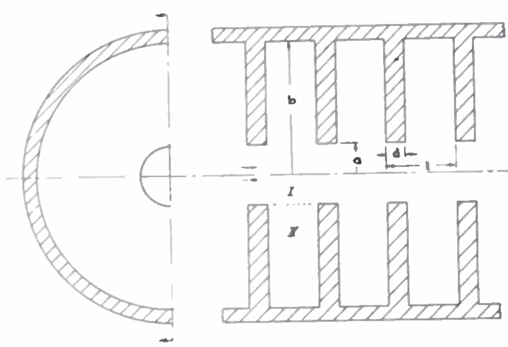


Fig. 9—Apertured-disk structure.

Define region I from $r=0$ to $r=a$ and region II from $r=a$ to $r=b$. In region I a TM-type wave will be assumed to exist for axial propagation. Then,

$$E_z = E_0 I_0(\gamma r) e^{j(\omega t - \beta z)} \tag{16}$$

$$E_r = jE_0 \frac{\beta}{\gamma} I_1(\gamma r) e^{j(\omega t - \beta z)} \tag{17}$$

$$B_\phi = \frac{jE_0\beta_0}{c\gamma} I_1(\gamma r) e^{j(\omega t - \beta z)}. \tag{18}$$

In region II, radial propagation by a TEM-type wave is assumed. Thus,

$$E_z = E_0[J_0(\beta_0 r)N_0(\beta_0 b) - N_0(\beta_0 r)J_0(\beta_0 b)]e^{j(\omega t)} \tag{19}$$

$$E_r = 0 \tag{20}$$

$$B_\phi = \frac{jE_0}{c}[J_1(\beta_0 r)N_0(\beta_0 b) - N_1(\beta_0 r)J_0(\beta_0 b)]e^{j(\omega t)}. \tag{21}$$

Here we have satisfied the boundary condition of zero electric field tangent to the conductor at $r=b$. The other boundary at $r=a$ requires continuous electric and magnetic fields, so we have

$$\frac{B_{\phi I}}{E_{zI}} = \frac{B_{\phi II}}{E_{zII}}. \tag{22}$$

[8] E. L. Chu and W. W. Hansen, "The theory of disk-loaded wave guides," *Jour. Appl. Phys.*, vol. 18, pp. 996–1008; November, 1947.

Substituting the above expressions into (22), we get, at $r=a$,

$$\frac{1}{\eta}\frac{J_1(\beta_0 a)N_0(\beta_0 b) - N_1(\beta_0 a)J_0(\beta_0 b)}{J_0(\beta_0 a)N_0(\beta_0 b) - N_0(\beta_0 a)J_0(\beta_0 b)} = \frac{\beta_0}{\gamma}\frac{I_1(\gamma a)}{I_0(\gamma a)} \tag{23}$$

where $\eta = (l-d)/l$ is the space factor. This is necessary to give equal line integrals of electric field in the two regions.

This transcendental equation may be solved for γ as a function of frequency for any set of dimensions, and therefore velocity versus frequency will be known. Typical curves plotted from such a solution are of the same general form as the $\zeta=0$ curves of Fig. 5, but have a low-frequency cutoff also.

The value of $E^2/\beta^2 P$ may be found just as described in Section III by evaluating P as $1/2\ Re\int(E\times H^*)\cdot ds$ over the cross section. A result for $E^2/\beta^2 P$ for this structure, where E is the field on the axis, is

$$(E^2/\beta^2 P)^{1/3} = (\beta/\beta_0)^{1/3}(\gamma/\beta)^{4/3}F(\gamma a). \tag{24}$$

$F(\gamma a)$ is plotted in Fig. 1 against the parameter γa where a is the hole radius. For a given hole through which a beam may be sent, and at the same velocity, it is evident that the apertured-disk structure should have approximately twice the number of db gain per unit length that a helix would have. However, the apertured-disk structure would make a rather narrow-band amplifier. As an example, with a gain parameter C of 0.03, which is typical, the apertured-disk circuit would be approximately 25 Mc wide at 9,000 Mc between 3-db-down points. The helix would not be limited by velocity variation, but rather is now limited to about 2,000 Mc bandwidth at 9,000 Mc by matching problems.

IV. The Helical Waveguide

It is evident that the structures of Sections II and III of this paper could be modified slightly, such that adjacent cavities are connected in a continuous spiral fashion. Fig. 10 shows the resulting configuration when this

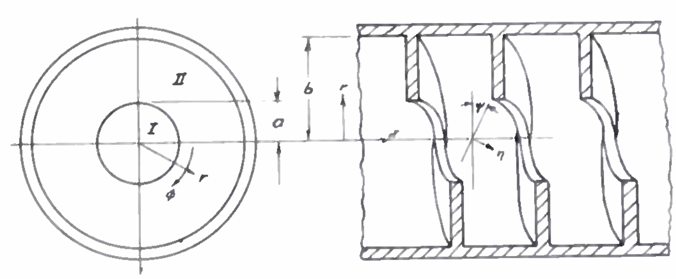


Fig. 10—Internal helical-waveguide structure.

process is applied to the apertured-disk structure of Section III. It might be hoped that much broader-band properties might result.

Preliminary attempts at analysis of velocity versus frequency for such a structure indicate that this is indeed likely, but the analysis to date involves approximations not valid for practical dimensions, and is not included here. However, one evident consideration makes the use of such a structure of doubtful value as compared with a helix. In analyzing the helical waveguide, the field configuration inside the hole could very well be represented by TM and TE waves of the same form, $I_0(\gamma r)$, as those used inside a helix. Outside the hole, in a helix tube the fields die off as $K_0(\gamma r)$, but in the helical waveguide the fields die off much more slowly, at least for dimensions relative to a wavelength at which the field in the half-waveguide looks like a normal waveguide field variation—that is, far from cutoff. Hence, for a given electric field on the axis, the helical waveguide will be transmitting much more power, and hence will have a much lower gain parameter. Initial results indicate that, as an extreme value, for equal aperture sizes the gain in db per unit length for the helical guide might be as low as one-sixth of that for a helix as shown for a typical calculation in Fig. 1. This result should be considered as merely suggestive because of the broad approximations used in deriving it. Near cutoff, gain and velocity-dispersion properties should approach those of the apertured-disk structure of Section III.

V. Comparison of Structures

It is instructive to compare gain for a helix, apertured disk, and spiraled waveguide, all with the same-sized aperture through which a beam may be passed. That the spiralled waveguide should have a lower gain parameter than a helix seems reasonable from the arguments of Section IV.

On the same basis, the apertured-disk structure should be better than a helix, since it transmits no power except through the apertures (assuming small cavity power losses). Since, over most of the useful operating range of Fig. 1, the helix transmits approximately equal amounts of power in the fields inside and outside of the helical surface, one might expect the apertured-disk structure to be better by a factor of two in power transmitted for a given axial field. Inasmuch as gain per unit length varies inversely as the one-third power of the total power transmitted for a unit axial field, this would seem to predict that the apertured-disk structure was better by a factor of $2^{1/3}$, or 1.26, whereas the curves of Fig. 1 show an improvement of 2 to 3 times.

The discrepancy is resolved when it is found that the extra power carried by the TE fields in the helix accounts for the difference exactly. These fields are necessary to meet the skew boundary of a helix or helical waveguide, but are not necessary in the apertured disk or disk-loaded rod.

Direct comparison of the disk-loaded rod to the other structures is difficult because of the different electron-beam configuration, but the beam densities required for other structures to approach equivalent gain as described in Section II are evidence of its usefully high gain properties. It would appear from the curves of Fig. 1 that the disk-loaded-rod structure would be useful for high-beam-current, high-power-output tubes, while the apertured-disk structure would be preferable for high-gain, low-limiting-power applications.

Since the improved gain properties of the apertured disk and disk-loaded rod as compared with the helix very largely result from the absence of skewed boundaries in these structures, the possibility of decreasing the dispersion to get wider-band amplification by providing coupling from cavity to cavity, in addition to that provided by fringing fields, would seem to have merit. The extra power transferred by coupling holes in the fins, or by some other coupling scheme, would raise the group velocity, and so decrease dispersion. However, this extra power transfer would need to be kept low enough to avoid using up all of the gain advantage obtained from the absence of skewed-structure boundaries.

Acknowledgment

The writer is indebted to D. A. Dunn and A. R. Margolin, graduate students at Stanford University, for completing integrations and computations of Sections II and III. The work described was carried out under the sponsorship of the Office of Naval Research.

Correction

The following errors in the paper, "Some Notes on Noise Figures," by Harold Goldberg, which appeared on pages 1205–1215, of the October, 1948, issue of the Proceedings of the I.R.E., have been brought to the attention of the editors by the author:

Page 1208, equation (17), should read

$$N = \frac{4KTBA^2}{4R_0}\left[R_{eq} + \frac{RR_1}{R + R_1}\right].$$

Page 1210, the line following equation (23) should read

"Recognizing the fact that $G_{m.q}$ is equal to $G_m \cdot G_{m+1} \cdots G_q$, one obtains . . . . "

Page 1210, equation (24), the last term should read

$$+ \frac{(F_n - 1)B_{n,n}}{G_{1,n-1}B_{1,n}}.$$

Page 1211, first column, the second last formula should read.

$$R_{eq} = \frac{R_{eq_p}}{\mu^2}\left[\frac{R_0 + R_p}{R_0 + R_{eq_p}}\right]^2.$$

A Theory on Radar Reflections from the Lower Atmosphere*

W. E. GORDON†, ASSOCIATE, IRE

Summary—A theory is presented, supported by several sample sets of data, which indicate that the curious phenomenon dubbed "angels" (radar reflections from the lower atmosphere) may be attributed to sharp changes in the dielectric constant. The required magnitudes of the changes are computed from reflection theory and compared to sample meteorological data obtained from rapid-response instruments. The near-discontinuities in the dielectric constant are produced by atmospheric turbulence. It is proposed that the observed radar reflections are the result of turbulent motion in the lower atmosphere.

Introduction

Curiosity was aroused when radar echoes from a clear sky were detected. Early observers suspected birds or stray weather balloons, but these were eliminated by visual checks. Radar reflections from the lower atmosphere, referred to as "angels," are detected on sensitive sets and are produced by sources invisible to the human eye. In the past few months certain measurements have been obtained which offer an explanation of the echoes. These measurements are meteorological in nature and were obtained from rapid-response instruments. This paper will propose that the radar reflections in the lower atmosphere are the result of atmospheric turbulence, which creates abrupt changes in the dielectric constant.

The existence of the echoes for microwave radars is confirmed by reports of Baldwin,[1,2] Friis,[3] and Gould.[4,5] Crawford and Sharpless[6] postulated in November, 1946, the existence of discontinuities of the dielectric constant in accounting for multiple-path transmissions. As early as 1940, Friend[7] attibuted radio echoes on a wave length of 123 meters to large gradients of dielectric constant in the lower atmosphere. Reference must be made to the remarkably similar reflections of sound signals,[8] which closely parallel the radar reflections.

This paper will summarize the microwave observations and present the pertinent properties of atmospheric turbulence, supplemented by data for the lower atmosphere obtained from rapid-response temperature and wind instruments. The material will be considered under three headings: first, the properties of the echoes as determined from radar observations; second, the turbulence measurements and their relation to the problem; and third, reflection coefficients in the atmosphere.

I. The Radar Observations

The reflections have been observed on 1-, 3-, and 10-cm radars at vertical incidence. The following are presented as a summary of the observed characteristics of the reflections received on a fixed antenna directed vertically upwards, and are based on conversations with Crawford[9] and on Gould's paper.[4,5]

1. The echoes are weak. For example, the power-reflection ratios are of the orders 10^{-9} to 10^{-12}. These limits are, of course, functions of the equipment, being determined by the saturation and sensitivity of the set used, but they serve to indicate orders of magnitude.

2. Angels are detected from the antenna height (at 15 feet) up to about 9,000 feet, but are most numerous below 3,000 feet.

3. The frequency of occurrence varies from no detectable echoes up to a number which cannot be counted by an observer, and the distribution with time is quite random.

4. The echo durations are from a fraction of a second to several seconds, perhaps ten. The duration is influenced by the beam size of the antenna and is not necessarily a property of the reflecting source. In fact, Baldwin[1] has tracked Angels for periods of 5 to 10 minutes, traveling with the wind.

5. The echoes appear in all seasons, and both day and night. However, they are more numerous during the summer and tend to occur with moderate winds. There is some evidence that the echoes stratify at heights corresponding to temperature inversions.

II. Atmospheric Turbulence

To radio engineers, remarks on atmospheric turbulence may appear as a digression to be regarded lightly, but this is a subject in which the meteorologists and radio engineers, especially those concerned with propagation, must have a mutual interest. Turbulence is an important basic factor in the study of microwave propagation in the atmosphere. For the present, the discussion is confined to an elementary description foregoing the mathematical treatment, except to note that at present a statistical approach is the most promising.

* Decimal classification: R537×R113.501.1. Original manuscript received by the Institute, April 15, 1948. Presented, 1948 IRE National Convention, New York, N. Y., March 25, 1948.

† Formerly, Electrical Engineering Research Laboratory, University of Texas, Austin, Tex.; now, School of Electrical Engineering, Cornell University, Ithaca, N. Y.

[1] M. W. Baldwin, Jr., "Radar echoes from the nearby atmosphere," Presented, joint URSI-IRE Meeting, Washington, D. C., May 2–4, 1946.

[2] M. W. Baldwin, Jr., "Radar echoes from the lower atmosphere," Correspondence, Proc. I.R.E., vol. 36, p. 363; March, 1948

[3] H T. Friis, "Radar reflections from the lower atmosphere," Correspondence, Proc. I.R.E., vol. 35, pp. 494–495; May, 1947.

[4] W. B. Gould, "Radar reflections from the lower atmosphere," Correspondence, Proc. I.R.E., vol. 35, p. 1105; October, 1947.

[5] H. B. Brooks, W. B. Gould, and R. Wexler, "What are angels?" Joint URSI-IRE Meeting, Washington, D. C., October 20–22, 1947.

[6] A. B. Crawford and W. M. Sharpless, "Further observations of the angle-of-arrival of microwaves," Proc. I.R.E., vol. 34, pp. 845–848; November, 1946.

[7] A. W. Friend, "Developments in meteorological soundings by radio waves," *Jour. Aero. Sci.*, vol. 7, pp. 347–352; June, 1940.

[8] G. W. Gilman, A. B. Coxhead, and F. H. Willis, "Reflection of sound signals in the troposphere," *Jour. Acous. Soc.*, vol. 18, pp. 274–283; October, 1946.

[9] A. B. Crawford, Bell Telephone Laboratories, Holmdel, N. J.

The rapid irregular fluctuations of wind velocity about a mean are indicative of the turbulent character of the atmosphere. Before the description may proceed very far, however, two words are encountered which form sort of a basic vocaublary for the study of turbulence. The two words, scale and intensity, will describe the turbulence. Simply, the scale is a measure of the average eddy size. The intensity is merely the ratio of the fluctuations to the mean wind speed. A few familiar examples of atmospheric turbulence are the gustiness of wind, the crisp cauliflower shape of active cumulus clouds, and the bumps encountered in air travel under certain conditions. These are readily observed because of their relatively large scales and intensities. With instruments of sufficiently small time lag, a virtually continuous spectrum of turbulence is shown to exist in the atmosphere to frequencies of 3 kc or higher. The intensity, of course, decreases rapidly with increasing frequency.

In general, turbulence is more pronounced near the ground, where its scale and intensity are determined by wind speeds, roughness of the surface, and thermal conditions. Turbulence is also associated with boundaries of different air masses aloft.

Less well known than the wind fluctuations are the very similar temperature fluctuations. Startling information is obtained with a suitable temperature-measuring element. Temperature fluctuations up to 6°C in periods of a fraction of a second to several seconds have been indicated.

As a point of interest, consider the suitable element. The element is a piece of platinum wire, finer than a human hair, and just a few millimeters in length. It is useful for both temperature and wind measurements. More precisely, it is a density-flow meter, but can be used to indicate either temperature or wind, depending on the current applied to it. When operated with a current which makes it glow, it becomes the conventional hot-wire anemometer or wind-measuring device; when operated cold, i.e., just sufficient current to obtain a resistance measurement, it becomes a resistance thermometer. Due to its small mass it can follow wind and/or temperature fluctuations, and is well suited to turbulence measurements.

Without going into the details of the bridges, galvanometers, and amplifiers necessary to obtain the indication of the fluctuations, consider how the elements are used. Single elements are used as hot-wire anemometers to measure the fluctuations in velocity. A pair of elements separated by 6 millimeters is used as a differential thermometer; that is, they indicate the temperature difference between the two elements. A separation of 6 millimeters is a length which is considerably smaller than a wavelength at 3 or 10 cm, and is approximately one-half wavelength at 1.25 cm. The data were taken up to heights of 300 feet above the ground. This is, admittedly, far short of the heights to which echoes are observed. However, theoretical considerations confirm that, while greater heights are desirable, the elements are, nevertheless, sampling the free atmosphere. In these two forms, the elements were used to obtain several sets of data. The measurements were made on five different days and represent approximately eight hours of observations. For heights above 100 feet, the maximum value of the temperature differential and intensity of turbulence were of the order of 1°C and 5 per cent, respectively.

In order to show the relationship of these data to the radar reflections, it is necessary to refer the hot-wire fluctuations and temperature differentials to corresponding values of dielectric constant. King[10] in his classical reference on hot-wire anemometry provides the basic equation (1) relating velocity (v) to density (ρ). The equation applies with certain limitations on the method of measurement and the magnitude of the fluctuations.

$$\frac{\Delta\rho}{\rho} = -\frac{\Delta V}{V}. \tag{1}$$

Using Debye's work,[11] the dielectric constant of air (ϵ), may be expressed in terms of the density of the dry air and the moisture concentration (m) as

$$\epsilon - 1 = a\rho + bm \tag{2}$$

where a is a constant and b is proportional to the reciprocal of temperature. Neglecting moisture for the present, and noting that the index of refraction n equals the square root of the dielectric constant, we may write

$$\frac{\Delta n}{n-1} = \frac{\Delta\rho}{\rho}. \tag{3}$$

Substituting from (1) and inserting a reasonable value for $n-1$, there results

$$\Delta n = -300\frac{\Delta V}{V}10^{-6}. \tag{4}$$

Note that (4) relates the change (or the magnitude of the fluctuation) in index to the *intensity* of turbulence. Corrsin[12] has described a method of measuring concentration fluctuations by means of hot wires. This appears to be an extremely useful method of obtaining the required data precisely, although, as yet, it has not been adapted to measurements in the atmosphere.

Again referring to Debye, express the index change in terms of the temperature differential (ΔT) and the moisture differential (Δe) as

$$\Delta n\ 10^6 = -1.4\Delta T + 4.0\Delta e. \tag{5}$$

This expression results from the substitution of certain average temperature and moisture values which are ap-

[10] L. V. King, "Air velocity by means of the hot wire anemometer," *Phil. Mag.*, vol. 29, pp. 556–577; 1915.

[11] P. Debye, "Polar Molecules," The Chemical Catalogue Publishing Company, 1929.

[12] S. Corrsin, "Extended applications of the hot-wire anemometer," *Rev. Sci. Instr.*, vol. 18, pp. 469–471; July, 1947.

propriate for the summer conditions under which the radar observations were made.

III. Reflection Coefficients in the Atmosphere

The power-reflection coefficient may be written in terms of the dielectric constants of media one and two as

$$R = \left(\frac{\sqrt{\epsilon_2} - \sqrt{\epsilon_1}}{\sqrt{\epsilon_2} + \sqrt{\epsilon_1}}\right)^2. \tag{6}$$

This equation applies to air, and may be immediately reduced to (7) using the relation of dielectric constant and index of refraction:

$$R = \left(\frac{\Delta m}{2}\right)^2. \tag{7}$$

The numerator is the difference or change in index between the two media. The value of the denominator of (6) for air is very closely two.

Equation (7) deserves a few comments. It represents the reflection coefficient for vertical incidence at the plane boundary of two media of infinite extent. The condition of a boundary is satisfied when the change occurs in a distance small compared to a wavelength. The requirement of infinite vertical extent is eliminated, since this equation also represents the upper limit of reflection for a dielectric sheet inserted in a homogeneous medium. Horizontal limits to our original infinite-extent assumption can be applied in terms of beam widths and Fresnel zones. An examination of reflection theory for more complicated models leads to similar results, and, in the light of our limited present knowledge of eddy shapes, a more complicated model is not warranted. We are thus led to accept (6) and (7) as a first approximation to the reflection coefficient for our problem.

Be reminded that the reflection coefficients as determined by the radio measurements on the basis of a flat reflecting surface were of the order of 10^{-9} to 10^{-12}, and that these values are functions of the particular equipment employed. The differential temperature and wind fluctuation data almost always produce reflection coefficients smaller than 10^{-12}; that is, they indicate changes in index of refraction which would create radar echoes weaker than those detected. However, on a few occasions reflection coefficients between 10^{-10} and 10^{-12} have been computed from meteorologically measured data. A maximum of four such items was recorded at a height of 250 feet in a single half-hour period. Normally the frequency of occurrence is much less. This is readily explained when it is considered that the elements are operated at a fixed point in the atmosphere. It follows that the chances are small of an eddy of the necessary characteristics flowing by the elements. There are, on a few occasions, differential-temperature and wind-fluctuation measurements which show that rapid changes in the index or dielectric constant do exist, and that these changes are sufficient to produce echoes equal in magnitude to the weaker echoes that have been detected by the radar sets.

The moisture changes and their contribution to index changes were neglected. This was done of necessity rather than by choice, for we simply did not have a moisture element with a response rapid enough for this type of work. A few general remarks on this subject may be made. From (5) it can be seen that the contribution of moisture change to the index change is very significant. The moisture change might conceivably add to the index change, tend to compensate for the temperature change giving practically no index change, or it might be the predominant term altogether. At the present there is no suitable way of establishing its significance for particular situations; however, the conclusion from the radio observations that the angels are more frequent in summer sheds some light. In summer the air is capable of holding more moisture, and hence there are potentially larger changes in index allowable. Very recently there have been reports of an instrument developed at the University of Chicago[13] which has measured rapid moisture changes in the atmosphere near the ground which correspond to as much as 20 to 30 units in $\Delta n \times 10^6$. One foot above warm ground there was observed a maximum temperature differential of 6°C, equivalent to approximately eight units in $\Delta n \times 10^6$. The suggestion is that this contribution must at times increase the index change, or become the predominant term.

At air-mass boundaries aloft, eddies are produced by the wind shear. With these eddies there are associated the physical properties of the two air masses involved. Hence, it is very reasonable that echoes have been observed at heights roughly corresponding to the height of air-mass boundaries aloft.

IV. Conclusion

The properties of the radar reflections have been reviewed, especially noting their weak magnitudes, their concentration in the lower atmosphere, their more or less random distribution with time, and their association with moderate winds and air-mass boundaries. There is a striking correspondence of these properties with the characteristics of atmospheric turbulence, and in particular we have measured differential temperatures and wind fluctuations which, when evaluated in terms of reflection coefficients, are capable of producing the weaker observed radar echoes. Of necessity, the moisture contributions were neglected, but radar observations lead us to believe that they reinforce the temperature contributions.

On the basis of the available evidence it is proposed that the radar reflections may be attributed to abrupt changes of dielectric constant, and are the result of the turbulent motion of atmosphere. Here, then, is a research tool which can be used to the mutual advantage of the meteorologist and radio engineer.

[13] Verner E. Suomi, Director, Instrument Laboratory, Department of Meterology, University of Chicago, Chicago, Ill.

Predicting Maximum Usable Frequency From Long-Distance Scatter*

ARTHUR H. BENNER†, STUDENT MEMBER, IRE

Summary—**A method of predicting the maximum usable frequency by means of observation of long-distance (2F) scatter is presented. The locations of the scattering sources are fixed by comparison of experimental observation with theoretical calculations.**

Introduction

THE KNOWLEDGE of the maximum usable frequency, muf, for a given distance at a given time is a vital factor in maintaining radio communication via sky-wave paths. The determination of the muf at the present time is usually made by reference to material published by the Central Radio Propagation Laboratories, National Bureau of Standards.[1] This material consists of charts of 0 and 4000-km muf's for three world zones. These charts are published each month, and each booklet contains muf predictions for that month, which is three months in advance of the publication date.

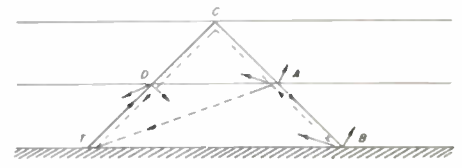
Although this method of predicting the muf is very useful, there are certain limitations in the accuracy. First, the predictions are made three months in advance and, therefore, are subject to errors due to unpredictable ionospheric changes in that period. Secondly, the world is divided into three zones in these reports, and a single chart covers an entire zone. This leads to errors in prediction, as the ionospheric variation within one zone may be of considerable magnitude.

It is the purpose of this paper to outline a method of predicting the muf that will minimize the two errors stated above. This method involves the observation of long-distance scatter.

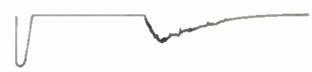
Description of Scatter

The scattering of electromagnetic waves in the frequency range of about 2 to 30 Mc may arise from ionospheric clouds in the E layer or from ground scattering. Eckersley[2] has denoted the different types of scattered echoes as short-distance E, $1F$, $2F$, and ground scatter. These types are illustrated in Fig. 1(a). It is these last two in which we are interested.

If a transmitter is sending out radio-frequency pulses at a carrier frequency higher than the vertical-incidence critical frequency of the F layer, we might expect one possible mode of transmission to be that as given by the solid line in Fig. 1(a). If, at the point A in the E layer, there exists some sharp irregularities in the ionization density, it is possible that some of the wave will be scattered. Part of this scattered energy may return to the transmission site by the dotted path, as shown in this figure. This is long-distance $2F$ scatter. If the wave



(a)



(b)

Fig. 1

(a) Scatter; TDT—short-distance E; $TCAT-1F$; $TCACT-2F$; $TCBCT$—ground.
(b) Typical oscilloscope pattern of scatter echo.

which arrives at the point B on the ground is scattered, part of the scattered signal may be returned to the transmitter by the same path; that is, ground, F, and back to the transmitting site. This is called ground scattering. The received echoes of these scattered signals as observed on a linear oscilloscope trace at the transmitting site will be similar to that in Fig. 1(b). In this figure, the pulse at the left edge of the trace is the transmitter pulse, while the echo in the middle is the received scatter echo.

In Fig. 2(a) is a plane-earth–plane-ionosphere geometric portrayal of the skip zone. If the point B is the edge of the skip zone, then the leading edge of the scatter pattern corresponds to the ray TAB. Hence the operating frequency f' is the maximum usable frequency for the distance D. Now, from the observed oscilloscope scatter pattern, we may observe the distance TAB or TAC, depending on the source of scatter. The problem, then, is to find a relation among f', TAB or TAC, and D.

Determination of Scatter Source Location

First, we will develop an expression to make a theoretical check of our observed data, and in this way attempt to fix the location of the scattering centers; that is, whether they are in the E layer or on the ground.

For developing an expression concerning this, curved-earth–plane-ionosphere geometry will be used. It has

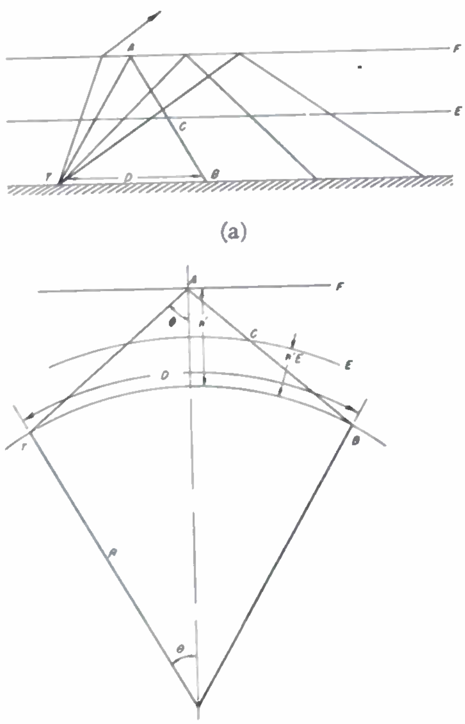
* Decimal classification: R113.616.3. Original manuscript received by the Institute, March 16, 1948; revised manuscript received, July 26, 1948.

† Pennsylvania State College, State College, Pa.

[1] "Basic Radio Propagation Predictions," CRPL-D Series, National Bureau of Standards, Government Printing Office, Washington, D. C.

[2] T. L. Eckersley, "Analysis of the effect of scattering in radio transmission," *Jour. I.E.E.* (London), vol. 86, pp. 548–563; June, 1940.

been shown by Smith[3] that curved-earth–plane-ionosphere geometry is valid as a second approximation in transmission-path calculations. We are justified in using this, since the data never exceeded a scatter distance of 2000 km. In this approximation, we consider the E layer as concentric with the earth, but ignore the effect of the curvature of the F layer. In a given experimental case, the data available from measurement include the height of the F layer, the F-region vertical-incidence critical frequencies, the radius of the earth (6320 km), the height of the E layer (assumed constant at 100 km), the frequency at which the scatter is observed, the distance to the leading edge of the $2F$ scatter group, and the distance to the maximum amplitude of the $2F$ scatter pattern.



(a)

(b)

Fig. 2

(a) Plane earth-plane ionosphere geometry of skip zone
(b) Curved earth geometry of scatter calculations.

Referring to Fig. 2(b), it is desired to find the distance TAB and TAC for transmission by the ordinary ray. From Martyn's theorem,[4] the minimum angle of incidence ϕ on the F layer for a wave of frequency f' may be determined from the relation

$$f' = f/\cos\phi$$

where f is the equivalent vertical-incidence frequency at the point of tangency of the transmission curve for a given distance with the $h'f$ curve. However, since the $h'f$ data was taken manually, it was possible only to record f_c, the vertical-incidence critical frequency of the F layer. In these first experiments, since the distances involved were relatively small, f_c was used in place of f in the above equation as an approximation. If the complete $h'f$ curves are available, f should be used. From the geometry of the figure,

$$R^2 = \overline{TA}^2 + (h' + R)^2 - 2\overline{TA}(h' + R)\cos\phi$$

where h' is the virtual height of reflection and R is the radius of the earth. Rearranging, substituting the cosine law for $\cos\phi$, and casting out the positive sign from physical reasoning, leads to

$$\overline{TA} = (h' + R)\frac{fc}{f'} - \left\{R^2\left(\frac{fc}{f'}\right)^2 + \left[\left(\frac{fc}{f'}\right)^2 - 1\right](2h'R + h'^2)\right\}^{1/2}.$$

By means of this equation, TA may be determined from known or measured quantities.

Similarly it can easily be shown that

$$\overline{AC} = (R + h')\frac{fc}{f'} - \left\{(R + h')^2\left[\left(\frac{fc}{f'}\right)^2 - 1\right] + (R + h'E)^2\right\}^{1/2}$$

where $h'E$ is the height of the E layer which is assumed known and constant.

The two desired relationships are

$$\overline{TAC} = \overline{TA} + \overline{AC} \quad (1)$$

$$\overline{TAB} = 2\overline{TA}. \quad (2)$$

Equipment

The equipment used consisted of a pulse transmitter yielding approximately 50-kw peak power rating using 100-microsecond pulses at a repetition rate of about 20 per second. This transmitter was part of $h'f$ equipment used for other experimental work. The receiver was a Hammarlund Super Pro modified for pulse reception. Observations of the delay time of the scatter were made either visually, or by photographing the signals displayed on an oscilloscope screen in conjunction with appropriate range marks. The antennas used for both transmitting and receiving were ordinary half-wave horizontal doublets one-quarter wavelength above ground. Although several antennas were used to cover the operating range, the individual antennas were used over a fairly large frequency range, so that the directivity characteristics are not accurately known.

It has been found that long-distance scatter is attenuated severely with increasing frequency. High system gain is necessary to observe scatter at a frequency considerably above the vertical-incidence critical frequency. With the apparatus used, readable echoes could be recorded with the equipment up to approximately twice the vertical-incidence critical frequency.

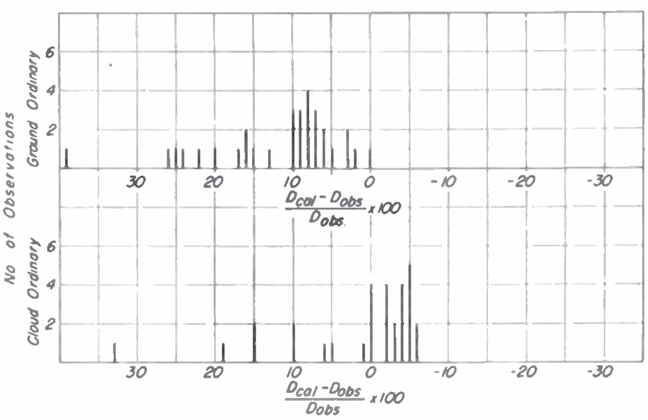
[3] N. Smith, "The relation of radio sky-wave transmission to ionosphere measurements," Proc. I.R.E., vol. 27, pp. 332–348; May, 1939.

[4] D. F. Martyn, "The propagation of the medium radio waves in the ionosphere," *Proc. Phys. Soc.* (London), vol. 47, p. 332; March, 1935.

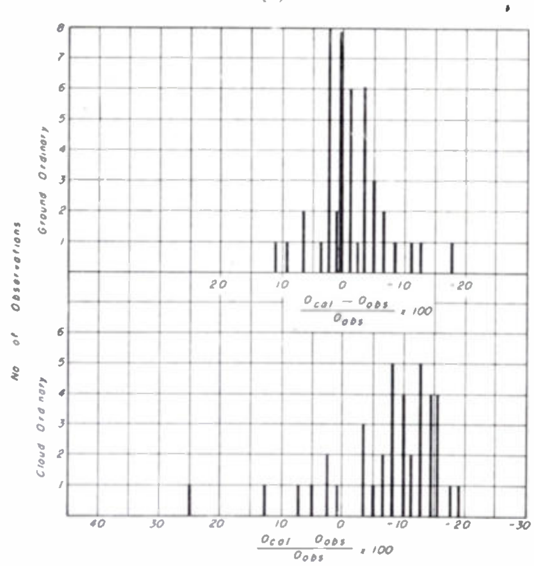
Experimental Data

From the $h'f$ data, the f^0F_2 and f^xF_2 critical frequencies were obtained along with the layer heights. The frequency of the observed scatter f' and delay to the scatter leading edge and maximum amplitude were recorded. TAB and TAC were calculated from (1) and (2) for both the ordinary and extraordinary rays. It was found that the ordinary-ray results checked much more closely than those for the extraordinary ray.

In Fig. 3(a) appears a frequency chart illustrating the number of observations versus the percentage deviation of the calculated theoretical values against the observed



(a)



(b)

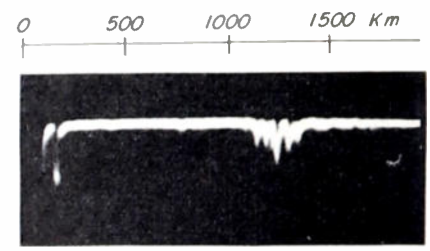
Fig. 3—Frequency diagrams.

(a) Measurements to the scatter-pattern leading edge.
(b) Measurements to the maximum amplitude of the scatter pattern.

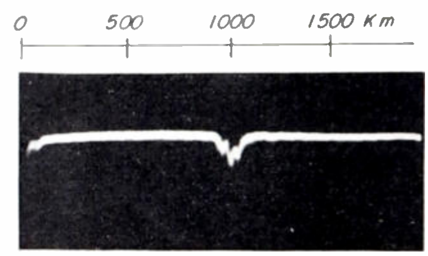
values to the leading edge of the scatter pattern. It is evident from this figure that, within the limits of the number of observations available, the leading edge of the scatter pattern is a result of scattering centered in the E region. This observation is based on the statistical centering of the frequency diagram for the cloud-scattering assumption and grouping of the observations in the ground-scattering frequency diagram at positive error percentage, indicating that D_{obs} is less than D_{cal}.

In Fig. 3(b) the data relating the location of scattering centers and the maximum amplitude of the scatter pattern are shown in a frequency diagram. From these charts it is apparent that the maximum of the scatter groups corresponds to the calculated ground-scattering distance. This statement is supported by noting that the center of the frequency distribution for the theoretical ground delay is closer to the zero per cent deviation point than is that for the theoretical cloud delay. Also, the frequency distribution for the cloud ordinary observations center on negative deviations, indicating that the observed values are greater than the theoretical distance to the assumed cloud sources.

These data were taken over a period of four months in the spring and summer of 1947. Observations were



(a)



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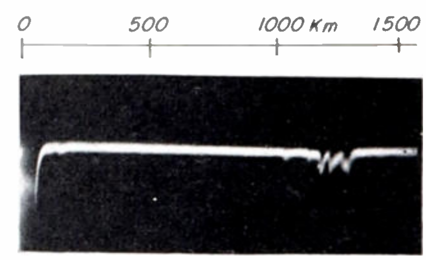
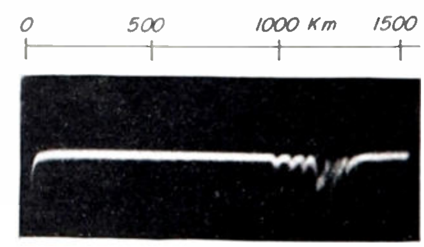


(c)



(d)

Fig. 4—Photographs of scatter signals.

(a) $\frac{f_c}{f'} = 0.84$; (b) $\frac{f_c}{f'} = 0.92$; (c) $\frac{f_c}{f'} = 0.87$; (d) $\frac{f_c}{f'} = 0.83$.

made during both the daytime and nighttime hours. Because of the decreased E-layer absorption, greater scatter distances were observed at night than during the day. Figs. 4(a), (b), (c), and (d) are examples of scatter signals recorded photographically. In all photographs, time increases from left to right. The ground pulse occurs at the left edge of the trace.

From the above data, one may conclude (1) that the $2F$ scatter range may be used to determine the maximum usable frequency, and (2) that the leading edge of the scatter group corresponds to cloud scatter, while the maximum of the scatter group corresponds to ground scatter.

Determining MUF's

A nomograph has been constructed to use the scatter distance in determining the muf. It has been shown that the maximum amplitude of the scatter group corresponds to the first ray striking the ground at the outer edge of the skip zone. Also, the ground distance from the transmitter to the outer edge of the skip zone is associated with a frequency called the maximum usable frequency for that ground distance. Referring to Fig. 2(b), an operator at the transmitter might record the ground-scatter distance $\overline{TAB}$, the vertical-incidence critical frequency of the ordinary ray fc; and the frequency of the observed scatter f'; then the frequency f' is the muf for the ground distance D. Again it should be pointed out that f_c was used in place of f in this experiment due to necessity, and if complete $h'f$ curves are available, f should be used in place of f_c in the nomograph.

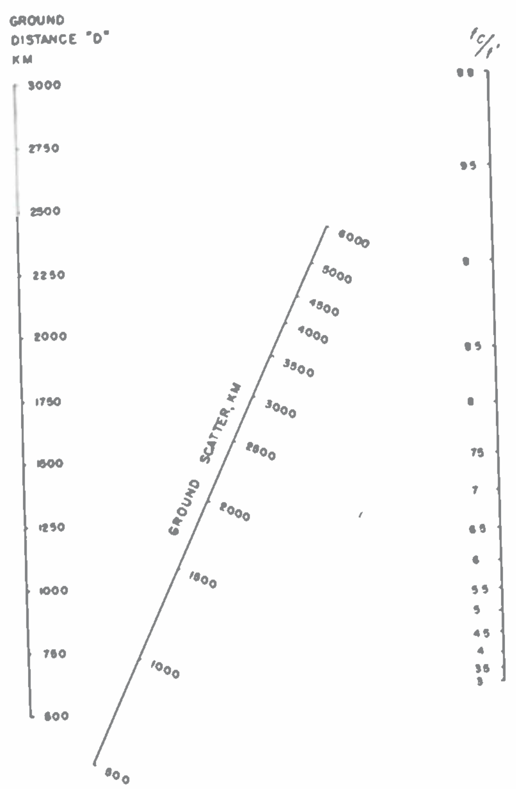


Fig. 5—Nomograph for determining the muf from scatter distance.

From the figure, $D/2=R\theta$, and $R \sin \theta = \overline{TA} \sin \phi$. Hence,

$$\overline{TAB} = \frac{2R \sin D/2R}{\sqrt{1-\left(\frac{fc}{f'}\right)^2}}$$

is the desired relation.

A nomograph based on this expression is shown in Fig. 5. As an example of its use, suppose observations indicate that fc is 6 Mc, f' is 8 Mc, and the observed scatter distance to the maximum amplitude is 1500 km. Placing a straight edge at 0.75 on the fc/f' line and at 1500 on the ground-distance line, the extended line will pass through 875 km on the ground-distance line. This, then, means that the maximum usable frequency for a distance of 875 km is 8 Mc.

Conclusions

(1) A correlation between maximum usable frequency and long-distance scatter has been established. This relationship could be useful to the operator of a long-distance radio station in determining the maximum usable frequency for a given distance. Thus, by utilizing equipment similar to that used in the present investigation, an hourly check on the muf by observation of the scatter distance as a function of frequency could be obtained. By determining the vertical-incidence critical frequency and using the nomograph, a daily record of the muf for the given operating distance could be obtained. From this muf record, the operating frequency for the following day's schedule could be estimated. This system has the considerable advantage of minimizing errors inherent in muf predictions made several months in advance.

(2) The experimental results reported indicate that the scatter is composed of echoes from E-layer clouds and from ground scatter. The former makes up that portion of the scatter pattern from the leading edge to just before the maximum amplitude of the scatter-spread echo. The maximum amplitude marks the beginning of the ground scatter and, following the maximum amplitude, the scatter pattern is a combination of echoes from both cloud and ground scattering sources.

Acknowledgment

The original idea for this investigation was suggested by Newbern Smith of the Central Radio Propagation Laboratory of the National Bureau of Standards. The author is grateful for the guidance and assistance of A. H. Waynick on this problem.

Square-Wave Analysis of Compensated Amplifiers*

PHILIP M. SEAL†, ASSOCIATE, IRE

Summary—A complete analysis of a single-stage compensated video-frequency amplifier is presented. In this amplifier high-frequency compensation is obtained by the use of a shunt peaking coil in series with the plate load resistance, and low-frequency compensation is obtained by using a resistance-capacitance network in series with the plate load resistance.

Both square-wave and sine-wave input voltages to the amplifier are considered. The square-wave analysis consists of a determination of the output wave shape when a symmetrical square wave is applied to the input. Equations for the output wave shape are derived and put in such a form that the output wave shape may be plotted for any desired frequency of input square wave. Output wave shapes are drawn for a number of typical operating conditions, and the corresponding frequency- and phase-response curves are drawn for comparison.

The effect of the cathode impedance, assumed negligible in the square-wave analysis, is considered briefly in a separate section, and its effect on low-frequency compensation is discussed. Derivations of many of the equations used are found in the Appendix.

I. Introduction

Wide-band amplifiers in which high-frequency compensation is obtained by means of a shunt peaking coil in series with the plate load resistance, and in which a resistance-capacitance network in series with the load resistance provides low-frequency compensation, are widely used in television and other applications in which it is important to preserve the shape of the applied input signal. In the actual testing of such amplifiers, square waves are commonly used. A square wave of voltage is applied to the input of the amplifier and the output wave shape is observed on an oscilloscope. In designing the amplifier, the object is to obtain an output voltage as nearly "square" as possible over a wide frequency range, consistent with a reasonable gain.

To the author's knowledge, a thorough square-wave analysis of a compensated amplifier has not been presented before. Many papers have been written dealing with the transient analysis, where a unit step function is applied to the input of the amplifier. This is especially true of the high-frequency-compensated case.[1] A few papers have been written dealing with the transient response at low frequencies, such as a recent paper by Schlesinger.[2] In this paper square waves are touched upon, but no complete analysis given.

A complete square-wave analysis of a compensated amplifier should prove useful in that it brings out the limitations inherent in such an amplifier and points out the optimum results which may be expected in performing a square-wave test on the amplifier. It should also prove useful in the design of the amplifier for optimum performance.

To obtain the complete analysis of the amplifier, the following three types of input and output voltages are used:

1. A dc step voltage suddenly applied at time $t=0$, and the resulting output voltage.
2. A symmetrical square wave of voltage applied at the input, and the resulting output wave shape.
3. Sinusoidal input and output voltages.

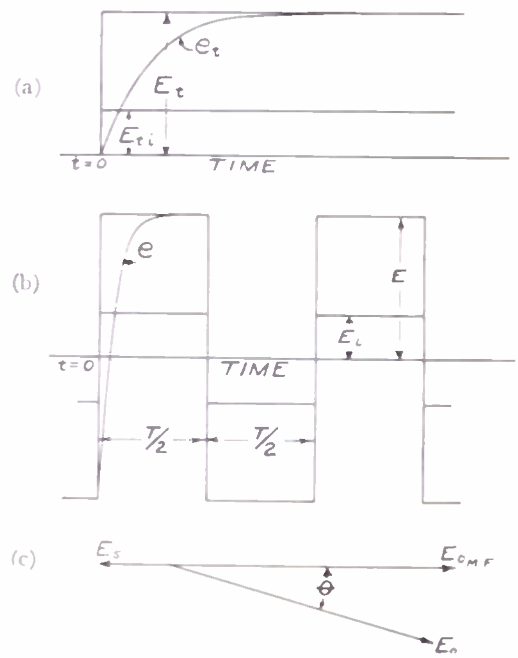


Fig. 1—Sketches illustrating the three types of input and output voltages being considered. (a) Step voltage input. (b) Square-wave voltage input. (c) Sine-wave voltage input.

To distinguish clearly between these three types, reference should be made to Fig. 1 and the list of symbols presented below. In Fig. 1(a):

E_{ti} = step function of voltage applied to the input of the amplifier

e_t = the resulting output voltage plotted as a function of time

E_t = the ideal output voltage. In magnitude, $E_t = g_m R_0' E_{ti}$

relative $e_t = e_t/E_t$.

In Fig. 1(b):

E_i = the amplitude of the square wave of voltage applied to the input of the amplifier

* Decimal classification: R255.13. Original manuscript received by the Institute, August 6, 1947; revised manuscript received, July 12, 1948. This paper represents material being presented to Purdue University in partial fulfillment for the degree of Doctor of Philosophy.

† University of Maine, Orono, Maine.

[1] H. E. Kallmann, R. E. Spencer, and C. P. Singer, "Transient response," Proc. I.R.E., vol. 33, pp. 169–195; March, 1945.

[2] Kurt Schlesinger, "Low-frequency compensation for amplifiers," *Electronics*, vol. 21, pp. 103–105; February, 1948.

e = the instantaneous value of the output voltage during a positive half cycle of the input voltage E_i

E = the amplitude of the ideal output square wave. In magnitude, $E = g_m R_0 E_i$

relative $e = e/E$.

T = the period of the square wave, or the time required to complete one cycle.

In Fig. 1(c):

E_s = the rms value of the input sine wave of voltage

E_0 = the rms value of the output sine wave of voltage

$E_{0M.F.}$ = the rms value of the output sine wave of voltage in the middle-frequency range. That is, $E_{0M.F.} = -g_m R_0 E_s$

θ = the angle by which E_0 lags or leads $E_{0M.F.}$. If E_0 lags $E_{0M.F.}$, θ is considered positive, and if E_0 leads $E_{0M.F.}$, θ is considered negative. This choice of signs is made deliberately in order to make the curves of relative phase shift versus frequency, Figs. 7 and 14, have a positive slope.

relative gain $= |E_0/E_{0M.F.}|$

relative phase shift $= \theta$.

A complete circuit diagram of the amplifier to be analyzed is shown in Fig. 2.

In the low-frequency analysis, the $R_K C_K$ network is neglected. That is, the reactance of C_K is considered negligibly small at the frequencies considered. Since this is not usually a valid assumption in video amplifiers at low frequencies, the effect of this network is considered in a separate section.

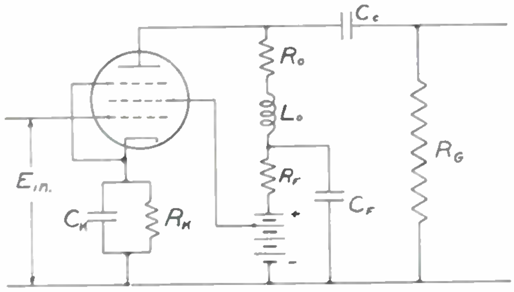


Fig. 2—Schematic diagram of a single-stage amplifier compensated for both low and high frequencies.

II. High-Frequency Compensation

At high frequencies the amplifier has the equivalent circuit diagram shown in Fig. 3. In this diagram C_t is the total shunt capacitance inherent in the amplifier. The shunting effect of the plate resistance of the tube, r_p, normally present in this type of equivalent circuit, may be neglected, since in video amplifiers r_p is much greater than the impedance of the network made up of R_0, L_0, and C_t. For the same reason, the shunting effect of the grid-leak resistance R_g may be neglected.

E_{in} is the input voltage applied to the amplifier, and may be E_{ti}, E_i, or E_s, depending on the type of input voltage considered. (See Fig. 1.)

In making a square-wave analysis of this circuit, it is assumed first that E_{in} is a step voltage having a value 0 up to time $t=0$, and a constant value E_{ti} thereafter. Hence the current I will be of the same nature, having a value 0 up to time $t=0$, and a constant value I thereafter.

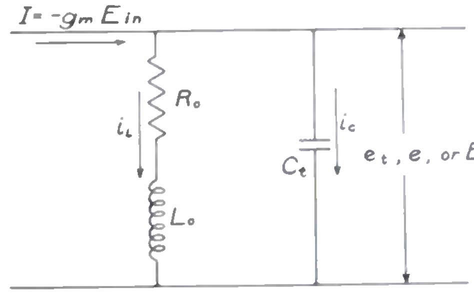


Fig. 3—Equivalent circuit diagram of the amplifier shown in Fig. 2 as it applies in the high-frequency case.

In the analysis two important parameters are used, as defined by the following equations[3]:

$$f_0 = \frac{1}{2\pi R_0 C_t} \tag{1}$$

$$n = \frac{2\pi f_0 L_0}{R_0}. \tag{2}$$

Physically, f_0 is the frequency at which the reactance of C_t is equal to R_0. It is also the frequency at which relative gain drops to 0.707 with no high-frequency compensation.

Derivation of the expression for e_t as a function of time is made with the use of Laplace transforms (see Appendix A for this derivation). The results indicate that three separate expressions may be obtained, depending on whether n is less than $\frac{1}{4}$, equal to $\frac{1}{4}$, or greater than $\frac{1}{4}$. The case where n is less than $\frac{1}{4}$ is the nonoscillatory case. Since the amount of compensation is too small to be of much practical value, this case is not considered further except for zero compensation where $n=0$ (see (19)).

Case where $n = \frac{1}{4}$:

This is the critical case. The result obtained is as follows:

When $n = \frac{1}{4}$,

$$e_t = IR_0[1 - (1 + \omega_0 t)\epsilon^{-2\omega_0 t}] \tag{3}$$

where,

$$\omega_0 = 2\pi f_0. \tag{4}$$

Hence,

$$e_t/E_t = \text{relative } e_t = 1 - (1 + \omega_0 t)\epsilon^{-2\omega_0 t}, \tag{5}$$

[3] D. G. Fink, "Principles of Television Engineering," McGraw-Hill Book Company, Inc., New York, N. Y., 1940; equation (113) on p. 221 and just above equation (147) on p. 223.

since

$$IR_0 = g_m R_0 E_{t_1} = E_t.$$

Equation (3) is an expression for the output voltage as a function of time when $n=\frac{1}{4}$. To express it in terms of the frequency of an applied square wave, the input voltage will be considered to be a square wave whose height above the axis is E_i.

By reference to Fig. 1(b), it is seen that $t=0$ is the time at which the wave goes suddenly from a value $-E_i$ to a value $+E_i$. It is seen, then, that the expression for the output voltage e in terms of the square-wave input of Fig. 1(b) is found by doubling (3) and then subtracting $E=g_m R_0 E_i = IR_0$ to take care of the shift in the position of the axis. This gives, when $n=1/4$,

$$e = IR_0[1 - 2(1 + \omega_0 t)\epsilon^{-2\omega_0 t}]. \tag{6}$$

Note that this assumes that the output voltage will reach the limiting voltage E by the time $t=T/2$. This is true within 0.1 per cent for square-wave frequencies less than $\frac{1}{2}$ of f_0.

Dividing both sides of (6) by $E=IR_0$ gives, when $n=1/4$,

$$e/E = \text{relative } e = 1 - 2(1 + \omega_0 t)\epsilon^{-2\omega_0 t}. \tag{7}$$

To obtain the equation for relative e in terms of the actual frequency of the square wave, the following equations are used:

$$T = 1/f \tag{8}$$

where f is the frequency of the applied square wave, and T is the period of the square wave

$$\gamma = f/f_0 \tag{9}$$

$$r = 2t/T. \tag{10}$$

That is, r represents the fractional part of a positive half cycle. It is equal to 0 at the beginning of the half cycle, and equal to 1 at the end of the half cycle. It is the variable along the horizontal axis used to plot all output wave shapes, and is independent of the actual frequency of the square wave.

If these substitutions are made, the following relation is obtained: When $n=1/4$,

$$\text{relative } e = 1 - 2\left(1 + \frac{\pi r}{\gamma}\right)\epsilon^{-2\pi r/\gamma}. \tag{11}$$

Case where n is greater than $\frac{1}{4}$:

When n is greater than $\frac{1}{4}$, damped oscillations inevitably occur in the output voltage, and hence there is always some overshoot. That is, the instantaneous value of the output voltage e will go through a peak value greater than the final value E.

General equations for this case are given below:

When the input voltage is a step voltage suddenly applied at $t=0$:

relative e_t

$$= 1 - \frac{2n}{\sqrt{4n-1}}\epsilon^{-\omega_0 t/2n}\cos\left(\omega_0 t\frac{\sqrt{4n-1}}{2n} + \phi\right) \tag{12}$$

where

$$\phi = \sin^{-1}\frac{2n-1}{2n}. \tag{13}$$

Equation (12) agrees with equation (7) in the paper by Kallman, Spencer, and Singer, when the difference in the definitions of the symbols is taken into account.[4]

When a square wave of voltage is applied, the equation becomes:

relative e

$$= 1 - \frac{4n}{\sqrt{4n-1}}\epsilon^{-\pi r/2n\gamma}\cos\left(\frac{\pi r}{\gamma}\frac{\sqrt{4n-1}}{2n} + \phi\right) \tag{14}$$

where r, γ, and ϕ are as previously defined.

Equation (14), like equations (6), (7), and (11), is true only if the voltage e coincides with the limiting value E by the time r reaches unity (i.e., within a half cycle). This is true within 0.5 per cent for values of n equal to 0.5 or less if the frequency of the square wave does not exceed 0.5 f_0. Since higher square-wave frequencies would be distorted too much to have a practical application in video amplifiers, and since values of n greater than 0.5 result in too much overshoot to be of practical use, (14) is considered accurate for all practical cases.

Special case where n = 0.5.

When $n=0.5$, (12) and (14) are very greatly simplified, as below:

For the step voltage,

$$\text{relative } e_t = 1 - \epsilon^{-\omega_0 t}\cos\omega_0 t. \tag{15}$$

For the applied square wave,

$$\text{relative } e = 1 - 2\epsilon^{-\pi r/\gamma}\cos\frac{\pi r}{\gamma}. \tag{16}$$

Frequency-Response and Phase-Shift Characteristics

If a sine wave of voltage is applied to the input of the amplifier, the relative gain of the amplifier at high frequencies and the relative phase shift are given by the following equations[5]:

$$\text{relative gain} = \frac{\sqrt{1+n^2\gamma^2}}{\sqrt{(1-n\gamma^2)^2+\gamma^2}} \tag{17}$$

$$\text{relative phase shift} = \theta = \tan^{-1}\frac{\gamma}{1-n\gamma^2} - \tan^{-1}n\gamma. \tag{18}$$

[4] See equation (7), p. 171 of footnote reference 1.

[5] See also equation (8), p. 172, of footnote reference 1. This checks equation (17) when differences in the definitions of symbols are taken into account.

Typical output wave shapes for square-wave inputs are shown in Figs. 4 and 5. In Fig. 4, the curve for zero compensation where $n=0$ is also shown for comparison.

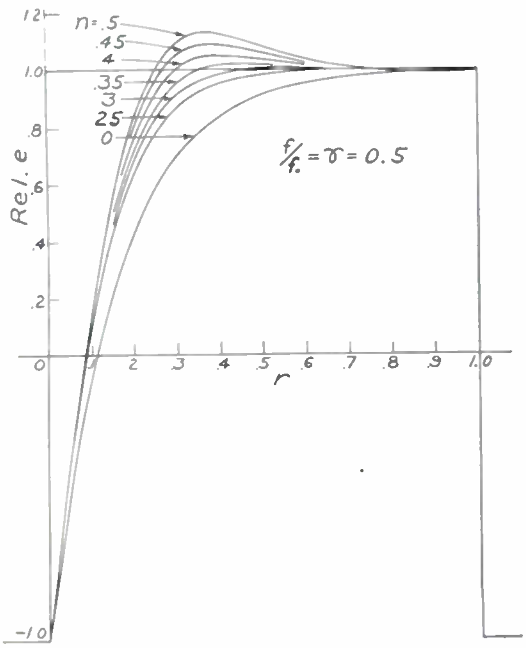


Fig. 4—Curves of relative e for various values of n. High-frequency-compensated amplifier, $\gamma=0.5$.

The equation for the uncompensated case is as follows: When

$$n = 0, \quad \text{relative } e = 1 - 2\epsilon^{-\pi r/\gamma}. \tag{19}$$

In Fig. 4 it is seen that as n is increased above the critical value, 0.25, the amount of overshoot increases.

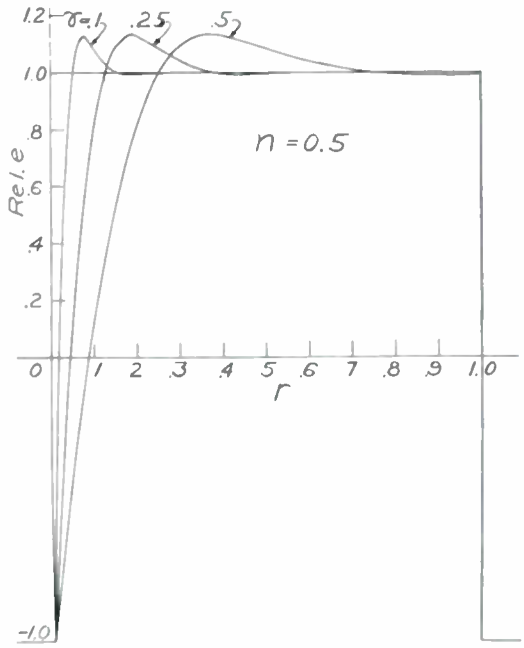


Fig. 5—Curves of relative e for applied square waves of various frequencies. High-frequency-compensated amplifier, $n=0.5$.

When $n=0.5$, the peak value of relative e is 1.133. It will be noted by referring to Fig. 5 that the peak value of relative e depends only on n, being independent of the frequency of the applied square wave.

In Fig. 4 it should be noted that the slope of the curve at $r=0$ is independent of the value of n.

Typical curves of relative gain and relative phase shift as a function of frequency are plotted in Figs. 6 and 7. The curve for $n=0.414$ (or $\sqrt{2}-1$) is included, because

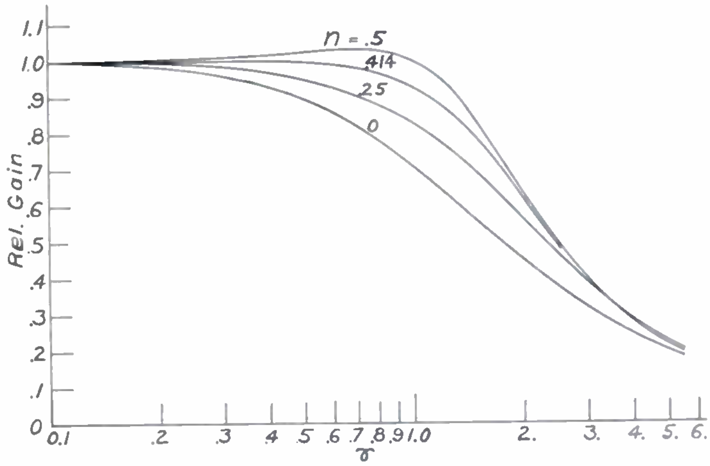


Fig. 6—Curves of relative gain and frequency for various values of n. High-frequency-compensated amplifier.

this is the largest values of n for which relative gain does not rise to a peak value greater than unity. When $n=0.5$, relative gain rises to a peak value of 1.03 when $\gamma=0.687$.

In comparing the curves for relative gain, Fig. 6, with the curves for a square-wave input, Fig. 4, it is interesting to note that, although no peak occurs in the gain characteristic until $n>0.414$, yet overshoot is obtained in the case of the square-wave input when $n>0.25$.

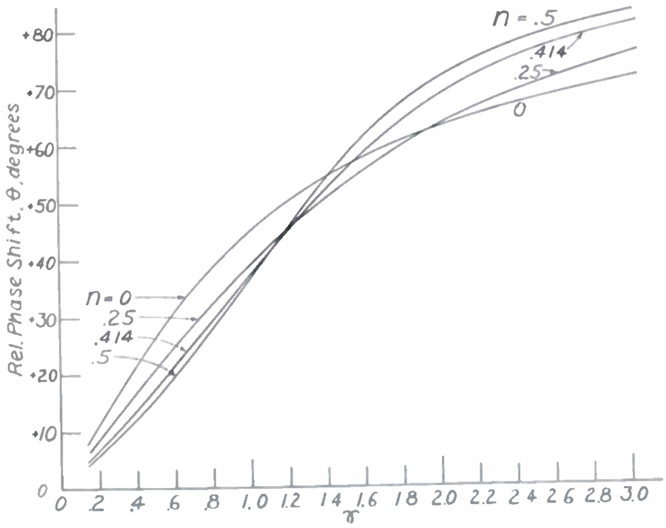


Fig. 7—Curves of relative phase shift θ and frequency for various values of n. High-frequency compensated amplifier.

Fig. 7 shows the relative phase shift as a function of γ. The ideal phase-shift curve would be a straight line through the origin. The curves for $n=0.414$ and $n=0.5$ closely approach this ideal when γ is less than 1.5, but for larger values of γ the curves become quite nonlinear.

III. Low-Frequency Compensation

At low frequencies the amplifier has the equivalent circuit diagram shown in Fig. 8. As in the high-frequency-compensation case, the square-wave analysis begins with the assumption that E_{in} is a step voltage having a value zero up to time $t=0$, and a constant value E_{ti} thereafter.

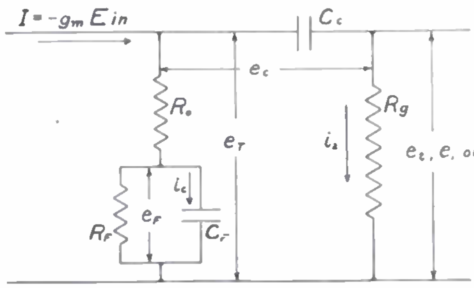


Fig. 8—Equivalent circuit diagram of the amplifier shown in Fig. 2 as it applies in the low-frequency case.

The analysis shows that three RC time constants are involved. The equations are greatly simplified if the following parameters are introduced:

$$f_1 = \frac{1}{2\pi C_F \dfrac{R_0 R_F}{R_0 + R_F}}, \tag{20}$$

$$f_2 = \frac{1}{2\pi C_F R_F}, \tag{21}$$

$$f_3 = \frac{1}{2\pi C_c R_g}, \tag{22}$$

$$\omega_1 = 2\pi f_1, \tag{23}$$

$$\omega_2 = 2\pi f_2, \tag{24}$$

$$\omega_3 = 2\pi f_3. \tag{25}$$

The expression for the output voltage e_t as a function of time is obtained as in the high-frequency case by the use of Laplace transforms. See Appendix B for the derivation. The final expression for relative e_t is as follows:

$$\text{relative } e_t = A\epsilon^{-\omega_2 t} - (A-1)\epsilon^{-\omega_3 t} \tag{26}$$

where

$$A = \frac{f_1 - f_2}{f_3 - f_2}. \tag{27}$$

If a square wave of voltage is applied to the input, the equations for the output wave shape cannot be obtained as easily as they can in the high-frequency case, because in general the output voltage will not have reached its limiting value at the end of a half cycle when the input voltage reverses. The method used here is taken from a paper by Chin.[6] The final equation obtained for relative e in terms of time and the period of the square wave is as follows:

relative e

$$= 2\left[\frac{A}{1+\epsilon^{-\omega_2 T/2}}\epsilon^{-\omega_2 t} - \frac{A-1}{1+\epsilon^{-\omega_3 T/2}}\epsilon^{-\omega_3 t}\right]. \tag{28}$$

By making use of (8) and (10), this equation is expressed in terms of the frequency f of the square wave and the variable r, with the following result:

relative e

$$= 2\left[\frac{A}{1+\epsilon^{-\pi f_2/f}}\epsilon^{-\pi r f_2/f} - \frac{A-1}{1+\epsilon^{-\pi f_3/f}}\epsilon^{-\pi r f_3/f}\right]. \tag{29}$$

Equation (29) may also be put into the following form, which is often more useful in making computations:

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

Frequency-Response and Phase-Shift Characteristics

When a low-frequency sinusoidal voltage is applied to the input of the amplifier, relative gain and relative phase shift may be obtained from the following formulas:

$$\text{relative gain} = \frac{\sqrt{1+(f_1/f)^2}}{\sqrt{1+(f_2/f)^2}\sqrt{1+(f_3/f)^2}}. \tag{31}$$

$$\text{relative phase shift, } \theta = -\tan^{-1}(f_1/f) + \tan^{-1}(f_2/f) + \tan^{-1}(f_3/f). \tag{32}$$

Application of Low-Frequency Equations

Equations (30) and (31) may be used to obtain the output wave shape and frequency-response curves for any desired combination of the three parameters f_1, f_2, and f_3. To determine suitable combinations for these parameters, (31) is most useful. It shows that, if f_1 is made equal to f_3, the low-frequency response curve will be the same as with no low-frequency compensation, except that relative gain will drop to 0.707 when $f=f_2$ instead of when $f=f_3$, as would be the case with no compensation. Thus, in designing the amplifier, it would seem desirable to make f_2 as small as possible by making the $R_F C_F$ product large, and then choosing R_F and C_F such that f_1 will equal f_3. However, if (20) is divided by (21), the following is obtained:

$$f_1/f_2 = \left(\frac{R_F}{R_0} + 1\right). \tag{33}$$

[6] P. T. Chin, "Circuit response to non-sinusoidal wave forms," *Electronics*, vol. 17, pp. 138–141; October, 1944.

This indicates that, the smaller f_2 is made with respect to f_1, the larger will be the resistance R_F compared to R_0. But too large a value of R_F is not practical because of the resulting dc voltage drop and power loss.

In practice it will be found desirable to make f_1 somewhat greater than f_3. This will introduce a small hump in the frequency-response curve at low frequencies, and improve the square-wave response.

The above considerations led to two choices of combinations of the parameters f_1, f_2, and f_3 to use as typical examples. In the first example, $f_1=5f_2$ (or $R_F=4R_0$) and $f_3=3.5f_2$. Fig. 9 shows output wave shapes for square-wave inputs of different frequencies, for this choice of parameters. The frequency of the input square wave is

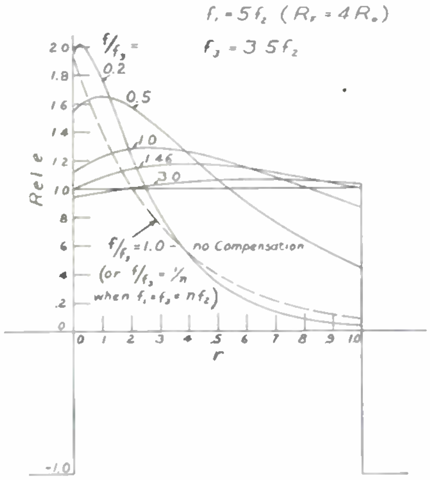


Fig. 9—Curves of relative e for various values of f/f_3. Low-frequency-compensated amplifier, $f_1=5f_2$ and $f_3=3.5f_2$.

expressed in terms of the ratio f/f_3. It will be noted that at very low frequencies relative e has a value greater than unity at the beginning of the half cycle, rises to a peak value, and then drops to a value lower than unity at the end of the half cycle. If the amouut by which relative e differs from unity at the beginning of the half cycle is referred to as ΔE, and the symmetry of the output wave about the axis is considered, the following equations result:

$$\text{relative } e = 1 + \Delta E \quad \text{when} \quad r = 0 \tag{34}$$

$$\text{relative } e = 1 - \Delta E \quad \text{when} \quad r = 1. \tag{35}$$

As the frequency of the input square wave is increased ΔE decreases, and in Fig. 9 reaches 0 when $f/f_3=1.46$. If there is no low-frequency compensation, ΔE never does reach 0. With compensation, ΔE will reach 0 only under certain conditions. To determine what these conditions are, an expression for ΔE is found. By setting $r=0$ in (29), $1+\Delta E$ is obtained. By setting $r=1$, $1-\Delta E$ is obtained. By subtracting the second from the first and dividing by 2, the following is obtained for ΔE:

$$\Delta E = A \tanh \frac{\pi}{2}\frac{f_2}{f} - (A-1)\tanh\frac{\pi}{2}\frac{f_3}{f}. \tag{36}$$

If ΔE is set equal to 0, and use is made of (27), the following is obtained:

$$\text{for } \Delta E = 0, \quad \frac{\tanh \frac{\pi}{2}\frac{f_3}{f}}{\tanh\frac{\pi}{2}\frac{f_2}{f}} = \frac{f_1 - f_2}{f_1 - f_3}. \tag{37}$$

From this equation it is found that ΔE cannot be 0 at some finite frequency unless the following relation holds:

$$\text{for } \Delta E \text{ to be } 0, \quad f_3/f_2 > \frac{f_1-f_2}{f_1-f_3}. \tag{38}$$

Or, if (38) is simplified further:

$$\text{for } \Delta E \text{ to be } 0, \quad f_1 > f_2 + f_3. \tag{39}$$

At very low frequencies, where ΔE is positive, the wave resembles the uncompensated case and may be described as "undercompensated." Where ΔE is negative, the wave has a rising characteristic, and hence may be described as "overcompensated." For the special case where $\Delta E=0$, the term "flat-compensated" could be used, although the wave may have a pronounced hump in it.

The dotted curve shown in Fig. 9, representing the output wave shape with no compensation when $f/f_3=1$, is included for comparison. It also gives the wave shape when $f/f_3=1/n$, provided the network is designed to make $f_1=f_3=nf_2$. For example, if $f_1=f_3=5f_2$, this curve represents the output wave shape for $f/f_3=0.2$. This compares closely with the solid curve for the same frequency.

For the second example, the parameters were chosen to make $f_1=11f_2$ (or $R_F=10R_0$) and $f_3=9f_2$. Fig. 10

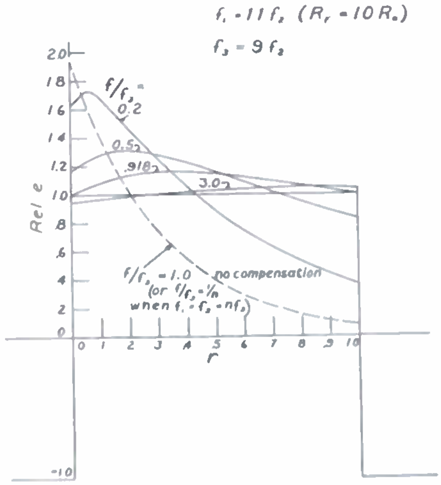


Fig. 10—Same as Fig. 9, except that $f_1=11f_2$ and $f_3=9f_2$.

shows typical output wave shapes for this case. The results show a definite improvement over those of Fig. 9, but a much larger value of R_F is required. For both cases it should be noted that ideal output wave shapes cannot be expected for square-wave frequencies much less than f_3. Thus it is desirable to make f_3 as low as possible. This means that the R_gC_c product should be made as large as possible. It must be remembered, however, that the size of R_g is limited by the grid-leak requirements of the following tube, while the size of C_c is limited by such factors as its leakage resistance, capacitance to ground, and the like.

The criterion for judging how nearly "square" the output wave is would seem to be the amount of variation in relative e throughout a half cycle. Fig. 11 shows this variation as a function of f/f_3 for the first set of parameters. The curve marked $1+\Delta E$ is the value of relative e at the beginning of the half cycle, the curve marked $1-\Delta E$ is the value of relative e at the end of the half cy-

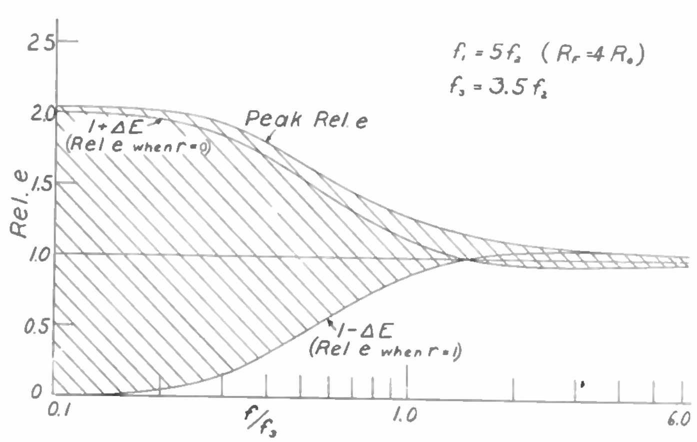


Fig. 11—Curves showing variation of relative e throughout a half cycle and f/f_3. Low-frequency-compensated amplifier, $f_1=5f_2$ and $f_3=3.5f_2$.

cle, and the curve marked peak relative e is the maximum value of relative e during the half cycle. The amount of variation in relative e throughout the half cycle for any desired frequency may be determined by noting the thickness of the cross-hatching at that frequency.

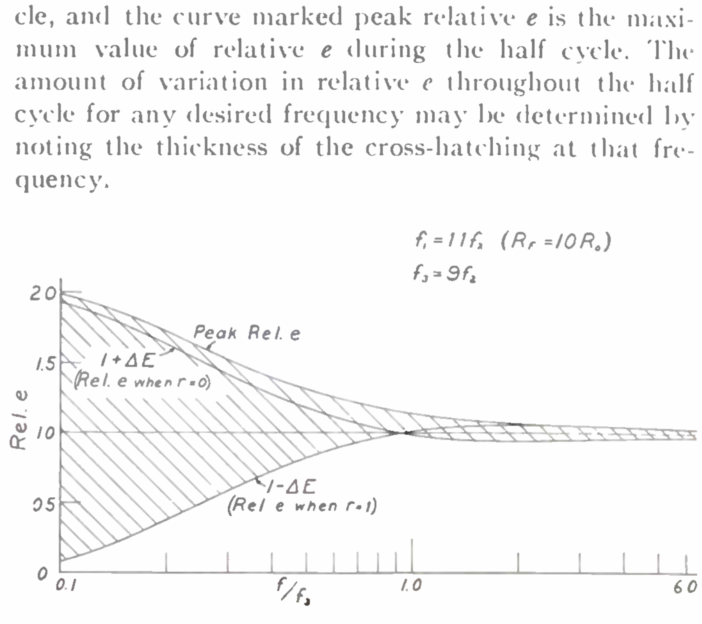


Fig. 12—Same as Fig. 11, except that $f_1=11f_2$ and $f_3=9f_2$.

Fig. 12 shows similar curves for the second set of parameters used. A certain amount of improvement is noted, especially for values of f/f_3 less than unity.

Curves of relative gain for a sine-wave input are shown in Fig. 13 for the two sets of parameters used and for the uncompensated case. For each of the two examples chosen, there is a pronounced hump in the response curve. Analysis shows that this hump will occur if the following inequality holds:

$$\text{For relative gain to have a peak, } f_1^2 > f_2^2 + f_3^2. \quad (40)$$

It is interesting to compare this inequality with the one shown in (39) for the square-wave case.

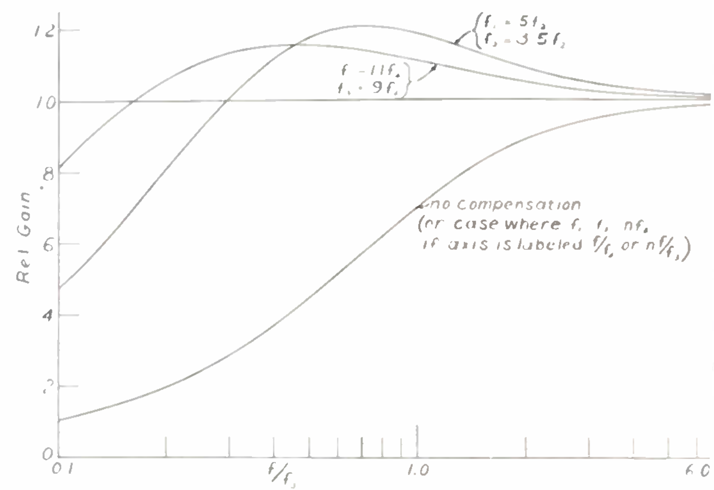


Fig. 13—Curves of relative gain and f/f_3. Low-frequency-compensated amplifier.

The curve labeled "no compensation" can also be used as the curve for the case where $f_1=f_3=nf_2$ if the abscissa is labeled $n\,f/f_3$. For example, if $n=10$, the values of f/f_3 along the axis would be 0.01 and 0.1, instead of 0.1 and 1.0.

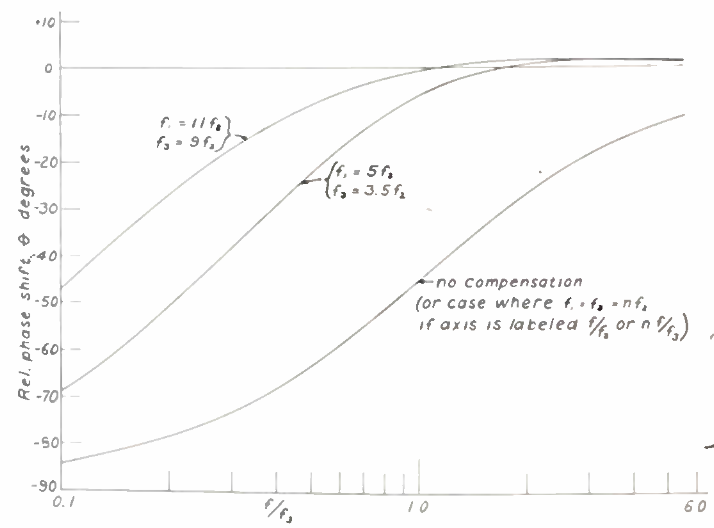


Fig. 14—Curves of relative phase shift θ and f/f. Low-frequency-compensated amplifier.

Fig. 14 shows curves of relative phase shift versus f/f_3 for the two examples chosen and for the uncompensated case. It is pointed out in the corresponding high-frequency case (Fig. 7) that the ideal phase characteristic is a straight line through the origin. In the low-frequency case, this means that the ideal phase shift is 0 for all frequencies. This is closely approached down to $f/f_3=1.0$ for the two examples chosen, but for lower frequencies no longer holds.

Effect of Cathode Impedance

All equations presented up to this point were derived under the assumption that the cathode resistor R_K was perfectly by-passed by the cathode by-pass capacitor C_K. This is not necessarily true, even for large values of C_K, partly because of the very low frequencies being considered, and partly because the use of high-g_m tubes tends to increase the degenerative effect of the cathode impedance.

Although a square-wave analysis of the network including the cathode impedance has not been attempted because the results would be extremely complex, formulas for relative gain and relative phase shift have been derived, and are presented below. (See Appendix C for the derivation.)

$$\text{relative gain} = \frac{\sqrt{1+(f_1/f)^2}\sqrt{1+(f_K/f)^2}}{\sqrt{1+(f_2/f)^2}\sqrt{1+(f_3/f)^2}\sqrt{1+(f_4/f)^2}} \qquad (41)$$

relative phase shift, θ

$$= \tan^{-1}(f_2/f)+\tan^{-1}(f_3/f)+\tan^{-1}(f_4/f) - \tan^{-1}(f_1/f)-\tan^{-1}(f_K/f) \qquad (42)$$

where

$$f_K = \frac{1}{2\pi C_K R_K} \qquad (43)$$

and

$$f_4 = f_K[1 + (g_p + g_m)R_K]. \qquad (44)$$

In (44), g_p is the plate conductance, or $1/r_p$. With pentode tubes, g_p is negligible compared to g_m.

Equation (41)may be greatly simplified if the amplifier is designed to make $f_K=f_3$ and $f_1=f_4$. For this special case,

$$\text{relative gain} = \frac{1}{\sqrt{1+(f_2/f)^2}}. \qquad (45)$$

If it is assumed that R_0, C_c, R_g, R_K, g_p, and g_m are known at the outset, the procedure would be as follows:

1. Calculate the required value of C_K by noting that, when $f_K=f_3$, $C_K=C_cR_g/R_K$.
2. Calculate f_K and f_4 by using (43) and (44).
3. Decide on a value for f_2 and calculate R_F as follows:

$$\text{when } f_1 = f_4, \qquad R_F = R_0\frac{f_4 - f_2}{f_2}. \qquad (46)$$

4. Calculate the required value of C_F by using (21). Note that the decision as to how low f_2 can be made depends on how large R_F can be without causing an excessive dc voltage drop.

Two numerical examples are presented for comparison. In the first, the cathode impedance is considered negligible, so that (31) holds, but the amplifier is designed to make $f_1=f_3$. In the second example, the cathode impedance is not neglected, but the amplifier is designed to make $f_K=f_3$ and $f_1=f_4$. In both examples f_2 is the same, and relative gain is given by (45).

Given values are:

$R_0=1{,}200$ ohms	$g_m=9{,}000$ micromhos
$C_c=0.05$ µf	$g_p=$ negligible
$R_g=0.5$ megohms	$R_K=160$ ohms

f_2 to be 1 cps

Results obtained for the two examples are as follows:

Cathode impedance neglected	*Cathode impedance considered*
$R_F=6{,}450$ ohms	$R_F=17{,}470$ ohms
$C_F=24.7$ µf	$C_F=9.1$ µf
	$C_K=156$ µf

The results show that, to counteract the degenerative effect of cathode impedance, R_F must be made much larger than would be required if the cathode resistance were perfectly by-passed.

Experimental Verification of Low-Frequency Analysis

To check the theoretical wave shapes shown in Fig. 9 a low-frequency compensated amplifier was set up, a square wave of voltage was applied to the input, and photographs were taken of the output wave shape as it appeared on the oscilloscope. The results are shown in Fig. 15. The circuit diagram of the amplifier used is that of Fig. 2, except that L_0 and C_K are omitted. C_K was omitted so that the cathode impedance could have no

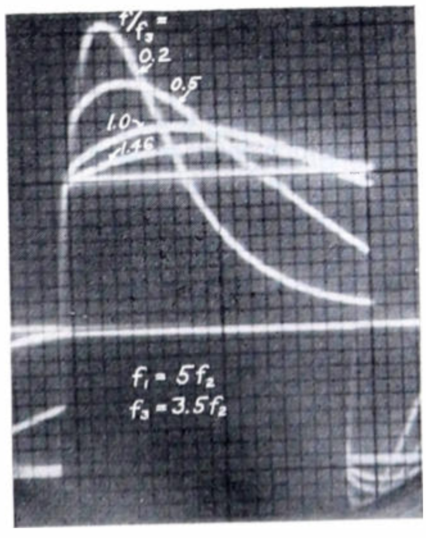


Fig. 15—Photograph of output wave shapes as obtained on an oscilloscope. Low-frequency-compensated amplifier.

effect on the output wave shape. The degenerative effect of R_K reduces the gain considerably, but cannot affect the wave shape since the effect is independent of frequency. The following values were used for the resistances:

$R_0 = 1000$ ohms
$R_F = 4000$ ohms
$R_g = 500{,}000$ ohms (including the input impedance of the oscilloscope)

These are typical values for a video amplifier, and, since $R_F = 4R_0$, $f_1 = 5f_2$ as shown in (33).

To obtain Fig. 15, a square wave of voltage having a frequency of 300 cps was applied to the input of the amplifier. The values of C_F and C_c were then adjusted to obtain the proper values of f/f_3 and the correct relation between the parameters f_3 and f_2. Table I shows the values of C_F and C_c used to obtain each wave.

TABLE I

f/f_3	C_F	C_c
0.2	0.093 μf	212 $\mu\mu f$
0.5	0.232 μf	530 $\mu\mu f$
1.0	0.465 μf	1060 $\mu\mu f$
1.46	0.680 μf	1550 $\mu\mu f$

The values of C_F and C_c used are much smaller than would normally be used in a video amplifier. This was done deliberately so that a frequency of 300 cps could be used. This frequency was chosen because it was easy to obtain on a square-wave generator, and because it could be fairly accurately reproduced on the screen of the oscilloscope.

Close examination will show that Fig. 15 is not an exact reproduction of Fig. 9. The differences occur mainly at the beginning of the half cycle, due probably to a certain amount of rounding off of the wave caused by distortion in the oscilloscope amplifier. When the possible sources of error are taken into account, Fig. 15 agrees very closely with Fig. 9 and verifies the square-wave analysis.

Conclusions

The analysis as presented is admittedly just one special case of many which could be made. It is confined to one particular type of high-frequency compensation and one type of low-frequency compensation. Furthermore, it is confined to a single stage of amplification. However, it is considered to be a rather complete analysis of this special case.

It is hoped that a simpler means of making a square-wave analysis can be worked out, possibly along the lines, suggested by Waidelich.[7] When this is done, it is hoped to extend the analysis to multistage amplifiers and more complicated types of networks.

Acknowledgment

The author wishes to acknowledge gratefully the help of R. P. Siskind, who is in charge of communications at Purdue University, in the preparation of this paper. The section entitled "Effect of Cathode Impedance," which includes hitherto unpublished formulas for relative gain and relative phase shift derived by him, is included with his permission.

Appendix A

High-Frequency Compensation—Square-Wave Analysis

In Fig. 3, consider E_{in} as a step voltage having a value zero up to time $t=0$ and a constant value E_{ti} thereafter. The current I will be of the same form, being zero up to time $t=0$ and having a constant value I thereafter. By applying Kirchhoff's laws, the following three equations may be written:

$$I = i_L + i_c \tag{47}$$

$$e_t = i_L R_0 + L_0 \frac{di_L}{dt} \tag{48}$$

$$i_c = C_t \frac{de_t}{dt} \tag{49}$$

where e_t is the instantaneous value of the output voltage.

By determining the Laplace transform of each of these three equations, the following three equations are obtained[8]:

$$I/s = \mathcal{L}(i_L) + \mathcal{L}(i_c) \tag{50}$$

$$\mathcal{L}(e_t) = R_0\mathcal{L}(i_L) + sL_0\mathcal{L}(i_L) \tag{51}$$

$$\mathcal{L}(i_c) = sC_t\mathcal{L}(e_t). \tag{52}$$

These three equations involve three unknowns, $\mathcal{L}(i_L)$, $\mathcal{L}(i_c)$, and $\mathcal{L}(e_t)$. When $\mathcal{L}(i_L)$ and $\mathcal{L}(i_c)$ are eliminated from the equations,

$$\mathcal{L}(e_t) = I \frac{L_0 + \dfrac{R_0}{s}}{s^2L_0C_t + sR_0C_t + 1} \tag{53}$$

where s is considered to be a complex variable.

If the parameters f_0, $\omega_0 = 2\pi f_0$, and n (see (1) and (2)) are now used,

$$L_0C_t = \frac{nR_0}{\omega_0} \times \frac{1}{\omega_0 R_0} = \frac{n}{\omega_0^2} \tag{54}$$

$$R_0C_t = 1/\omega_0. \tag{55}$$

When (54) and (55) are substituted in (53) and both numerator and denominator are multiplied by ω_0^2/n,

$$\mathcal{L}(e_t) = \frac{I\omega_0^2}{n} \frac{L_0 + R_0/s}{s^2 + \dfrac{\omega_0 s}{n} + \dfrac{\omega_0^2}{n}} \tag{56}$$

[7] D. L. Waidelich, "The steady-state operational calculus," Proc. I.R.E., vol. 34, pp. 78P–83P; February, 1946.

[8] See, for example, M. F. Gardner and J. L. Barnes, "Transients in Linear Systems, Volume I," John Wiley and Sons, Inc., New York, N. Y., 1942.

The denominator is now set equal to zero and solved for s to obtain the roots:

$$s = -\frac{\omega_0}{2n}[1 \mp \sqrt{1-4n}]. \tag{57}$$

This leads to three special regions:

1. $n<1/4$, in which s has two real roots
2. $n=1/4$, in which s has two equal roots, $=-2\omega_0$
3. $n>1/4$, in which s has two complex roots,

$$= -\frac{\omega_0}{2n}[1 \mp j\sqrt{4n-1}].$$

The first region is not considered further, as the amount of compensation is too small to be of practical value. By setting $L_0=0$, (53) could be used to derive (19) for the uncompensated case.

Case where $n=\frac{1}{4}$

For this case, (56) simplifies to the following:

$$\mathcal{L}(e_t) = 4I\omega_0^2\left[\frac{L_0+R_0/s}{(s+2\omega_0)^2}\right] \tag{58}$$

or

$$\mathcal{L}(e_t) = \omega_0 R_0 I\frac{1}{(s+2\omega_0)^2} + 4\omega_0^2 R_0 I\frac{1}{s(s+2\omega_0)^2} \tag{59}$$

since $4\,\omega_0^2\,L_0 = 4\,\omega_0 \times nR_0 = \omega_0 R_0$ when $n=1/4$.

When the inverse transform is found and the proper substitutions made, the following equation is obtained[9]:

$$e_t = IR_0[1-(1+\omega_0 t)\epsilon^{-2\omega_0 t}]. \tag{60}$$

This is the same as (3).

Case where $n>\frac{1}{4}$

For this case, (56) is put in the following form:

$$\mathcal{L}(e_t) = \frac{I\omega_0^2}{n}\left[\frac{L_0}{(s+\alpha)(s+\beta)} + \frac{R_0}{s(s+\alpha)(s+\beta)}\right] \tag{61}$$

where

$$\alpha = \frac{\omega_0}{2n}[1-j\sqrt{4n-1}] \tag{62}$$

and

$$\beta = \frac{\omega_0}{2n}[1+j\sqrt{4n-1}]. \tag{63}$$

From tables of inverse transforms,[10]

$$e_t = \frac{I\omega_0 L_0}{j\sqrt{4n-1}}(\epsilon^{-\alpha t}-\epsilon^{-\beta t}) + IR_0 - \frac{IR_0 n}{j\omega_0\sqrt{4n-1}}(\beta\epsilon^{-\alpha t}-\alpha\epsilon^{-\beta t}). \tag{64}$$

[9] See formulas 2.118 and 2.137, p. 346, in footnote reference 8.

[10] See formulas 1.105 and 1.108, pp. 338, 339, in footnote reference 8.

When the values for α and β given in (62) and (63) are substituted in (64), and a number of trigonometric identities are resorted to, the equations below are finally obtained:

relative e_t

$$= \frac{e_t}{IR_0} = 1 - \frac{2n}{\sqrt{4n-1}}\epsilon^{-\omega_0 t/2n}\cos\left(\omega_0 t\frac{\sqrt{4n-1}}{2n}+\phi\right) \tag{65}$$

where

$$\phi = \sin^{-1}\frac{2n-1}{2n}. \tag{66}$$

These equations are the same as (12) and (13).

Appendix B

Low-Frequency Compensation—Square-Wave Analysis

As in the high-frequency-compensation case, the low-frequency analysis begins with the assumption that the input voltage is a step voltage having a value zero up to time $t=0$ and a constant value E_i thereafter. In the equivalent circuit diagram, Fig. 8, the current I is also a step current, having a value zero up to time $t=0$ and a constant value I thereafter. Hence, the following differential equations may be set up:

$$e_T = IR_0 + e_F \tag{67}$$

$$i_1 = C_F\frac{de_F}{dt} = I - \frac{e_F}{R_F} \tag{68}$$

$$e_t = e_T - e_c \tag{69}$$

$$i_2 = C_c\frac{de_c}{dt} = e_t/R_g. \tag{70}$$

In order for these equations to be true, i_2 must be assumed negligible compared to I. This is true in the normal video amplifier.

These four equations involve four unknowns, e_T, e_F, e_c, and e_t, and it is desired to eliminate all but the output voltage e_t. To do this, the Laplace transforms of these four equations are first obtained, as follows:

$$\mathcal{L}(e_T) = IR_0/s + \mathcal{L}(e_F) \tag{71}$$

$$sC_F\mathcal{L}(e_F) = I/s - \frac{\mathcal{L}(e_F)}{R_F} \tag{72}$$

$$\mathcal{L}(e_t) = \mathcal{L}(e_T) - \mathcal{L}(e_c) \tag{73}$$

$$sC_c\mathcal{L}(e_c) = \frac{\mathcal{L}(e_t)}{R_g}. \tag{74}$$

When $\mathcal{L}(e_T)$, $\mathcal{L}(e_F)$, and $\mathcal{L}(e_c)$ are eliminated from these equations, the following is obtained:

$$\mathcal{L}(e_t) = \frac{IR_0}{s+\dfrac{1}{R_gC_c}}$$

$$+\frac{I}{C_F}\frac{1}{\left(s+\frac{1}{R_gC_c}\right)\left(s+\frac{1}{R_FC_F}\right)}. \tag{75}$$

The inverse transform is as follows:

$$e_t = IR_0\epsilon^{-t/R_gC_c} + \frac{I}{C_F}\frac{\epsilon^{-t/R_gC_c}-\epsilon^{-t/R_FC_F}}{\frac{1}{R_FC_F}-\frac{1}{R_gC_c}} \tag{76}$$

relative $e_t = e_t/IR_0$

$$= \epsilon^{-t/R_gC_c} + \frac{\frac{1}{R_0C_F}}{\frac{1}{R_FC_F}-\frac{1}{R_gC_c}}(\epsilon^{-t/R_gC_c}-\epsilon^{-t/R_FC_F}). \tag{77}$$

If the parameters f_1, f_2, f_3, ω_1, ω_2, and ω_3 (equations (20) through (25)) are now introduced, (77) becomes:

$$\text{relative } e_t = \frac{f_1-f_2}{f_3-f_2}\epsilon^{-\omega_2 t} - \frac{f_1-f_3}{f_3-f_2}\epsilon^{-\omega_3 t} \tag{78}$$

$$= A\epsilon^{-\omega_2 t} - (A-1)\epsilon^{-\omega_3 t} \tag{79}$$

when (27) is used.

This is the same as (26).

Appendix C

Low-Frequency Compensation—Relative Gain Formulas when Effect of Cathode Impedance is Considered

Fig. 16 shows two forms of the equivalent circuit diagram of an amplifier in which Z_L represents the load impedance, and Z_K the impedance of the cathode circuit. Since, in the video amplifier, Z_L is very small compared to r_p, it can be seen by comparing Fig. 16(b) with Fig. 8 that the essential difference between the two, as far as relative gain is concerned, lies in the expression for the input current I. That is,

$$R = \frac{1}{1+(g_p+g_m)Z_K} \tag{80}$$

where R is defined as

$$\frac{\text{relative gain when } Z_K \text{ is considered}}{\text{relative gain when } Z_K \text{ is neglected}}.$$

In the video amplifier being considered,

$$Z_K = \frac{1}{\frac{1}{R_K}+j\omega C_K} \tag{81}$$

$$= \frac{R_K}{1+j\omega C_KR_K} \tag{82}$$

$$= \frac{R_K}{1+j\frac{f}{f_K}} \tag{83}$$

when (43) for f_K is used.

From this relation, (80) becomes

$$R = \frac{1+j\frac{f}{f_K}}{1+(g_p+g_m)R_K+j\frac{f}{f_K}}. \tag{84}$$

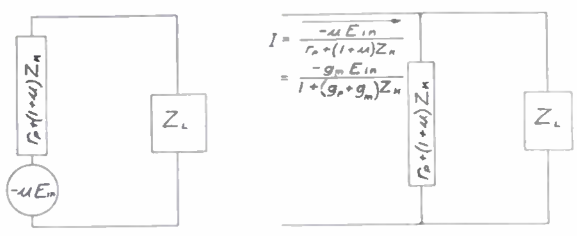
If both numerator and denominator are multiplied by $-jf_K/f$,

$$R = \frac{1-j\frac{f_K}{f}}{1-j\frac{[1+(g_p+g_m)R_K]f_K}{f}} \tag{85}$$

$$= \frac{1-j\frac{f_K}{f}}{1-j\frac{f_4}{f}} \tag{86}$$

when equation (44) for f_4 is used.

From this relation, (41) and (42) are easily obtained.



(a) (b)

Fig. 16—Equivalent circuit diagram of an amplifier when cathode impedance is considered. (a) Constant-voltage form of circuit. (b) Constant-current form of circuit.

Correction

C. T. Tai, author of the paper, "High-Frequency Polyphase Transmission Line," which appeared on pages 1370–1376 of the November, 1948, issue of the Proceedings of the I.R.E., has brought the following error to the attention of the editors:

Equation (49) on page 1374 should be corrected to read

$$R_{cn} = \frac{240}{n}\, ln\left(\frac{2D}{na}\right) = \frac{240}{n}\, ln\left(\frac{2b_{12}}{na\sin\phi}\right).$$

Contributors to the Proceedings of the I.R.E.

Carl C. Bath (A'46) was born in Abilene, Kan., on July 24, 1912. After graduating from Abilene High School in 1930, he was engaged in installing and servicing radio and sound equipment in Abilene and Salina, Kan. In 1942 and 1943, he was an instructor in the Signal Corps radar schools in Lexington, Ky., and Baltimore, Md. Early in 1944, Mr. Bath became associated with the Office of the Chief Signal Officer in Washington, D. C., in writing and revising technical manuals and instructional aids. In the latter part of 1944 he was sent to Europe as a field service consultant with the Office of Scientific Research and Development, specializing in the SCR-584 radar. His duties involved field instruction to Army maintenance personnel, and assisting in developing field modifications for extending the utility of the equipment. For his overseas work, Mr. Bath was awarded the Medal of Freedom.

CARL C. BATH

During the summer of 1945, Mr. Bath was associated with the Signal Corps radio relay tests which were held in California. He joined the research and development section of the Bendix Radio Division of the Bendix Aviation Corporation in the fall of 1945 as a research engineer, specializing in the development of microwave and special-purpose electronic equipment.

❖

Arthur H. Benner (S'43) was born on March 29, 1922, in Leavenworth, Kan. He received the B.S. degree in electrical engineering from the University of Kansas in 1944, and the M.S. degree from the Pennsylvania State College in 1948. He was an instructor in electrical engineering at the University of Kansas in 1944. From 1944 until 1946 he was an officer in the Signal Corps, and in 1946 he joined the staff of the Ordnance Research Laboratory, State College, Pa., as a research engineer.

ARTHUR H. BENNER

Mr. Benner is a member of Tau Beta Pi, and an associate member of Sigma Xi.

❖

For a photograph and biography of HAROLD GOLDBERG, see page 1259 of the October, 1948, issue of the PROCEEDINGS OF THE I.R.E.

Lester M. Field was born on February 9, 1918, at Chicago, Ill. He received the B.S. degree from Purdue University in 1939, and the Ph.D. degree from Stanford University in 1944. He was acting instructor in 1941, and acting assistant professor from 1942 to 1944, both in electrical engineering, at Stanford University.

LESTER M. FIELD

In 1944 Dr. Field joined the Bell Telephone Laboratories as a member of the magnetron development group, and later the electron dynamics group, of the physical research department. In September, 1946, Dr. Field returned to Stanford University as acting associate professor of electrical engineering, and became associate professor in September, 1948.

❖

For a biography and photograph of W. E. GORDON, see page 1259 of the October, 1948, issue of the PROCEEDINGS OF THE I.R.E.

❖

Andrew V. Haeff (A'34–M'40–SM'43–F'49) was born in Moscow, Russia, on December 30, 1904. In 1928, he received the degree of Electrical and Mechanical Engineer from the Russian Polytechnic Institute at Harbin, China. In 1928 he majored in electrical engineering at the California Institute of Technology, where he obtained the M.S. degree in 1929 and the Ph.D. degree in 1932.

ANDREW V. HAEFF

Until the end of 1933, Dr. Haeff was a Special Research Fellow at the C.I.T., engaging in research work on ultra-high-frequency problems. From 1934 to 1941, he was a member of the vacuum-tube research section of the RCA Manufacturing Company, specializing in research on uhf tubes and circuits. On March 1, 1941, he joined the research staff of the Naval Research Laboratory at Washington, D. C., in order to devote his full time to naval research problems, particularly in connection with wartime development of radar. His present position is that of consultant in electronics and head of the Vacuum Tube Research Section at the Naval Research Laboratory.

Leon Riebman (S'43–A'44) was born on April 22, 1920, in Coatesville, Pa. He received the B.S. degree in electrical engineering in 1943 and the M.S. degree in 1947 from the Moore School of Electrical Engineering, University of Pennsylvania.

LEON RIEBMAN

After attending the United States Navy Midshipman School, he was assigned to duty in radar development at the Naval Research Laboratory in 1944. In March, 1946, he joined the research and development staff of Philco Corporation as a senior engineer. Since October 1, 1948, he has been attending the Moore School of Electrical Engineering, University of Pennsylvania, under an Atomic Energy Commission predoctoral fellowship.

Mr. Riebman is a member of Sigma Xi, Tau Beta Pi, Eta Kappa Nu, Pi Mu Epsilon, and Sigma Tau.

❖

Philip M. Seal (A'41) was born on September 3, 1907, in Springfield, Mass. He received the B.S. degree in electrical engineering from the Worcester Polytechnic Institute in 1930, and the M.S. degree from the same institution in 1932.

PHILIP M. SEAL

From 1934 to 1937 Mr. Seal was employed by the Westinghouse Electric and Manufacturing Company in their radio test department. He was employed by the same company during the summers of 1939, 1940, and 1941, the last two summers being spent in the radio engineering department. His teaching career began as an instructor in electrical engineering at the University of Maine during the year 1937–1938. From there he went to Purdue University and was an instructor in electrical engineering from 1938 to the fall of 1947. In September, 1947, he returned to the University of Maine, where he is now an assistant professor in electrical engineering. He is a member of Tau Beta Pi, Sigma Xi, and the American Society for Engineering Education.

❖

For a photograph ane biography of CLAUDE E. SHANNON, see page 1389 of the November, 1948, issue of the PROCEEDINGS OF THE I.R.E.

Correspondence

Degenerative Feedback*

In the case known as "degenerative feedback" or "current feedback," we have found two different expressions for the amplification:

In Terman's book[1] there is given the general expression

$$A' = \frac{A}{1 - A\beta},$$

while in the handbook of Batcher and Moulic[2] there is a special expression for this case

$$A' = \frac{A}{1 - \beta(1 + A)}; \qquad \left[-\beta = \frac{R_K}{R_L}\right].$$

The circuit is shown in Fig. 1 and the equivalent circuit in Fig. 2.

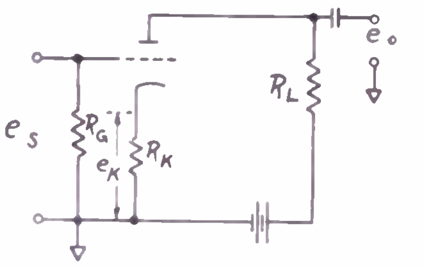


Fig. 1

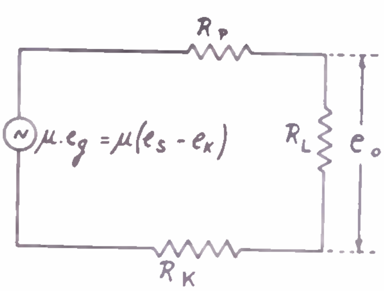


Fig. 2

$$e_0 = i_p \cdot R_L = \frac{\mu\, e_G \cdot R_L}{R_p + R_L + R_K}$$

$$= \frac{\mu\, e_s\, R_L}{R_p + R_L + R_K + \mu\, R_K},$$

$$A = \frac{e_0}{e_s} = \frac{\mu\, R_L}{R_p + R_L + R_K(1 + \mu)},$$

$$A = \frac{e_0}{e_s}(R_K = 0) = \frac{\mu\, R_L}{R_p + R_L},$$

and

$$\mu = \frac{A(R_p + R_L)}{R_L}.$$

For $R_K \neq 0$,

$$A' = \frac{A}{1 + \dfrac{R_K}{R_p + R_L}(1 + \mu)}$$

$$= \frac{A}{1 + R_K\left(\dfrac{1}{R_p + R_L} + \dfrac{A}{R_L}\right)}$$

$$A' = \frac{A}{1 + \dfrac{R_K}{R_L}\left(A + \dfrac{R}{R_p + R_L}\right)}.$$

A second form, eliminating μ, becomes

$$A' = \frac{A}{1 + A\dfrac{R_K}{R_L}\dfrac{M + 1}{\mu}}.$$

Following Terman,

$$A' = \frac{A}{1 - A\beta},$$

and, consequently

$$-\beta = \frac{R_K}{R_L}\,\frac{\mu + 1}{\mu}.$$

For large enough μ this agrees with the definition of β implicit in Terman's discussion of current feedback[3] $-\beta = Rk/R_L$ also given explicitly in the work of Batcher and Moulic.

Consequently one can correct the definition of β by the factor $(\mu+1)/\mu$ keeping the general definition for A or one can conserve the definition of β in which case the gain formula becomes

$$A = \frac{A}{1 - \beta\left(1 + \dfrac{R_L}{R_p + R_L}\right)}$$

which only for $R_p \ll R_L$ leads to the relation given by Batcher and Moulic:

$$A = \frac{A}{1 - \beta(1 + A)}.$$

MORTON NADLER and VICTOR TOMA
Tesla Vyzkum
Prague Strasnice
Czechoslovakia

* Received by the Institute, August 4, 1948.

[1] F. E. Terman, "Radio Engineering," McGraw-Hill Book Co., Inc., New York, N. Y., 1947; p. 311.

[2] R. R. Batcher and W. Moulic, "Electronic Control Handbook," Caldwell-Clements, Inc., New York, N. Y., 1946.

[3] F. E. Terman, "Radio Engineers Handbook," McGraw Hill Book Co., Inc., New York, N. Y., 1943, p. 403.

Simplified Analysis of Reactance Tube*

In Fig. 1,

r = plate resistance

a, b, c = generalized impedances

u = amplification factor

Then

$$z_{11} = \frac{e}{i_1} = r + c(1 + u)$$

$$= \frac{[u(b + c) + r + c][r + c(1 + u)]}{r + a + (b + c)(1 + u)}$$

$$= \frac{(a + b)[r + c(1 + u)]}{r + a + (b + c)(1 + u)}.$$

This may be written in the form:

$$z_{11} = \frac{1}{\dfrac{1}{a+b} + \dfrac{1}{r+c(1+u)} + \dfrac{u}{r+c(1+u)}\,\dfrac{b}{(a+b)}},$$

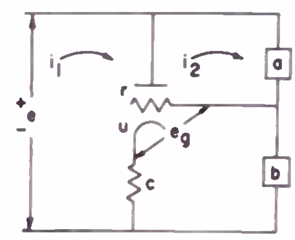


Fig. 1

which says that the reactance tube may be represented by three parallel impedances, as in Fig. 2. The impedance c may be disposed of by suitable by-passing or omission, and a and b may be made resistive or reactive to suit.

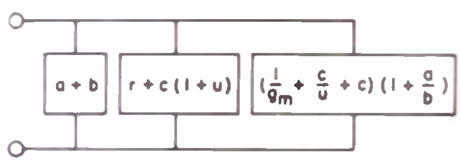


Fig 2

It should be noted that, on the basis of linear circuit theory, the above solution involves no approximations.

B. B. DRISKO
Transducer Corp.
63 Melcher St.
Boston 10
Mass.

* Received by the Institute, July 26, 1948.

Empirical Formulas for Amplification Factor*

It has recently been suggested by Herold[1] that analytical convenience is given by using an empirical power series for b in the expression for the amplification factor of a triode

$$\mu = (2\pi NS + c)b$$

where S is the grid to anode spacing, and b and c are functions of Nd (the fractional coverage N being the number of grid wires per unit length, and d the grid wire diameter.

From Herne's[2] accurate results, Herold obtains the formula

$$b = 0.2 + 6.8Nd + 680(Nd)^4$$

which is accurate to better than 8 per cent over the range $Nd = 0.002$ to $Nd = 0.5$, thus more than adequately covering normal usage. To the same accuracy, the correction term c is negligible.

* Received by the Institute October 1, 1948.

[1] E. W. Herold, correspondence on "Empirical formula for amplification factor," PROC. I.R.E., vol. 35, p. 473, May, 1947.

[2] H. Herne, "Valve amplification factor," Wireless Eng., vol. 21, pp. 59–65, February, 1944.

Any number of such empirical formulas may be found, depending on the index of the term used to start the series, and the writer has used the form

$$b = 2.4(Nd)^{1/2} + 90(Nd)^3$$

which is accurate to within 5 per cent over the range $Nd = 0.02$ to 0.33, or within 2 per cent from $Nd = 0.03$ to 0.3, the deviations within this range being rather less than with Herold's result. A further term, $7000\,(Nd)^6$, may be added to give accuracy to values of Nd in excess of 0.5.

Whether it is necessary to have an empirical formula which is accurate over such a wide range is questionable. Normally, for analytical purposes, it is only required to know the rate of variation of μ with electrode dimensions over relatively narrow ranges. For this purpose it would appear simpler to express b as a power of Nd, say

$$b = k(Nd)^m$$

where the coefficient k and index m are both functions of Nd, being chosen to give the true values of b and its first derivative at the point concerned. It is found empirically that such an expression represents the variation of b with all the accuracy normally required. Thus, at $Nd = 0.133$ evaluation of the constants gives $b = 0.82\,Nd$, and it is found that this particular formula only departs from Herne's values by 0.1 per cent at $Nd = 0.133 \pm 5$ per cent and by 1 per cent at $Nd = 0.133 \pm 15$ per cent. The errors introduced become a little greater for higher values of Nd, and correspondingly smaller for lower values.

Values of k and m have been determined from Herne's figures and are given in Table I for Nd from 0.005 to 0.40.

TABLE I

Nd	k	m	Nd	k	m
0.005	0.78	0.24	0.20	18.7	1.46
0.01	1.10	0.29	0.22	24.1	1.62
0.02	1.48	0.36	0.24	30.2	1.79
0.04	2.26	0.48	0.26	39.5	1.97
0.06	3.13	0.59	0.28	50.7	2.16
0.08	4.09	0.69	0.30	66.7	2.38
0.10	5.46	0.81	0.32	87.1	2.61
0.12	6.95	0.92	0.34	114	2.86
0.14	8.97	1.04	0.36	153	3.13
0.16	11.48	1.18	0.38	205	3.43
0.18	14.60	1.31	0.40	285	3.78

We then have, to the first order (since c has been neglected),

$$\mu = 2\pi NSk(Nd)^m$$
$$= 2\pi kSd^m N^{m+1}$$

which, when $m = 1$, is the form of the early empirical formula of van der Bijl.[3]

The other purpose for an empirical formula, namely, for calculation in the absence of the full curves, is again most readily served by simple power expressions which will be valid over limited ranges. It has been found that three simple expressions may be used to cover the whole normal working range to an accuracy which is better than 5 per cent. These expressions are:

$b = 2.5(Nd)^{1/2}$ for use from $Nd = 0.02$ to 0.09

$b = 8.5Nd$ for use from $Nd = 0.09$ to 0.20

$b = 43(Nd)^2$ for use from $Nd = 0.20$ to 0.34,

[3] H. van der Bijl, "The Thermionic Vacuum Tube," McGraw-Hill Book Co., New York, N. Y., 1920; p. 231.

the percentage error from Herne's results being shown in Fig. 1.

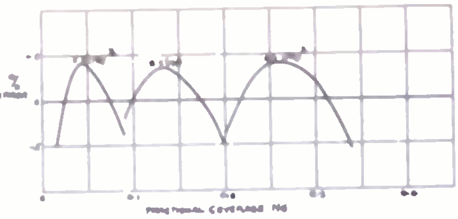


Fig. 1—Curves showing errors of proposed simple formulas.

The convenience of working with such simple approximate formulas, readily amenable to slide rule calculation, will be appreciated by those who have made much use of the older formulas, such as that of Vogdes and Elder.[4]

W. H. Aldous
M. O. Valve Company
G.E.C. Research Laboratories
Wembley, England

[4] F. B. Vogdes and F. R. Elder, *Phys. Rev.*, vol. 24, p. 683, December, 1924.

Television Time-Base Linearization*

It is interesting to compare the treatment by J. Haantjes and F. Kerkhof of their television time base linearization problem[1] with the alternative analysis and solution of the same problem given by K. Schlesinger.[2]

I would point out that the correction circuits derived in the two papers are potentially equivalent and represent a class of networks capable of linearizing exponential waveshapes by the integration method. The latter appears to have been first used by G. Hawkins, whose British Patent Specification No. 511,600 (accepted 22/8/39) includes Schlesinger's circuit, which has been used consistently in Murphy home television receivers since before the war, together with two equivalents. I checked the performance of these networks while developing television receivers for R.F. Equipment Ltd., and derived a number of equivalents including that given by Haantjes and Kerkhof. A brief descriptive account of this work was accepted for publication (17/6/47) by *Electronic Engineering*, and remains to be published. In the meantime, I have noted the use of the last-mentioned circuit in a General Electric Co. (U.S.A.) television receiver.

I have drawn the correction networks of Haantjes and Kerkhof and Schlesinger in the accompanying figure for ease of comparison together with their RL equivalents; all are characterized by the voltage transfer operator

$$\frac{1 + \tau p}{1 + (1 + k)\tau p}$$

where the constants τ, k are defined for each network in Fig. 1.

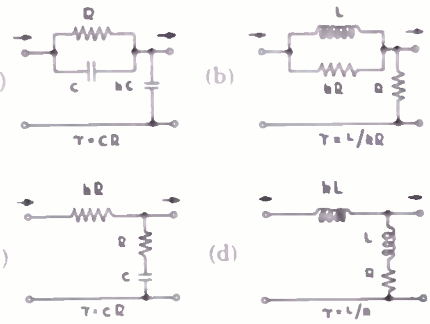


Fig. 1.—A group of linearity-correction networks:
(a) From Haantjes' and Kerkhof's paper.
(b) RL equivalent of (a).
(c) From Schlesinger's paper.
(d) RL equivalent of (c).

The following illustration of the integration method of linearization may be of interest to those who are not familiar with integral equations of the Volterra type as used by Schlesinger.

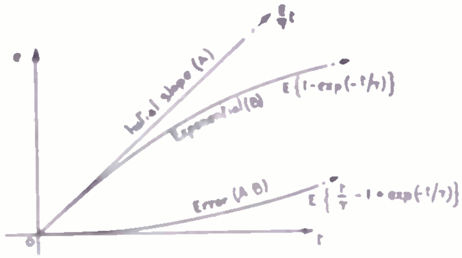


Fig. 2.

The voltage (see Fig. 2)

$$e(t) = E\{1 - \exp(-t/\tau)\} \quad (t \nless 0)$$

differs from the linear continuation of its initial slope

$$e'(0) = Et/\tau$$

by the quantity

$$E\{t/\tau - 1 + \exp(-t/\tau)\}$$
$$= E\{t^2/\tau^2\,\underline{|2} - t^3/\tau^3\,\underline{|3} + t^4/\tau^4\,\underline{|4} - \cdots\}$$
$$= E \cdot 1/\tau \cdot \int_0^t \{1 - \exp(-\lambda/\tau)\}\,d\lambda,$$

i.e., the required correction voltage is proportional to the integral of the voltage needing correction.

Finally, the method given by Schlesinger for correcting what he terms "acceleration" distortion due to the effect of resistance parallel to the deflector coil appears to be unnecessary, since the nonlinearity of the sawtooth generator output is usually of the "deceleration" type and is controllable by choice of the charging circuit constants. Expressed alternatively, a variable resistor parallel to the deflector coil may be used to compensate the exponential output of the sawtooth generator.

A. W. Keen
95, Pinner Rd.
Harrow, Middx.,
England

* Received by the Institute, May 3, 1948.

[1] J. Haantjes and F. Kerkhof, "Home projection television: Part III, Deflection circuits," Proc. IRE, vol. 36, pp. 407–412; March, 1948.

[2] Kurt Schlesinger, "Magnetic deflection of kinescopes," Proc. IRE, vol. 35, pp. 813–821; August, 1947.

Institute News and Radio Notes

West Coast Convention an Outstanding Success

At the West Coast Convention: Left to right, F. E. Terman, B. E. Shackelford, Harry Lubcke, Walter Kenworth, and Lloyd Sigmon

A total attendance of 7114 at the four-day IRE West Coast Convention made this one of the biggest electronic events ever held. Sponsored by IRE, it was held in Los Angeles, in conjunction with the AML and the WCEMA, all of which convened during the exhibit period, extending from September 30 through October 2. Third among American cities in IRE membership, Los Angeles has become a key center of electronics, owing both to the large number of radio manufacturers and scientists situated there, and to the U. S. Army's West Coast experiments with guided missiles and A-bomb research. Under the guidance of Walter Kenworth, Chairman, the Los Angeles Section acted as hosts to the IRE Convention.

In 1947, 3100 attended the Pacific Electronic Exhibit, and it is expected that at the fifth Exhibit, to be held in San Francisco in September, 1949, there will be an equally significant increase in attendance. Seventy-three members of the West Coast Electronic Manufacturers Association displayed outstanding developments in the electronic field. Among the new inventions of unusual interest were a hearing aid based on the proximity fuze which is one-quarter the size of those currently available, and printed circuits in which patterns traced in India ink take the place of wires.

Atomic bomb control was the principal topic of the convention. World-famous physicist Robert A. Millikan lectured on "The Release and Utilization of Atomic Energy." Because of the scarcity of fissionable material, he said, he does not believe that atomic power will become a major factor in world economy. "The greatest service of the atom bomb has been to make as clear as crystal the necessity for finding a substitute for war in international relations."

Benjamin E. Shackelford, former President of the IRE, attracted much public attention at a press conference, when he declared that the fear of Russian scientists to make mistakes would bring victory to the United States in the event of war. George W. Bailey, Executive Secretary of the Institute, also attended the Convention, as did Clinton B. DeSoto, Technical Editor of the PROCEEDINGS, and a host of other IRE notables.

Many papers on new developments in the radio, television, and electronic fields were read. One which caught a good deal of public interest was C. E. Nobles' speech on "stratovision." Westinghouse engineer Nobles told delegates that a coast to coast television network using airplanes to relay programs is now a technical reality.

Calendar of COMING EVENTS

AIEE-IRE-NBS Symposium on High-Frequency Measurements, Washington, D. C., Jan. 10–12.

American Physical Society Meeting, New York City, Jan. 27–29, 1949.

1949 IRE National Convention, New York City, March 7–10.

RMA—IRE Spring Meeting, Philadelphia, Pa., Apr., 25–27.

IRE Holds Rochester Fall Meeting

Attendance totals exceeded 900 at the highly successful twentieth Fall Meeting of The Institute of Radio Engineers and the Radio Manufacturers Association, which was held on November 8, 9, and 10, 1948, in Rochester, N. Y. Six technical sessions were presented, under the leadership of Benjamin E. Shackelford, 1948 President of the IRE; Stuart L. Bailey, 1949 President; David B. Smith and Dorman D. Israel, Institute Fellows; Oliver L. Angevine, Senior Member; and Kenneth J. Gardner, Chairman of the Rochester Section. E. Finley Carter presided over one of the general sessions, addressed by Kenneth W. Jarvis. Both are IRE Fellows. Arthur L. Schoen headed the photographic session.

The climax of the numerous social events was the Fall Meeting Dinner, at which B. DeForest Bayly acted as toastmaster, introducing RMA President Max F. Balcolm, who made the principal banquet address, and Dr. Shackelford and S. L. Bailey, who also spoke. Virgil M. Graham presented the annual RMA-IRE award to Dorman D. Israel, executive vice-president of the Emerson Radio and Phonograph Corporation, for his services in receiver standardization.

Mr. Graham has announced that the next engineering Fall Meeting, in 1949, will be held at Syracuse, N. Y.

Three Sections Meet in Olean

At a joint dinner meeting of the Buffalo, Emporium, and Rochester Sections held in Olean, N. Y., on October 14, 1948, Stuart L. Bailey, President Elect of the IRE, spoke on "Engineering Considerations in the Location of Broadcast Transmitters." Mr. Bailey was introduced by L. C. F. Horle, chief engineer of the RMA. Virgil M. Graham, Fellow of the Institute, was also present. The meeting was a result of the joint efforts of the three sections, directed by their chairmen: J. F. Myers, Buffalo; E. F. Kahl, Emporium, and K. J. Gardner, Rochester.

1949 National Convention News

"Radio Electronics—Servant of Mankind" will be the theme of the 1949 IRE Convention, because so much of modern life and living is based upon electronics that the radio art can be said to underlie present day civilization. The manufacturers' displays at Grand Central Palace will be grouped under the theme of "Spotlighting the New," and it is expected that many outstanding developments and products of postwar research in electronics will be on display for the first time to provide an exhibit of unusual interest to layman and engineer alike.

The technical program too will be of major significance. Among the numerous special features contemplated are symposia on nuclear science, navigation aids, and marketing. Efforts are being made to touch on every important aspect of the radio art.

Plans for the social events are proceeding apace. B. E. Shackelford will act as toastmaster at the President's Luncheon, and Delos W. Rentzel, Civil Aeronautics Administrator has been asked to speak at that event. Toastmaster at the Banquet will be Raymond Guy, and Karl Spangenberg will deliver the acceptance on behalf of the newly elected Fellows. B. E. Shackelford has been asked to take charge of Women's Activities, and Ivan S. Coggeshall will speak at the Annual Meeting of the Institute. Stanton Vanderbilt will head the Hotel Arrangements Committee.

Technical Committee Notes

The **Standards Committee** approved and published the following standards during 1948:

Standards on Antennas: Methods of Testing, 1948
Standards on Antennas, Modulation Systems, and Transmitters: Definitions of Terms, 1948
Standards on Abbreviations, Graphical Symbols, Letter Symbols, and Mathematical Signs, 1948
Standards on Television: Definitions of Terms, 1948
Standards on Television: Methods of Testing Television Receivers, 1948

Material submitted to the **Annual Review Committee** from all other technical committees concerning the developments in the field of radio engineering will be reviewed by this Committee and published in a forthcoming issue of the PROCEEDINGS. . . . The **Antennas Committee** is now preparing definitions of waveguide and transmission line terms for standardization. The Committee has initiated a petition for the formation of a Professional Group on Antennas and Wave Propagation. . . . The **Audio and Video Techniques Committee** at its December 8 meeting reviewed standard proposals prepared by subcommittees on definitions and methods of measurement and test. . . . Subcommittees in the New York, Philadelphia, Washington, D. C., and Boston areas are preparing definitions for terms relative to pulse-code modulation and modulation theory for the **Circuits Committee**. . . . The "Glossary of Acoustical Terms," which has been prepared by the American Standards Association in co-operation with the **Electroacoustics Committee**, is now being printed as an ASA Standard Proposal. Within the year a final revised and approved glossary will be available to IRE members. . . . The **Electron Tubes Committee** held a highly successful conference at Cornell University during June. It is planned to hold the 1949 conference at Princeton University and the 1950 conference at the University of Michigan. Proposed standards on testing gas tubes have been forwarded to the Standards Committee for approval. New subcommittees have been formed on semiconductor devices and solid-state devices. . . . The **Electronic Computers Committee** has prepared a computer bibliography which will be set up on punched cards at the Watson Laboratories. A file will also be kept at Headquarters. Work on proposed computer definitions has been started. . . . Among many other activities the **Industrial Electronics Committee** is preparing definitions for standardization in close co-operation with AIEE and NEMA. . . . The **Modulation Systems Committee** is preparing definitions in the fields of pulse code modulation. . . . Henri Busignies of FTL has recently taken over the chairmanship of the **Navigation Aids Committee**, which has planned an ambitious program. Standard material containing definitions is currently being considered by the Standards Committee. A symposium on Radio Aids to Navigation is planned for the 1949 IRE Convention. . . . The National Research Council is organizing a joint group to supervise a general glossary of nuclear terms. The **IRE Nuclear Studies Committee** is contributing to that work, principally in connection with electronics. The Committee in conjunction with the AIEE sponsored a Conference on Electronic Instrumentation in Nucleonics and Medicine in New York City on November 29 and 30 and December 1, 1948. It plans to hold a symposium on Nuclear Science during the March Convention. Papers solicited by this Committee are currently appearing in the PROCEEDINGS, in a series which will run through 1949. Other nuclear papers are being solicited for possible publication in monograph form. . . . The **Radio Receivers Committee** has revised the 1938 standards on methods of testing AM broadcast receivers and will publish the new standard at an early date. . . . Although several subcommittees have been doing active work, the **Radio Transmitters Committee** has not yet been called upon to review any material. . . . The **Railroad and Vehicular Communications Committee** is preparing work on definitions and methods of testing for standardization. Part of this work has been submitted to the Standards Committee for approval. . . . The new Chairman of the **Symbols Committee**, A. F. Pomeroy, reports that at the last meeting, on November 19, 1948, four task groups were authorized, (1) Editing Proposal for Reference Designations, (2) Graphical Symbols for Semiconductors, (3) Graphical Symbols for Single-Line Diagrams, (4) Changes and Additions to Standard Number 10 on Symbols (excluding graphical symbols for semiconductors and for single-line diagrams). Reference designations are combinations of letters and numbers used to identify objects, or circuits or portions thereof, or schematics and parts lists. Semiconductors refer to crystal detectors, metallic rectifiers, and transistors. Graphical symbols for single-line diagrams are already standardized for coaxial and waveguide circuits. The proposal of the new task group is to work with a group from RMA to cover applications to all types of circuits. . . . The scope of the **Television Systems Committee** has been altered to encompass matters of a systems nature. Work will continue on definitions, but the work of methods of testing components will be turned over to appropriate committees in line with the revised scope. The Committee has actively assisted JTAC in gathering information for its reports to the FCC. . . . The **Wave Propagation Committee** has formed subcommittees on Standards and Practices, Theory and Application of Tropospheric Propagation, Theory and Application of Ionospheric Propagation, and on Publication. The subcommittee on Publication will have proposed definitions completed this year. The Committee is assisting JTAC and the Television Systems Committee in gathering information for its second report to the FCC.

Industrial Engineering Notes[1]

British Television in Denmark

The British Radio Industry Council, which met at Copenhagen, Denmark, from September 16 to 21, 1948, has furnished the following technical description of its television equipment there:

Camera

This is a new turret camera carrying four lenses. These lenses can be fitted in any order and can have local lengths of $1\frac{1}{2}$ to 20 inches.

Since the focus control operates over the same arc for all lenses, the focus range over which the image tube moves is adjusted automatically. The turret is rotated from the rear and the lens aperture is adjusted from the same position. Both the aperture and the lens in use are registered through for the operator's information.

The viewfinder is electronic and uses a special crt. The picture pick-up tube is of the supericonoscope type recently perfected to operate over large variations in light intensity.

The camera also has a self-contained time base giving accurate geometric cutouts.

A talk-back system is incorporated to assist instruction between operator and the control van.

The Camera Control Unit

This unit supplies the camera with sweep wave forms, suppression pulses, etc., developed by special circuits from the master control pulses received from the wave-form generator. The pulses are selected and shaped so that the variation in delay time caused by using camera cables of up to 1,000 feet in length, are automatically compensated for.

The camera control unit also supplies all voltages for the camera image tube, the high voltages being derived from pulse-operated circuits. The various other circuits are supplied by either current or voltage-stabilized power from the camera control unit.

The camera video output voltages are fed to the camera control unit via the camera cable over a balanced and screened pair of wires to minimize extraneous cross talk and interference to the low voltages involved. In the control unit, the video voltages are amplified and frequency corrected and then, after shading injection, are further amplified by the main video amplifier chain with a frequency response flat to 5 Mc and fed into the mixing channel where dc restoration and blanking insertion are carried out. The video wave forms are then split to feed the various outputs, such as video, without synchronizing pulses—video with sync pulses—and balanced to feed via the camera cable and the camera electronic viewfinder. Provision is made so that an artificial test signal supplied by the wave-form generator can be mixed with the sync and suppression pulses and fed out to the transmitter while a camera picture is set up on the camera control unit monitor tube. The front panel controls are reduced to the minimum to allow ease of adjustment and efficiency. The spacing and layout making for the most comfortable operation, the shading wave forms are under immediate control although the operation is simplified by the use of the new camera tube.

The unit also houses an electronic peak voltmeter to enable the operator to monitor and control the voltage fed to the transmitter.

A microphone is fitted as an integral part of the control unit and this is connected into the talk-back and monitoring circuits. Headphones are plugged into a socket on the front panel.

Associated with the camera control unit is the control unit power pack, a unit which supplies all the low voltages for the control unit and camera and a monitor unit. This monitor houses a 9-inch crt on which the operator sees the output from his camera control unit as a complete television picture, while a smaller crt continuously monitors the output voltage wave form. The two tubes in conjunction allow the operator to set the appropriate controls to keep the outgoing television picture at the peak of clarity at all times.

Camera Mixer Unit

The camera mixer unit provides for 3 main requirements:

(a) For the mixing of pictures supplied from three cameras, which are known as the

[1] The data on which these NOTES are based was selected, by permission, from "Industry Reports," issues of October 22 and 29 and November 5 and 12, from the Canadian RMA "News" of October 26, and from a British Radio Industry report. We hereby gladly acknowledge all of these organizations' helpful attitude in this matter.

sync group. This may be done in two ways as follows:

(1) The camera pictures may be "faded up" separately or at the same time. The fading action is used where an instantaneous change from one camera to another is not desirable. Outside broadcasts from theaters make much use of this requirement. Ghost effects can be readily made by superimposing 1 camera picture on another.

(2) The camera pictures may be "cut up." This means an instantaneous change from 1 camera to another may be made. This action is used with good effect in outside broadcasts of sporting events where speed of change of camera is an outstanding requirement.

(b) For the acceptance and re-transmission of a television signal from a remote source which is not in synchronism with requirement (a); i.e., another outside broadcast or film-scanner equipment, etc. This is known as the nonsync group.

(c) For the switching in and transmission of a test signal provided by the wave-form generator for testing purposes or in the event of a breakdown of certain parts of the equipment.

The test signals which are used are described under the wave-form generator.

Picture and Wave-Form Monitor

Two types of monitor are available, one for use in conjunction with a camera mixer unit and the one for operation with a camera control unit.

They provide the operator of the respective camera unit with a picture of the outgoing signal on a $7\frac{1}{2}\times 6$-inch screen, while simultaneous operation of a 3-inch crt facilitates the monitoring of either the line or frame component of the wave form.

These monitors have been designed to ensure that high-quality pictures are obtained, thus providing a constant check on the fidelity of the outgoing signal and enabling the operator to make the necessary adjustments to the camera unit under his control.

Controls are provided on the front panels for the independent adjustment of brilliancy, focus, and video input to both tubes, while a key switch enables the selection of either line- or frame-component analysis on the 3-inch tube.

These units are entirely self-contained, a stabilized power pack, and extra-high-voltage generator being incorporated in the monitor.

Television Wave-Form Generator

The unit provides accurate and precise wave forms, completely automatically, independent of all conditions of operation. It provides the following signals for the camera chain system:

(1) Complete frame- and line-synchronizing pulses to produce the standard wave form transmitted by the British Broadcasting Corp.

(2) A black cross on a white background, for use as test signal, or as an interval signal.

(3) A sawtooth wave form for testing the complete transmitting chain.

(4) Provisions for a monoscope picture, as a test signal or for use during intervals.

(5) Frame- and line-synchronizing signals for driving the camera chains.

The unit incorporates the latest circuit improvements, designed to provide the most accurate wave form. Each line pulse width is determined by a passive circuit, and each frame pulse width is held accurately by a new type of counter circuit. The main divider, for reducing the line frequency to frame, consists of a binary counter embodying many new features. The frame frequency is held to the frequency of the main power supply by a correction motor driving the master oscillator capacitor.

The unit has a built-in oscilloscope for monitoring outgoing wave forms, which may also be used for testing purposes. It does not require an operator, as all the pulse widths and wave-form arrangements are fixed.

Commentator's Monitor

The timebase circuits on the monitors give good geometrical linearity with an rf unit of wide bandwidth. For improved picture quality, the video bandwidth is flat to 4 Mc. Sound response is flat from 25 cps to 10 kc.

The unit receives picture signals from the camera via cable from 0.5 to 1.5 volts peak-to-peak, or rf signals from the transmitter. The output from the video and rf channels can be changed by a 3-position switch on the back of the rf chassis to give rf, video, and video terminated with 75 ohms.

The sound to this unit is fed via a balanced cable through a transformer to the gain control which operates on the grid of the output tube.

Vision Transmitter

The vision transmitter is designed to transmit television signals at a maximum power of 25 watts at peak white level. It is fed with the video-frequency wave form from the camera mixer unit containing both picture information and synchronizing pulses.

This unit separates the synchronizing pulses from the complete wave form and converts them into clamp pulses for clamping the "black level" in the rf output tube to a reference potential; this prevents any crushing of the transmitted sync pulses or picture information. The separated sync pulses are also used to switch off the rf driving tube during the sync period to ensure that there is no transmission during that period.

This unit includes a built-in oscilloscope in order that the operator may see the transmitted rf envelope while transmission is taking place, and also for use when testing is being carried out.

There is a meter which, by turning a switch, shows the amount of cathode current which is being taken by certain tubes. Another position of the switch makes the meter into a triple-range voltmeter for testing purposes.

Sound Transmitter

The sound transmitter is designed to transmit audio signals at a maximum peak power of 20 watts. It has wide af bandwidth with a flat response from 30 cps to 15 kc.

It contains an oscilloscope for setting up modulation depth, etc. and for the operator to watch during transmission, and for test purposes.

There is a meter which is used for checking cathode current as in the vision transmitter, which can be used in the same way as a voltmeter.

Sound System

This system consists of 2 main parts:

(a) The program section.

(b) The talk-back or intercommunication section.

The program section is that part of the apparatus which from low-level inputs supplies an output of required program content suitable for transmitter modulation.

The talk-back section provides general instructions from the producer to the operators of both sound and vision equipment and also to artists in a stage rehearsal. Telephone conversation can also be obtained between camera control unit and camera by isolating this part of the circuit from the producer's line.

(a) *Program Section*

Provision is made for eight sound channels by the use of two mixer units (MX/18). Six of these channels are available for microphones for general program requirements while one microphone is for the use of the commentator. The final channel is fed by a pickup, the output of which is attenuated to a suitable level before being fed to the mixer.

In each mixer, the various channel signals are suitably faded and mixed together according to program requirements and the outputs are fed to a high-quality amplifier (OBA/8). The amplified signal is then supplied for modulating the transmitter.

A loudspeaker unit in the van is used for monitoring the sound and is fed also from the output of the amplifier. A peak program meter is incorporated in the amplifier and this gives an indication of the approximate power sent to the transmitter.

(b) *Talk-back section*

A high-quality microphone is installed on the producer's desk, and this feeds into an amplifier OBA/8. The output is connected to all points that require instruction from the producer during both setting up and operating, and also to a switched external loudspeaker unit which is used for instruction to artists during rehearsals.

When the external loudspeaker unit is switched into the circuit, the loudspeaker unit in the van is disconnected by the same key to avoid acoustical oscillation in the system as the latter unit is in close proximity with the producer's microphone.

For local instruction between camera control unit and camera, a key on the camera control unit is operated and this disconnects these units from the talk-back circuit as a whole and allows two-way conversation.

An auxiliary system of standard field telephones allows conversation between the producer and commentator, van transmitter operators, and set operators.

Film Scanner

The film scanner enables any normal 35 mm cinematograph film to be transmitted by television and received in the home as part of a television service. Special prints of the film are not necessary.

The equipment in use at Copenhagen is provided by EMI. The type of film transmission used is new. It gives far better results than the older method, is simpler to operate and the BBC is, at present, negotiating for the supply of film transmitters working on new principle.

The equipment comprises a wave form generator and a film channel, which are two of the units from which EMI television transmitting stations can be built. up The units are built for permanent installation at a fixed station and compactness has been sacrificed to ensure robustness and ease of maintenance.

The film scanner works on the "flying-spot" principle. An intense but finely focussed spot of light is moved or "scanned" in a regular pattern of line over the face of a special crt. An image of the light spot is focussed by a lens on to the film, which is moved continuously, and not, as in an ordinary projector, intermittently. The light passing through the film depends upon the density of the film at the place where the light is focussed. This transmitted light is collected by an electron multiplier photo cell of very high sensitivity. The output of the photo cell is thus caused to vary with the light and shade in the picture on the film. The photo-cell output is amplified and synchronizing signals are added to form the complete television signal, ready to feed the transmitter.

The equipment is in the charge of a single operator, who sits at a desk which carries all the necessary controls, including a "picture monitor."

The equipment can be adjusted to operate at any of the standards adopted in Europe or the United States, up to 605 lines. It is, at present, working on the British standard of 405 lines, 50 frames interlaced.

Synthetic Quartz Production Now Feasible

In September, 1949, the largest known single crystal of synthetic quartz essentially free from defects was delivered to the U. S. Signal Corps' Engineering Laboratories. This crystal, one of the synthetic quartz crystals of high-grade quality which several commercial firms and colleges have been producing under research contracts with the Army, is about the diameter of a silver dollar.

Although the mass production of synthetic quartz crystals may now be feasible, they cannot yet be manufactured on an economical basis. Once optimum manufacturing procedures have been established, it is possible that synthetic quartz can compete with the natural product.

Microwave Measurement Standards Developed

As part of a broad program for the establishment of national standards and calibration services for all electrical quantities at radio frequencies, the U. S. Bureau of Standards is developing microwave measurement standards in the range from 300 to 100,000 Mc and above. This work has resulted not only in extremely precise and accurate standards of frequency, power, attenuation, and other quantities, but has also made possible precision measurements in a whole new field of microwave spectroscopy formerly inaccessible to investigation because of the limitations of infrared and optical equipment.

Of basic importance in the microwave program has been the development and continued improvement of a primary standard of frequency accurate to one part in a hundred million. Further problems involved in the development of microwave standards of power, attenuation, impedance, field intensity, and noise, and of dielectric and magnetic measurement, are also being attacked. A detailed description of the Bureau's research in microwave measurements has been published in the December issue of "The Technical News Bulletin," which may be obtained from the Superintendent of Documents, Washington 25, D. C., for 10 cents.

Signal Corps Presents Exhibit to Smithsonian

Major General Spencer B. Akin, Chief Signal Officer of the U. S. Army, presented a Signal Corps exhibit to Dr. Alexander Wetmore, Secretary of the Smithsonian Institution, last fall. How radar reached the moon is one of the displays, as well as the use of microwave radio relay communication, as shown in a miniature battle scene from World War II.

Telemetering Systems Report

The basic problems of designing telemetering systems are discussed in a report released by the Office of Technical Services, U. S. Department of Commerce. Prepared by the Signal Corps Engineering Laboratories, the report (PB93938) is available in photostat form for $10, or in microfilm for $3.50, from the Library of Congress, Photoduplication Service, Publication Board Project, Washington 25, D. C.

Tin Allocations Frozen

Although the rising production of television receivers by radio manufacturers has increased the industry's demand for tin allocations, these allocations are still covered by a materials order which has been in effect since 1942 and which will continue until June 30, 1949. Despite the increase in the over-all supply of tin, industrial users are being allocated no more than they were during 1947, according to officials of the Department of Commerce. Any surplus is being held for strategic stockpiling. Exceptions, when allowed, are made on an individual basis following a presentation of complete information showing a need for increased allotments.

Government Studies Reduced Distribution Costs

"How Manufacturers Reduce Their Distribution Costs," a 150-page study of the sales practices of a representative group of 45 different producers who have effectively reduced their marketing expenses and at this time substantially increased their over-all profits, has been published by the Department of Commerce, and may be obtained from the Superintendent of Documents, Government Printing Office, Washington 25, D. C., for 35 cents each copy. Case histories summarized in the booklet show that "selective distribution," rather than attempts to achieve "100 per cent coverage" is the key not only to reduced distribution costs but also to increased sales.

FCC Activities

A change in FCC administrative procedure to expedite the handling of requests from FM broadcast stations to transmit multiplex facsimile has been made. The Commission delegated authority to act on this type of application to its secretary. . . . The FCC has published its **objection to Mexico's proposed use of the 540-kc frequency** for a standard broadcast station. Although the 540-kc frequency is not presently a part of the standard broadcast band, under agreements reached at the International Radio Conference in Atlantic City in 1947 it was made available for such use, subject to agreement among the interested countries. Mexico has agreed to defer its plans for use of the controversial frequency, and the subject is expected to be one of the major items under consideration at the next North American Regional Broadcasting Conference, to be held in Canada in September, 1949. . . . The Commission has announced a postponement of the effective date of two of its rulings. The **regulations governing welding** devices using radio-frequency energy have been deferred until April 30, 1939, following petitions filed by users of this type of equipment asking for additional time to solve engineering problems involved in bringing the devices in conformity with FCC rules. . . . The Commission has also extended its waiver of the provision requiring a person to hold **a radio operator license in order to operate radar stations aboard ships** until April 15, 1949. This relaxation of FCC rules does not permit unlicensed personnel to make adjustments or do any servicing or maintenance to ship radar stations. . . . Complete tabulations showing **standard broadcast employment and compensation** for seven networks and 1,260 stations for a sample week in October, 1947, have been released.

Export Forms Revised

A revision of the Shipper's Export Declaration Form "in the interest of more effective export control" was announced by the Office of International Trade, U. S. Department of Commerce. In the revised form, the exporter is defined as the actual seller of the materials being shipped and as the person who holds an OIT export license, where it is required, to replace the looser usage in the old form, in which the word "exporter" was often interpreted to mean the forwarding agent. Supplies of the new form may be obtained at 50 cents per pad of 100 sheets from the Superintendent of Documents, Washington 25, D. C., as well as from Customs Offices and from field offices of the Department of Commerce.

FM News

Under its newly adopted rules providing for low-power FM educational outlets, the FCC has granted its first noncommercial educational FM station construction permit to Syracuse University, which was authorized to construct an FM station with a power output of 2.5 watts. Estimated cost of equipment is $2,088, exclusive of studios and equipment already installed for radio training.

FM stations now in operation number

689, including 25 noncommercial educational outlets. New FM stations began operations in the following states:

Cal., Alameda (KONG) and Modesto (KTRB-FM); *Conn.*, Greenwich (WGCH-FM); *Fla.*, Jacksonville (WJAX-FM) and Miami Beach (WLRD); *Ga.*, Macon (WNEX-FM); *Ill.*, Centralia (WCNT-FM) and Chicago (WCFL); *Iowa*, Clinton (KROS-FM); *Ind.*, Terre Haute (WTHI-FM); *Ky.*, Owensboro (WOMI-FM); *Mass.*, Boston (WCOP-FM) and North Adams (WMFM); *Md.*, Cumberland (WCUM-FM); *N. J.*, Newark (WNJR-FM); *N. Y.*, Lockport (WUSJ); and *Ohio*, Cleveland (WERE-FM).

Television Developments Here and Abroad

Although Chairman Wayne Coy of the FCC told the Radio Executives Club of New York that "a thousand television stations within the next seven or eight years is altogether reasonable," nevertheless the FCC "freeze order" continues in force and no more applications for television stations are being approved, although applicants with construction permits are authorized to proceed with the construction and opening of their stations. At present there are 45 commercial television stations on the air and 77 construction permits outstanding. Five new stations have begun operations recently: WAVE at Louisville, Ky.; WAAM at Baltimore, Md.; WJBK-TV and WXYX, at Detroit, Mich.; and WNBK at Cleveland, Ohio.

The FCC has published a television channel study showing the effects of ground wave and tropospheric interference on representative service areas of the television stations allocated under the FCC allocation plan.

Britain and Argentina are both developing their television industries rapidly. Two British radio manufacturers have publicly announced their intention of producing television receivers in Canada. Cosser of Canada Ltd., is making arrangements for a plant at Halifax, and Pye Radio plans to manufacture television sets at Ajax, Ont.

A national television industry to be run by Argentine specialists is the goal of the Argentine government. At the start, however, French engineers, under the leadership of Rene Barthelemy, will direct the laboratories and train their personnel.

Television and Radio Set Production Rise

Television receiver production soared to a new monthly peak in September, as RMA member-companies reported manufacturing 88,195 television sets, thus bringing their output for the first three-quarters of 1948 to 488,133. The average weekly output during the third quarter of 1948 was 50 per cent greater than it was during the first half of the year, while the September weekly average was 64 per cent greater than the weekly average for the same six-month period.

FM-AM radio set production also reached a new high of 171,573 in September, and brought the year's record to more than a million and close to the total output of this type of receiver in 1947. General radio receiver production recovered from a seasonal summer decline to reach 1,192,251 in September and bring the first three-quarters of 1948's total output by RMA set manufacturers to more than 10,000,000.

In Canada, also, radio production rose in September. A total of 76,623 units with a list price value of $5,983,635 made this the highest September on record for the Canadian radio industry, and the fourth highest for any month on record in dollars (only the last three months in 1947 exceeded these figures) and the fifth in units (exceeded by the three months mentioned, plus December 1946).

Radio Industry Mobilizes

Present plans of the Office of Civil Defense Planning, National Military Establishment, call for the utilization of existing communications and radio equipment as far as possible in the event of an emergency, according to a report submitted by the agency to Defense Secretary Forrestal. Under the plan, local radio and communications agencies will procure, with the aid of the military, any needed radio equipments.

Meanwhile, the radio industry itself is getting ready, so that, if war should come, the industry will not be unprepared. The following 27 radio industry representatives were selected by officials of the Munitions Board and the NSRB without any recommendations from the RMA, in accordance with government policy not to consult trade associations on such appointments:

W. R. G. Baker, General Electric; Max F. Balcolm, Sylvania Electric Products; John Ballantine, Philco; F. D. Bliley, Bliley Electric Co.; Monte Cohen, F. W. Sickles Co.; R. O. Driver, Wilbur H. Driver Co.; H. A. Ehle, International Resistance Co.; Ray C. Ellis, Raytheon; Walter Evans, Westinghouse; Frank M. Folsom, RCA Victor; Paul V. Galvin, Motorola; G. M. Gardner, Wells-Gardner Co.; S. A. Gilfillan, Gilfillan Bros.; W. J. Halligan, Hallicrafters; W. P. Hilliard, Bendix; H. L. Hoffman, Hoffman Radio Corp.; Paul J. Kruesi, American Lava; F. R. Lack, Western Electric; E. H. Locke, General Radio; W. A. MacDonald, Hazeltine Electronics; J. H. Miller, Weston Electrical Instruments; D. R. G. Palmer, General Cable; A. D. Plamondon, Jr., Indiana Steel Products; R. C. Sprague, Sprague Electric; and C. A. Warde, Jr., Superior Tube.

Leighton H. Peebles, who served during part of 1948 as communications consultant to the NSRB, was elevated to the post of Assistant Director of Manufacturing, U. S. Office of Production. There he will direct eight NSRB divisions, representing eight fields of industrial production, which include radio and communications.

An RMA task committee, headed by Ray C. Ellis, drew up a 16-page report which is now being used by Mr. Peebles as an example for other industries to follow. Its recommendations are divided into four sections: The philosophy underlying the development of an electronic division of a production co-ordinating agency in the event of mobilization; a general limitation order for the electronics industry in the event of mobilization; the discussion of controlled materials plan and priorities regulations; and suggested government organizations.

Books

Radio Industry Red Book Replacement Parts Buyers Guide, edited by E. L. Jordan

Published (1948) by Howard W. Sams and Co., Inc., Indianapolis 7, Ind., 446 pages, 8½×11.

This is a single volume which lists the proper radio replacement parts used during the past ten years, for the convenience of the radio service technician. Besides containing data on the replacement parts of 17 manufacturers, it has material on 9 major replacements components. Tubes and dial lights; capacitors; transformers; controls; if coils; vibrators; speakers; batteries and phonograph chartridges; if peaks; and installation and servicing methods are all covered.

Publication News

The National Association of Corrosion Engineers, 905 Southern Standard Building, Houston 2, Tex., has recently put out two new volumes of interest to radio engineers: *A Bibliographic Survey of Corrosion—1945*, by R. D. Misch, J. T. Weber, and H. J. McDonald; and the *Directory of the American Co-ordinating Committee on Corrosion.* The 129-page bibliography contains approximately 1100 references to published articles relating to corrosion, and 170 references to patents, all of which appeared from January, 1945, to February, 1946, inclusive, in *Chemical Abstracts, Corrosion, Corrosion and Material Protection, Engineering Index, Industrial Arts Index*, and *Metals Review*. The book is $4.00 to NACE members, $5.00 to others. . . . The directory, which is the fourth edition of a work actually published by the American Co-ordinating Committee on Corrosion but distributed by the NACE, lists in its 62 pages the names, addresses, and fields of special endeavor of many of the principal corrosion workers in the United States and Canada. There is a cross-index system so arranged that the name and address of a specialist in some particular phase of corrosion may readily be found. . . . The first volume of *Advances in Electronics*, edited by L. Marton of the National Bureau of Standards, has recently been published by the Academic Press, Inc., 125 E. 23 St., New York 10, N. Y. In view of the steadily increasing number of publications in the broad field of electronics, this yearly publication, devoted to reviews of specific topics in physical and engineering electronics, has been planned for research workers who wish to keep abreast of developments in the various branches of the subject. . . . The British Institute of Electrical Engineers in London has announced that the subscription rates for complete sets of their *Proceedings*, Parts I, II, and III, will now be £3. 13s. 6d., instead of £3. 3s., thus bringing the half-price cost to members of the Institute to £1. 16s. 9d. The subscription rate for each of the separate parts will remain the same, only the subscription rate for the whole having been altered. Changes in the *Proceedings* itself will include the omission from Part I of any material meant especially for the use of members, this being scheduled to appear in a new monthly *Journal* for members only. Part I will be issued in alternate months instead of monthly.

Sections

Chairman		Secretary
V. A. Edson eorgia School of Tech. tlanta, Ga.	ATLANTA January 21	M. S. Alexander 2289 Memorial Dr., S.E. Atlanta, Ga.
	BALTIMORE	J. W. Hammond 4 Alabama Ct. Baltimore 28, Md.
ohn Petkovsek 015 Ave. E eaumont, Texas	BEAUMONT—PORT ARTHUR	C. E. Laughlin 1292 Liberty Beaumont, Texas
. W. Hickman ruft Laboratory arvard University ambridge, Mass.	BOSTON	A. F. Coleman Mass. Inst. of Technology 77 Massachusetts Ave. Cambridge, Mass.
. E. Van Spankeren an Martin 379 uenos Aires, Arg.	BUENOS AIRES	A. C. Cambre San Martin 379 Buenos Aires, Arg.
. F. Myers 49 Linwood Ave. Buffalo 9, N. Y.	BUFFALO-NIAGARA January 19	R. F. Blinzler 558 Crescent Ave. Buffalo 14, N. Y.
. P. Hixenbaugh Radio Station WMT Cedar Rapids, Iowa	CEDAR RAPIDS	V. R. Hudek Collins Radio Co. Cedar Rapids, Iowa
K. W. Jarvis 5058 W. Fullerton Ave. Chicago 39, Ill.	CHICAGO January 21	Kipling Adams General Radio Co. 920 S. Michigan Ave. Chicago 5, Ill.
C. K. Gieringer 3016 Lischer Ave. Cincinnati, Ohio	CINCINNATI January 18	F. W. King RR 9 Box 263 College Hill Cincinnati 24, Ohio
F. B. Schramm 2403 Channing Way Cleveland 18, Ohio	CLEVELAND January 27	J. B. Epperson Box 228 Berea, Ohio
Warren Bauer 376 Crestview Rd. Columbus 2, Ohio	COLUMBUS February 11	George Mueller Electrical Eng. Dept. Ohio State University Columbus, Ohio
S. E. Warner Aircraft Electronics Assoc. 1031 New Britain Ave. Hartford 10, Conn.	CONNECTICUT VALLEY January 20	H. L. Krauss Dunham Laboratory Yale University New Haven, Conn.
J. G. Rountree 4333 South Western Blvd. Dallas 5, Texas	DALLAS-FT. WORTH	J. H. Homsy Box 5238 Dallas, Texas
George Rappaport 132 East Court Harshman Homes Dayton 3, Ohio	DAYTON January 20	C. J. Marshall 1 Twain Place Dayton 10, Ohio
C. F. Quentin Radio Station KRNT Des Moines 4, Iowa	DES MOINES-AMES	F. E. Bartlett Radio Station KSO Old Colony Bldg. Des Moines 9, Iowa
A. Friedenthal 5396 Oregon Detroit 4, Mich.	DETROIT January 21	N. C. Fisk 3005 W. Chicago Ave. Detroit 6, Mich.
E. F. Kahl Sylvania Electric Products Emporium, Pa.	EMPORIUM	R. W. Slinkman Sylvania Electric Products Emporium, Pa.
W. H. Carter 1309 Marshall Ave. Houston 6, Texas	HOUSTON	J. C. Robinson 1422 San Jacinto St. Houston 2, Texas
R. E. McCormick 3466 Carrollton Ave. Indianapolis, Ind.	INDIANAPOLIS	Eugene Pulliam 931 N. Parker Ave. Indianapolis, Ind.
Karl Troeglen KCMO Broadcasting Co. Commerce Bldg. Kansas City 6, Mo.	KANSAS CITY	Mrs. G. L. Curtis 6005 El Monte Mission, Kan.
R. W Wilton 71 Carling St. London, Ont., Canada	LONDON, ONTARIO	G. H. Hadden 35 Becher St. London, Ont., Canada
Walter Kenworth 1427 Lafayette St. San Gabriel, Calif.	LOS ANGELES January 18	R. A. Monfort L. A. Times 202 W. First St. Los Angeles 12, Calif.
O. W. Towner Radio Station WHAS Third & Liberty Louisville, Ky.	LOUISVILLE	D. C. Summerford Radio Station WKLO Henry Clay Hotel Louisville, Ky.
F. J. Van Zeeland Milwaukee School of Eng. 1020 N. Broadway Milwaukee, Wis.	MILWAUKEE	H. F. Loeffler Wisconsin Telephone Co. 722 N. Broadway Milwaukee 1, Wis.
K. R. Patrick RCA Victor Div. 1001 Lenoir St. Montreal, Canada	MONTREAL, QUEBEC February 9	S. F. Knights Canadian Marconi Co. P.O. Box 1690 Montreal, P. Q., Canada
L. A. Hopkins, Jr. 1711 17th Loop Sandia Base Branch Albuquerque, N. M.	NEW MEXICO	T. S. Church 637 La Vega Rd. Albuquerque, N. M.
J. W. McRae Bell Telephone Labs. Murray Hill, N. J.	NEW YORK February 2	R. D. Chipp DuMont Telev. Lab. 515 Madison Ave. New York, N. Y.
C. G. Brennecke Dept. of Electrical Eng. North Carolina State College Raleigh, N. C.	NORTH CAROLINA-VIRGINIA	C. M. Smith Radio Station WMIT Winston-Salem, N. C.
W. L. Haney 117 Bourque St. Hull, P. Q.	OTTAWA, ONTARIO January 20	G. A. Davis 78 Holland Ave. Ottawa, Canada
A. N. Curtiss Radio Corp. of America Camden, N. J.	PHILADELPHIA February 3	C. A. Gunther Radio Corp. of America Front & Cooper Sts. Camden, N. J.
M. A. Schultz 635 Cascade Rd. Forest Hills Borough Pittsburgh, Pa.	PITTSBURGH February 14	E. W. Marlowe Union Switch & Sig. Co. Swissvale P.O. Pittsburgh 18, Pa.
O. A. Steele 1506 S.W. Montgomery St. Portland 1, Ore.	PORTLAND	F. E. Miller 3122 S.E. 73 Ave. Portland 6, Ore.
A. V. Bedford RCA Laboratories Princeton, N. J.	PRINCETON	L. J. Giacoletto 9 Villa Pl. Eatontown, N. J.
K. J. Gardner 111 East Ave. Rochester 4, N. Y.	ROCHESTER January 20	Gerrard Mountjoy Stromberg-Carlson Co. 100 Carlton Rd. Rochester, N. Y.
E. S. Naschke 1073-57 St. Sacramento 16, Calif.	SACRAMENTO	W. F. Koch 1340 33rd St. Sacramento 14, Calif.
G. M. Cummings 7200 Delta Ave. Richmond Height 17, Mo.	ST. LOUIS	C. E. Harrison 818 S. Kings Highway Blvd. St. Louis 10, Mo.
C. L. Jeffers Radio Station WOAI 1031 Navarro St. San Antonio, Texas	SAN ANTONIO	H. G. Campbell 233 Lotus Ave. San Antonio 3, Texas
C. N. Tirrell U. S. Navy Electronics Lab. San Diego 52, Calif.	SAN DIEGO February 1	S. H. Sessions U. S. Navy Electronics Lab. San Diego 52, Calif.
F. R. Brace 955 Jones St. San Francisco 9, Calif.	SAN FRANCISCO	R. A. Isberg Radio Station KRON 901 Mission St. San Francisco 19, Calif.
W. R. Hill University of Washington Seattle 5, Wash.	SEATTLE February 10	W. R. Triplett 3840—44 Ave. S.W. Seattle 6, Wash.
F. M. Deerhake 600 Oakwood St. Fayetteville, N. Y.	SYRACUSE	S. E. Clements Dept. of Electrical Eng. Syracuse University Syracuse 10, N. Y.
A. R. Bitter 4292 Monroe St. Toledo 6, Ohio	TOLEDO	J. K. Beins 435 Kenilworth Ave. Toledo 10, Ohio

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G. P. Adair 1833 "M" St. N.W. Washington, D. C.	Washington	H. W. Wells Dept. of Terrestrial Magnetism Carnegie Inst. of Washington Washington, D. C.
J. C. Starks Box 307 Sunbury, Pa.	Williamsport	R. G. Petts Sylvania Electric Products, Inc. 1004 Cherry St. Montoursville, Pa.

SUBSECTIONS

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E. Olson 162 Haddon Ave., N Hamilton, Ont., Canada	Hamilton (Toronto Subsection)	E. Ruse 195 Ferguson Ave., S. Hamilton, Ont., Canada
A. M. Glover RCA Victor Div. Lancaster, Pa.	Lancaster (Philadelphia Subsection)	C. E. Burnett RCA Victor Div. Lancaster, Pa.
H. A. Wheeler Wheeler Laboratories 259-09 Northern Blvd. Great Neck, L. I., N. Y.	Long Island (New York Subsection)	M. Lebenbaum Airborne Inst. Lab. 160 Old Country Rd. Box 111 Mineola, L. I., N. Y.
L. E. Hunt Bell Telephone Labs. Deal, N. J.	Monmouth (New York Subsection)	G. E. Reynolds, Jr. Electronics Associates, Inc. Long Branch, N. J.
J. B. Minter Box 1 Boonton, N. J.	Northern N. J. (New York Subsection)	A. W. Parkes, Jr. 47 Cobb Rd. Mountain Lakes, N. J.
A. R. Kahn Electro-Voice, Inc. Buchanan, Mich.	South Bend (Chicago Subsection) January 20	A. M. Wiggins Electro-Voice, Inc. Buchanan, Mich.
R. M. Wainwright Elec. Eng. Department University of Illinois Urbana, Illinois	Urbana (Chicago Subsection)	M. H. Crothers Elec. Eng. Department University of Illinois Urbana, Illinois
S. S. Stevens Trans Canada Airlines Box 2973 Winnipeg, Manit., Canada	Winnipeg (Toronto Subsection)	S. G. L. Horner Hudson's Bay Co. Brandon Ave. Winnipeg, Manit., Canada

IRE People

Harold A. Zahl (A'39–SM'46), holder of the Legion of Merit for contributions made in the fields of radar and vacuum tubes, was appointed director of research of the Army Signal Corps Engineering laboratories at Fort Monmouth, N. J., one of the top-flight civilian posts in the U. S. Army.

Born in Chattsworth, Ill., on August 24, 1904, Dr. Zahl attended primary and secondary schools in Porterville, Calif. He received the B.A. degree in physics and mathematics from North Central College in 1928, and the M.S. and Ph.D. degrees in physics and mathematics from the State University of Iowa in 1929 and 1931, respectively.

In the latter year he accepted an appointment as physicist with the Signal Corps Laboratories at Fort Monmouth, N. J. There until 1942 he participated in many research and development projects, including work on sound, infrared, vacuum tubes, and radar. He has been intimately connected with the army program on radar since its inception, and was civilian engineer-in-charge of the first service tests on army radar. Dr. Zahl has been especially active in connection with the general Army program on vacuum tubes, being responsible for the development of a number of types used in U. S. Army equipment.

In 1942 Dr. Zahl entered active military duty as a major, and was promoted to the rank of lieutenant colonel in 1945. As an Army officer he continued to serve with the Signal Corps Engineering Laboratories, dividing his time between technical and administrative matters pertaining to the development of electronic equipment and vacuum tubes for use by the Armed Forces.

Upon separation from the Army in 1946, Dr. Zahl re-entered the Signal Corps Engineering laboratories as a civilian, and was shortly thereafter named chief of the engineering staff. He acted as official Signal Corps observer at the atomic bomb test at Bakini during the summer of 1946. On his return from Bikini he was given the assignment of building up a long-range postwar research program for the signal corps. In his new appointment Dr. Zahl will continue to emphasize research directed toward improving the defenses of this country.

He is the author of many scientific articles dealing with various research investigations conducted in the field of physics, and is holder of a large number of patents pertaining to vacuum tubes, radar, and communications. He is a member of the American Physical Society, the New York Academy of Sciencs, Sigma Xi, and Gamma Alpha, and is on the board of directors of the Armed Forces Communications Association.

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Newbern Smith (A'41–SM'46) has been advanced to the rank of Chief of the Central Radio Propagation Laboratory of the National Bureau of Standards, following the retirement of J. Howard Dellinger (see July Proceedings, page 889). Dr. Smith's new post will call for him to plan and direct basic theoretical and experimental radio wave propagation research, head the operation of the world-wide network of radio propagation observatories, and direct development of radio measurement standards at frequencies from 10 kc to 300,000 Mc.

Born in Philadelphia, Pa., in 1909, Dr. Smith attended the University of Pennsylvania in that city, receiving the bachelor's degree in electrical engineering, the master's degree, also in electrical engineering, in 1931, and the doctorate in physics in 1935. Awarded the Moore Fellowship in electrical engineering in 1930, the following year he taught that subject at the University and in 1933 he served as a research assistant. From 1934 to 1935 he taught physics at the Philadelphia College of Osteopathy.

After securing the Ph.D. degree, Dr. Smith joined the staff of the National Bureau of Standards as a physicist in the radio section of the electricity division. When the Central Radio Propagation Laboratory was organized in 1946, he was appointed assistant chief.

Dr. Smith has conducted considerable research in ionosphere measurements, oblique-incidence radio transmission, and radio critical frequencies in relation to solar eclipses and sunspot cycles. During the war he served as technical head of the Interservice Radio Propagation Laboratory set up by the U. S. Joint Chiefs of Staff.

Author of many technical papers, Dr. Smith has been a faculty member of George Washington University, and is a member of a number of professional fraternities.

W. E. Gordon (A'46) former associate director of the University of Texas' Electrical Engineering Research Laboratory, was recently appointed director of Cornell University's Microwave Astronomy Project.

Born in Paterson, N. J., on January 8, 1918, Mr. Gordon received the B.A. degree from Montclair State Teachers College in 1939. Seven years later he was awarded the M.S. degree from New York University.

During World War II, Mr. Gordon served with the Air Weather Services, and was associated with research on radar range forecasting and microwave propagation in the lower atmosphere at the Massachusetts Institute of Technology's Radiation Laboratory. In 1945 he joined the Electrical Engineering Research Laboratory at the University of Texas as a meteorologist, and became associate director in 1946.

Mr. Gordon is a member of Sigma Xi, the New York Academy of Science, and the American Meteorological Society.

Ricardo Muniz (M'43) has been appointed to the post of general manager of the Allen B. Du Mont Laboratories' television receiver division.

Previously Mr. Muniz had served as technical assistant to the vice-president of the company. Before joining Du Mont, he had extensive administrative and technical experience in electronic manufacturing with various firms.

C. C. Richelieu (A'42), secretary and general manager of Dairyland's Broadcasting Service, Inc., has resigned to become general sales manager of the Simplex Time Recorder Co. at Gardner, Mass. Prior to coming to Dairylands, Mr. Richelieu was Simplex's district sales manager for the State of Wisconsin. He has also been a radio engineer with the Civil Aeronautics Administration in Washington, and is currently director of the Central Division of the ARRL.

Harry Warren Wells (A'30–M'36–SM'43) of the Carnegie Institution of Washington was the recipient of the Washington Academy of Sciences' annual award for scientific achievement in the engineering sciences. The citation accompanying the award stated that it was given "in recognition of his distinguished upper air research and organization of a world-wide network of ionospheric stations."

Born in Washington, D. C., on January 13, 1907, Mr. Wells received the B.S. degree in electrical engineering in 1928 from the University of Maryland and the E.E. degree nine years later from the same institution. Between 1928 and 1932 he was associated with the Westinghouse Electric and Manufacturing Co., the All-American Malaysian Expedition to Borneo, Heintz and Kaufman, Ltd., and the Army Air Forces. Since 1932 he has been a member of the scientific staff of the Carnegie Institute's Department of Terrestrial Magnetism. His investigations, both here and abroad, have contributed materially to knowledge of the ionosphere, radio wave propagation, and related geophysical subjects.

Mr. Wells is a member of the Committee on Wave Propagation and Utilization of the IRE and Secretary-Treasurer of the Washington Section. He also belongs to the Washington Academy of Sciences, the American Geophysical Union, and the Philosophical Society of Washington.

Timothy E. Shea (SM'46), assistant engineer of manufacture of the Western Electric Co., has been elected president and a director of the Teletype Corp., a subsidiary of Western Electric.

A native of Newton, Mass., Mr. Shea received the bachelor's and the master's degrees from the Massachusetts Institute of Engineering and the B.S. from Harvard University just before he joined the Hawthorne Works, Chicago, in 1920 as a development engineer. A year later he transferred to the company's engineering department, which became the Bell Telephone Laboratories in 1924.

Early in his career he developed filters and networks used for transatlantic radio, carrier telephony, and television. In 1929 Mr. Shea was placed in charge of acoustical, optical, and electrical development in connection with sound motion pictures. He also supervised the development of public address, train despatching, and allied forms of equipment.

In 1939 Mr. Shea was elected vice president of Electrical Research Products, Inc., a subsidiary of Western Electric which became a division of the company in 1941. From 1941 to 1945 he was granted leave of absence to serve as director of war research at Columbia University, and as a member of Division Six, Undersea Warfare Committee, National Defense Research Committee. In this capacity he directed the New London, Conn., submarine research laboratory and supervised a large force of engineers and scientists working with the Pacific Fleet at Pearl Harbor.

Mr. Shea returned to Western Electric n 1945 as superintendent of manufacturing engineering at the company's electronic shops in New York City and became assistant engineer of manufacture at Broadway headquarters in 1945. The following year he was awarded the U. S. Government Medal of Merit, the highest civilian award and a presidential citation for exceptional services to the Navy submarine forces. Later that year Columbia University awarded him the honorary degree of Doctor of Science.

The author of "Transmission Networks and Wave Filters," Mr. Shea has also written a number of technical papers on various aspects of sound-motion-picture engineering. He is a former treasurer of the Society of Motion Picture Engineers and a former member of its Board of Governors, as well as being a Fellow of the Acoustical Society.

Eugene H. Fritschel (A'40—SM'46) was recently named sales manager in the industrial and transmitting tube division of the General Electric Co.'s tube division at Schenectady, N. Y.

A native of Waverly, Iowa, Mr. Fritschel was graduated from Iowa State College in 1926 with the B.S. degree in electrical engineering, and later that year joined General Electric as a student engineer. The following year he travelled to Uruguay as construction foreman to install radio transmitting equipment for the company. Upon his return to the United States, he became engaged in development work at Schenectady. Early in 1929 he was transferred to the Radio (now Electronics) Department, where he has taken charge of radio transmitter and industrial electronic tube sales. Prior to his new appointment he was sales manager for the tube division.

Donald G. Wilson (S'38–A'40–M'48) has been appointed chairman of the University of Kansas department of electrical engineering. He succeeds Dr. Victor P. Hessler, who is leaving to become professor of electrical engineering at the University of Illinois.

Dr. Wilson completed his undergraduate studies at the Rensselaer Polytechnic Institute of Troy, N. Y., and received the M.S. degree from Harvard University in 1939. After two years as manager of the fire-alarm department of an electrical company, he returned to Rensselaer to teach.

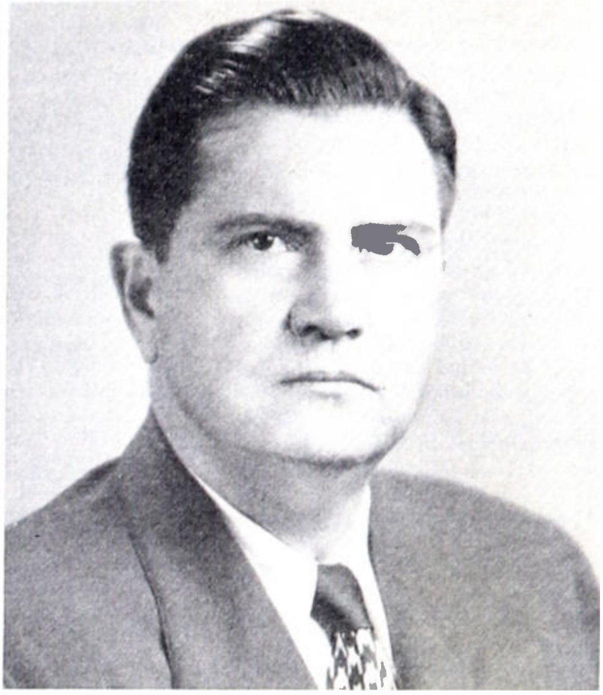
In 1942 he joined a group at the Massachusetts Institute of Technology's radiation laboratory which was working on microwave propagation research. Three years later he returned to Harvard to complete the requirements for the degrees of master of engineering science and doctor of philosophy in electrical engineering. In the fall of 1947 he joined the staff of the University of Kansas.

Albert C. Gable (SM'46), newly appointed division engineer of the General Electric Co.'s tube division, was formerly assistant engineer of the division.

After receiving the B.S.E.E. degree in 1929 from the Georgia School of Technology, Mr. Gable entered the employment of the General Electric Testing Department late in the same year. The following year he was transferred to the vacuum-tube engineering department, becoming a section leader on rectifier tubes in 1931, and later being assigned to an administrative post, which he held until his present promotion.

Alfred K. Wright (A'37–SM'43), chief radio engineer of the Tungsol Lamp Works, Inc., in Bloomfield, N. J., has been appointed a member of the Joint Electron Tube Engineering Council. Active in the council's standardization program since its beginning, Dr. Wright was formerly chairman of the receiving tube committee.

Dr. Wright's education includes the E.E. degree from Northeastern University in 1931, and the M.A. and D.Sc. from Harvard in 1932 and 1934, respectively.



Kenneth J. Gardner

Chairman, Rochester Section

Kenneth J. Gardner was born in Irondequoit, N. Y., on April 7, 1907. Virtually a "pioneer" radio worker, he entered the field in its earliest days, becoming an amateur radio operator in 1924. The call he received then as 8BGN he still holds today as W2BGN. At that time he helped to establish the old 80- and 40-meter bands, being one of the first to work regular 40-meter DX across the Atlantic and Pacific. In 1925 he entered the commercial side, working as an operator at WHAM in the evenings—which was the only time broadcasts were made in those days—and going to school during the day. The year 1926 found him both matriculating at the Rochester Institute of Technology for courses in electrical engineering and associating with General Electric engineers in establishing the first statewide radio broadcast network.

Early in that same year the Stromberg Carlson Co. took over the operation of WHAM, increasing its power to 5 kw and establishing connections with the old Blue Network. Mr. Gardner, who still continued to attend school, was placed in charge of all audio facilities and operation. It was at this time that he was required to learn the Morse code because, for the next few years until teletype service became established, all traffic was handled by telegraph at a pace of thirty-five words per minute.

Mr. Gardner finished his studies at the Rochester Institute of Technology in 1928 and later attended all the Ohio State broadcast conferences. Station WHAM grew to 50 kw and then into an FM and now into a Television station, and Mr. Gardner went along with it, first as assistant chief engineer and then as chief engineer of the company's broadcast facilities, which he designs and plans.

During the World War II, Mr. Gardner served with the Rochester War Research Corporation and was active in the RTPB. In 1945 he became a Member of the IRE, serving successively as Secretary-Treasurer, Vice-Chairman, and finally Chairman of the Rochester Section. He is also a director of the Rochester Engineering Society.

Charles L. Jeffers

Chairman, San Antonio Section

Charles L. Jeffers was born in San Antonio, Tex., on February 9, 1908. After having received the B.S. degree in electrical engineering from the University of Texas in 1929, he entered the employment of the Whites Uvalde Mines, near Uvalde, Tex., the following year.

Mr. Jeffers spent two years at the mines, then transferred to the sales department of the same company. In 1934 he left engineering in order to engage in other work, but in 1937 he returned to join the staff of radio station WOAI, and was appointed director of the engineering staff two years later.

In 1942, on leave of absence from the station, Mr. Jeffers became Assistant Radio Engineer, and, in 1944, Chief Radio Engineer of the Communication Facilities Bureau, Overseas Branch, Office of War Information, Washington, D. C. There he planned and secured equipment for a number of medium-wave and short-wave broadcast stations in the United States and overseas. He also assisted in the design and installation of rhombic antennas for several short-wave international broadcast transmitters with powers up to 200 kw. In 1945 he returned to his former position at WOAI.

Mr. Jeffers has been granted one patent and has two pending. On one of the latter, he has written a paper entitled "An Antenna for Controlling the Nonfading Range of Broadcasting Stations," which appeared in the November, 1948, issue of the PROCEEDINGS. The paper covers the design and describes model tests made on a broadcast antenna with very little high-angle radiation above an angle from the horizon which can be electrically adjusted from 40 to 60 degrees.

In 1938 Mr. Jeffers joined The Institute of Radio Engineers as an Associate and was upgraded to Senior Member in 1946. Two years later he was elected the first Chairman of the newly formed San Antonio Section. He is a registered professional engineer of the State of Texas, and a member of Tau Beta Pi and Eta Kappa Nu.

Problems in Connection with the Sale of Commercial Electronic Equipment*

WALTER A. KNOOP, JR.†, MEMBER, IRE

Since engineering includes the application of scientific principles, methods, and equipment to the meeting of human needs, it generally involves the production and distribution of apparatus and services. The response of the public to such a distribution is therefore of major interest to the engineer, since it indicates to him the degree of acceptance by the public of the products of his toil. The engineer is accordingly concerned to some extent with the broader aspects of distribution and sales of engineering products. Recognizing this fact, the Board of Directors of The Institute of Radio Engineers decided that, on a trial basis, there might appear in the PROCEEDINGS OF THE I.R.E., each year, a limited number of papers dealing with merchandising problems and methods. The number of such papers will be small and their subject matter will be selected, it is believed, so as to be of maximum interest and assistance to engineers. The following paper is the first of this type.

Since the publication of such papers in the PROCEEDINGS OF THE I.R.E. constitutes a novel departure, the members of the Institute are urged to write to the Editor expressing their reaction to the general policy of such publication and to any specific paper published in accordance with that policy. Such communications will be welcomed; an impression of the membership reaction can be obtained primarily through them.—*The Editor.*

INTRODUCTION

MOST OF US associated with the electronic industry recognize a growing trend toward earnest competition resulting from a more normal market condition. The preparation of this paper was prompted by two conditions observed in our industry of today by the writer.

1. There are many small new companies which will face extinction when their market seemingly begins to dry up. Little regard has been given to intelligent sales plans because, until recently, if something could be made it could be sold.

2. Some of the larger, better-known concerns have been drifting along at a happy pace. Management waxes complacent with good operating statistics on their desks and stockholders rest comfortably in the magic of "electronics." In spite of their size, scant attention has been placed on future sales plans.

Obviously it is impossible to present in this paper a complete blueprint of what should be done in each particular line of endeavor, but there follow a number of basic points that are considered significant.

PART I

The Product

Perhaps most companies are born because of a product. Someone has what is believed to be a "sure-fire" item because of several factors. These factors should be checked for factual accuracy.

1. Is there anything available now that does the same or a similar job?

2. If not, would any group other than friends consider its purchase for the work it is intended to accomplish? Is it possible to prepare a comparative cost analysis of performing the operation the old way as compared with the new?

3. If there is a similar item, analyze the product as to its competitive advantage.

4. Survey competition as to:
   (a) reputation and good will, and
   (b) merchandising ability.

These factors may either help or hinder you. Your product may be no better than your competitor's, but his poor selling effort may provide excellent opportunities for others.

The Customer

Something should also be known about prospective customers. Who they are will have a lot to do with what method ought to be used to sell them, since buying habits vary.[1]

Prepare a list of the various customer classes.[2] Then attempt to indicate that proportion of the sales volume which would be absorbed by each classification. This will show the more important customers and help to mold final sales plans to serve most effectively the largest market.

What Does the Customer Want

Future sales are lost at the very inception of a new-product program because the customer's desires are considered of only secondary importance. In too many firms there are evidences of "the doctor knows best" attitude. It is strongly recommended that the new product specifications, or a preproduction model of same, be submitted to some of the more important prospective customers.

What Sales Volume and When

More than one company has become bankrupt because it started off with "magnificent" ideas about what a short time it would take to sell a specific large volume of a product. Finished products went into stock, immediate orders did not materialize, and working capital was not sufficient to pay creditors. Then there is the other extreme; e.g., the firms that lost most of the punch in their original sales promotion, and consequently have lost important sales, because the product was not available for delivery.

A part of the product-analysis program must include a determination of production volume and scheduling to strike a good balance between the two extremes mentioned above.

As more experience is obtained in the volume of received orders, it will be possible to judge future volume with greater accuracy when combined with honest estimates from the field sales force.

The results of the product survey on dollar sales volume will have a very practical bearing on the size of the organization, the capital required, and the sales method. There will be cases where a good product might be conceived, but there would not be sufficient sales volume to support an organization to design, manufacture, and sell it.

Distribution and Sales Method

After completion of the above preliminaries, the producer should have an excellent collection of data to assist in deciding the most efficient form of selling and advertising.

In general, the following alternatives are available:

1. No sales force
2. Exclusive distributors
3. Manufacturer's representatives
4. Radio parts distributors and jobbers
5. Company-employed sales force.

There follows a brief description of the activities of each of these groups.

The term "no sales force" is meant to describe an operation which includes no field sales force of any kind. Dependence is placed on space advertising, direct-mail advertising, and customer recommendation to develop leads. Sales result generally from correspondence with the customer.

"Exclusive distributor" arrangements permit the manufacturer to eliminate the credit risks of dubious customers and provide a method of selling factory output in bulk quantities. It is usually reasonable to expect the exclusive distributor to commit himself for a minimum volume of business over some stated interval, such as a year or more, and to carry a stipulated minimum

* Decimal classification: R740. Original manuscript received by the Institute, September 7, 1948.
† Gawler-Knoop, Inc., Newark 2, N. J.

[1] "Electronics Asks the Buyer," McGraw-Hill Publishing Co., New York, N. Y.; 1944.
[2] "Pulse Beat of Industry," McGraw-Hill Publishing Co., New York, N. Y. October, 1945.

stock. The exclusive distributor may be expected to do a minimum amount of advertising and promotion in his territory, and should provide service and repair facilities. Using exclusive distributors, the manufacturer eliminates the difficulties of dealing with many customers, but at the same time he does not become intimately associated with the eventual purchaser of his product.

The question of "manufacturer's representatives" will be considered separately.

The radio parts distributor, or jobber, as he is often called, really performs the function of a dealer, insofar as industrial electronic equipment is concerned. In the case of test equipment, he is selling to the radio service man who is the user, or he sells to the industrials, schools, or local governments in his area. The operations of the "parts jobber" vary from a "storekeeper," who will sell what the customer asks for, to an aggressive operation in which certain products will be recommended on the basis of superior performance, by sales personnel who are technically well versed.

A directory of radio parts jobbers and their addresses is available from the National Electronics Distributors Association, 221 N. LaSalle St., Chicago 1, Ill., and from The Parts Jobber, 412 S. Michigan Ave., Chicago, Ill.

The Manufacturer's Representative

Somehow, the manufacturer must present his sales story to the jobber, if he sells to the jobber, and to the industrial and educational institutions, if he sells directly to them, or to both. For the most part, this function is performed by the manufacturer's representative, who is an individual or group of individuals active in rather well-defined territorial areas. The representative pays his own office, utility, travel, and entertainment expenses, and receives as his total compensation a commission on sales made in his territory. Most representatives handle several noncompetitive but complementary lines. Unfortunately, many representatives do not have a comprehensive technical background in electronics, and are only sufficiently versed in their particular lines to handle routine specification information. Conducting an intelligent presentation of a test or process equipment to engineering or management personnel requires a thorough knowledge of the product, its capabilities, and, as a matter of fact, its limitations. The wide-spread lack of technical knowledge on the part of representatives, jobbers, and dealers is one of the industry's present problems, and will not remain unsolved. It is heartening to note the increase in the number of engineering-trained men who recognize the opportunities that exist in selling electronic equipment.

The same problems exist as to the choice of representatives that appear in the choice of dealers. It will behoove the manufacturer to investigate thoroughly the type of operation, the nature, quantity, and quality of other lines handled, how the prospective customer values the services of the group under study, and which types of customers receive the most attention.

Listings of manufacturer's representatives, number of salesmen, territory covered, and products handled may be secured from "The Representatives," c/o William E. McFadden, 85 E. Gay St., No. 409, Columbus 15, Ohio.

Company-Employed Sales Force

A company-employed sales force provides the most complete control over a sales activity, and probably results in a more intimate knowledge of customers' problems and desires. It is a costly method, and for most small to medium-sized concerns, if sales costs are to be kept at a reasonable level, the number of men employed cannot possibly contact the number of accounts that should be frequently seen. Few successful salesmen are willing to work on a salary income only, and thus difficulties arise in establishing a fair means of remuneration.[3] As a firm grows in size and can afford fixed selling overhead charges, it is a management function to determine whether a company-employed sales force will be more effective and less costly than some other form of selling.

The usual sales method employs a combination of two or more of the groups described above, depending upon the particular abilities of the available groups, the class of product, financial considerations, and possible geographical buying preferences.

The above treatment objectively describes the preliminary market analysis and the available distribution channels for the manufacturer of electronic equipment. Part II will deal with the sales program and the effective utilization of these distribution channels.

Part II

Introducing the Product

The effective introduction of a new product will result in considerable savings in sales cost and effort as time passes. A regular time table should be planned so that the customer's natural enthusiasm for something new can be most efficiently converted to firm orders on the books.

Preliminary considerations involve the preparation of descriptive bulletins and photographs of a model, representative of the final product. The descriptive bulletin should at least contain complete performance specifications of the product. Nothing is so exasperating in analyzing equipment as to have only half the story told. Obviously, it is in order to accentuate the most desirable characteristics, but the omission of essential data is inexcusable and only provokes the intelligent prospect. If the product is a measuring instrument, provide accuracy data and quantitative information on frequency and other limitations. Show typical response curves, stability information, distortion data, transient-response characteristics, noise levels, amplifier and sweep linearity information, and, in the case of oscillographs, writing speeds. If this type of information is not known, then forget the bulletin and finish the engineering. Lack of pertinent data only serves to destroy confidence in the equipment. Too many manufacturers of measuring instruments completely omit reference in their bulletins to the accuracy with which the single measurement may be made for which the instrument is intended.

[3] "Trends in Paying Salesmen," McClure, Hadden & Ortman, 75 E. Wacker Drive, Chicago, Ill.

Space Advertising

At the same time that descriptive literature is under preparation, or even before, plans should be made for space advertising in trade journals. If the market analysis were done as described previously, the type of customer is known. Advertising representatives of the trade journals are always willing to present an analysis of their distribution to assist the advertiser in selecting his media.

If an extensive advertising program is contemplated, it would be worth while to retain the services of an advertising agency to plan the program, arrange for space contracts, art work, layout, and general publicity releases.

Publishers will be glad to furnish names and addresses of agencies placing ads in their publication. Often this information appears in the "Advertisers Index" in the rear of periodicals.

Far too many firms lose part of the value of their advertising dollar by not planning their program well in advance. Last-minute copy preparation and head-scratching for ideas is not conducive to effective advertising. There is a serious lack of presentation of interesting and useful data in technical advertising. Perhaps one of the reasons is the inability of the "general" advertising agencies to prepare good technical work. As in the case of the sales representatives, there is an excellent future for technically trained men with imagination in the advertising field.

Trade Exhibition

It is very advantageous to plan the introduction of the product at a trade show sponsored by the most logical customer class.

The questions asked about the product will aid in the preparation of future advertising and may indicate a few simple changes in design which will enhance its salability.

Plan space-advertising copy so that it relates to the convention and will create a desire on the part of the reader to visit the booth.

A dynamic display is much more effective than a static one. Simulate operating conditions, if possible. Be prepared to distribute descriptive bulletins, either at the show or by mail not more than a few weeks later.

Paper Presentation

An effective method of reaching those who might specify your product is the presentation of technical papers before local meetings of the appropriate engineering society. This type of promotion will go a long way toward building up a desirable reputation for ability. Obviously, the matter discussed must either be unique or contain interesting technical features, to be a worthwhile subject for such presentation.

Editorial Feature

Engineering and trade periodicals are always receptive to the publication of editorial material which contains information of interest to their readers. As in the case of the engineering-society paper, the matter presented must have reader interest.

New-Product Listings

Most engineering and trade publications contain a section cataloging new products. A usual practice is to condense the more pertinent performance specifications into a few short sentences, and briefly describe the product's purpose or features. This information, prepared in the forms of a "news release," is sent to the various interested publications with a photograph, stating the availability of suitable electros.

Direct Mail

If the prospective users of the equipment are well defined and not too numerous, the introduction can be done by mailing the descriptive bulletin with a covering sales letter to the individuals who should have use for it. This introduction letter may well be the first of a series of letters comprising a sales campaign.

The Sales Campaign

The process of introducing a new product may easily take a period of a year or more before it becomes generally known. Do not become discouraged by failure to receive a flood of orders immediately after the first publicity. Securing firm orders takes time.

Make plans to follow up inquiries received as a result of the introductory publicity. The follow-up can be accomplished by letter, by notifying your representative, or by furnishing leads to the jobber or dealer in the area. If you are using the services of a local representative, or a jobber who by his reputation can instill confidence in your product, give the customer his name and address.

Inquire of those who have already made their purchase as to their satisfaction. A tremendous wealth of information is available from this source: new uses and applications of the item, suggestions for improvement, why the customer selected your product, and any dissatisfaction he may be harboring and spreading to other prospects without your knowledge.

As new and interesting applications unfold, tell other customers about them. This type of information can be disseminated in the form of a direct-mail piece if it is only of limited interest, in your space advertising if of general interest, or by means of a house organ.

The House Organ

The house organ, properly prepared and edited, is a powerful sales tool. It can accomplish the following results:

1. Promote a wider use of your product by describing worth-while but generally unknown applications.

2. Establish a reputation for technical competence of your firm by presenting unusual and clever engineering "know-how" techniques embodied in your product.

3. Create a desire to deal with your firm because of its good personnel and customer relations.

There are individuals in our industry who consider their files of *The General Radio Experimenter* and *Du Mont Oscillographer* as prized possessions. Machlett Laboratories' *Cathode-Press* is an outstanding example of a combination external-internal house organ which contains not only technical features but also personal news of the plant employees and items of company policy. These types of publications are very effective in creating customer confidence in those who make the product.

First subscribers can be selected from your customer list, and then watch the subscription list grow by popular demand if the contents are of value.

The Catalog

The same comments apply to the catalog as to the descriptive bulletin. Many firms prepare their catalog from descriptive bulletins, adding pertinent commercial or "how-to-order" information.

The catalog should contain the following information:

1. An introduction to the company.
2. A complete description of the product.
3. Adequate price information and terms.
4. The guarantee policy.
5. Service information, where and how to arrange for same.
6. Complete company address and that of local representatives or dealers.

Instruction Book

The instruction book is of greater importance to selling than most concerns seem to realize. Not only does it assist in keeping a customer sold, but it is continually used as a source of equipment analysis by prospective customers.

At this point the author desires to decry the present insufficiency of so many instruction books that contain much "how it works" information, even to the extent of rather complex design considerations, but then singularly omit "how to use it" suggestions.

Selling Aids

Perhaps the most important and generally least considered phase of the selling operation is exemplified by "what not to do." It is the writer's observation that many firms in this industry go ahead and set up a workable and intelligent merchandising organization and then let it "die on the vine" by not keeping the field group informed about what is going on at the factory.

The entire sales organization must themselves be continually "sold" on the company and its products. Representatives should be regularly informed by means of at least a monthly newsletter as to company activities and developments. Periodic visits should be arranged between "reps" and the home-office people, both at the factory and in the field. Many companies have an annual or semiannual sales meeting, but that in itself is not enough to maintain enthusiasm throughout the entire year.

The same enthusiasm must be imparted to dealers or distributors. This is a function of the representative, assisted by the factory with such material as counter display, salesman's training booklet, lectures to dealers, and distributor's salesmen.

Another effective aid to enthusiasm is the contest between field sales groups, which may be part of a specific sales campaign or an annual proposition. Quotas can be established based on the sales potential for each geographical or trading area, and friendly rivalry among personnel thus encouraged.

The Service Policy

The manner in which returned or defective equipment is handled has a marked effect on the general reputation of a firm. One firm in the industry, making a commonplace item priced about the same as its competitors, is securing considerably more than its fair share of the total business because an enlightened service policy, among other things, provides for complete settlement in less time than their competitors require to reply to the customer's first letter of request for permission to return goods.

One company made the unfortunate error of transferring a man to the head of the service department, when he lacked tact in dealing with his fellow employees as a foreman.

A customer with defective equipment is a violent source of possible ill-will. Regardless of whether or not it is his fault, the matter should be disposed of with great speed, and judgment weighted in favor of the customer.

Commercial Policies

So often the accounting, billing, and credit departments conduct their business as if it were entirely unrelated to sales, and the sales department shies away from the unpleasantries of credit clearance, collections, and correspondence with customers affecting accounting problems. Until the sales group is satisfied that the accountants recognize and control difficulties that could cause poor customer relations, it is recommended that letters to customers be reviewed by responsible sales personnel as a matter of routine.

General Suggestions

It has been the author's observation that many new firms—in particular, those established by engineers or inventors who have seen an opportunity for their own enterprise—have neglected to appreciate the significance of the many small things which contribute to long-term success.

Noteworthy of these deficiencies seem to be:

1. Tardiness in response to correspondence.[4]
2. Unsightly preparation of descriptive bulletins and direct mail pieces. Poor composition and appearance of letters.
3. Lack of tact in appreciating the customer's problems and viewpoints.
4. Boasting rather than displaying restrained pride in descriptions of the product in the literature and instruction books.
5. Lack of effective co-ordination with representatives or dealers.

Conclusion

The author has attempted to show that the most successful sales program must be approached in an analytical fashion, administered with keen logic, and executed with close co-ordination by personnel technically competent with respect to the product and the industry in which it is used.

[4] "Very Promptly Yours," Hammermill Paper Company, Erie, Pa., 1943.

Instrumentation in the Field of Health Physics*

KARL Z. MORGAN†

Summary—**Instruments for personnel protection from radiation exposure are described. Many new developments have been made, and photographs and brief descriptions of instruments are given.**

Introduction

HEALTH PHYSICS is a new science organized for the purpose of studying radiation problems and of developing instruments and techniques, and is dedicated to the protection of persons from radiation exposure. The principal Health Physics organizations at the present time are located at Hanford Works, Oak Ridge National Laboratory, and Argonne National Laboratory. Altogether, about 350 persons are employed by these three organizations. Senior health physicists in these organizations are physicists who have specialized in radiation-protection problems. The large majority of the personnel are junior physicists, chemists, and chemical engineers who have completed only the B.S. degree. Laboratory technicians are used for routine operations in counter rooms and in the personnel-monitoring sections. The principal work of the Health Physics Divisions is divided into survey and monitoring functions, and into research and development.

An interesting phase of the Health Physics Program can be described by giving a short review of some of the thirty principal instruments that are used in making these radiation measurements. These instruments may be divided into three classes according to their use: (1) personnel monitoring, (2) area monitoring, and (3) survey instruments. The principal types of instruments include film badges, pocket capacitors, electroscopes, electronic circuits, Geiger-Muller counters, and proportional counters.

Personnel-Monitoring Instruments

Each person who enters areas where radiation exposure is possible is required to wear two pocket meters and one film badge. Two pocket meters are worn instead of one in order to provide a check on false readings due to electrical leakage. Two meters are always worn side by side, and since there is no mechanism that would charge the meters while they are being worn, all errors are in one direction. The lower of the pair is considered to be the true reading and is recorded as the radiation exposure. Three types of pocket meters have been used on the Atomic Energy Commission projects. The early type (see Fig. 1(a)) which was used at Oak Ridge National Laboratory gave one pair of false readings for about every 1,000 pairs of meters worn. Recent improvements (see Fig. 1(b)) have been made in this meter, and it is so nearly perfect that it is no longer necessary for a person to wear two meters if 0.16 per cent false readings can be tolerated. At present only one false pair of readings is obtained in 400,000 pairs of meters worn.

Fig. 1—Pocket meters worn in areas where radiation exposure is possible. (a) Type used originally at Oak Ridge National Laboratory. (b) Improved meter of type (a). (c) Rose pocket type.

Fig. 2—Racks for filing pocket meters and film badges at laboratory entrance.

* Decimal classification: 621.375.628. Original manuscript received by the Institute, April 21, 1948. Presented, 1948 IRE National Convention, New York, N. Y., March 23, 1948.
† Oak Ridge National Laboratory, Oak Ridge, Tenn.

These pocket meters have a useful range from 5 to 300 mr (milliroentgens). The Rose pocket meter (shown in Fig. 1(c)) was developed by J. E. Rose of Argonne National Laboratory. It has not been widely used, but it has many improvements in design. The most important improvement is the flexible plastic diaphragm in one end containing a charging button. This meter is completely sealed against dust and moisture.

Fig. 2 shows one of the portals through which persons pass when entering Oak Ridge National Laboratory. Each person picks up two meters from the far lower rack and his film badge beside his number in one of the other racks. In the evening, when leaving the Laboratory, the person places the film badge and both pocket meters in the slots adjacent to his number. After each shift change health-physics technicians collect all the pocket meters and read them on a Victoreen minometer similar to the one shown in Fig. 3. This meter is a projection-type

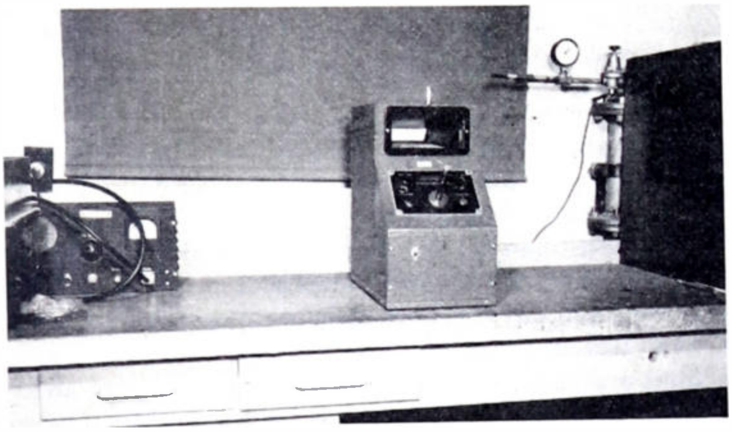


Fig. 3—Victoreen minometer.

fiber electrometer. The position of the fiber image on the scale indicates the discharge of this capacitor-type pocket meter while it was worn. Then it is recharged for use the next day by pressing the "charge" button, which applies 150 volts. Fig. 4 indicates that this voltage is suf-

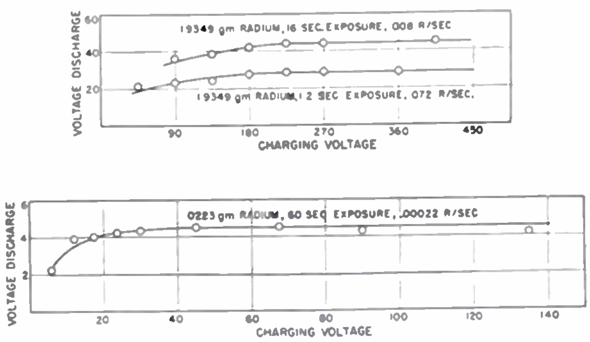


Fig. 4—Daily discharge plotted against charging voltage.

ficient to collect all the ions, and there is negligible loss due to recombination of ions unless a radiation flux of 250 roentgens per hour is exceeded. This is quite satisfactory, since personnel are usually not permitted to work in areas where the dosage rate exceeds 5 roentgens per hour.

Whenever 100 mr or more is recorded against a person in a single day, his film badge is collected, and the film is developed and read on a Weston photometer (see Fig. 5). A daily tolerance dose to X or gamma radiation is considered to be 100 mr. All film badges are processed on a routine basis once a week and the exposures recorded. Fig. 6 shows a close-up view of this film badge and the

Fig. 5—Weston photometer used for reading milliroentgens.

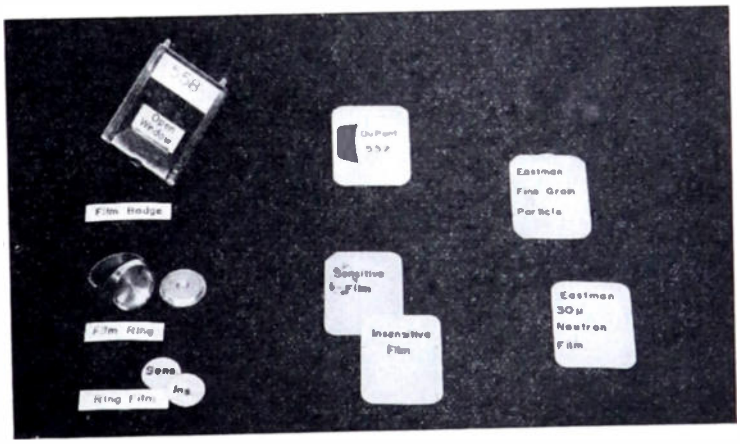


Fig. 6—Film badge and film packets; also film ring which is worn by personnel working with beta radiation.

two dental-size film packets it contains. The film badge has an open window and a middle section that shields both sides of the film by 1 mm of cadmium. The film packet worn in this badge contains a sensitive film that covers a useful range from 20 to 20,000 mr and an insensitive film covering a range from 1000 to 40,000 mr. Every time a batch of films is developed, a set of calibration films (which have received known exposures from a radium source) are developed with it in order to check errors due to the development process. The blackening of the center portion of the film which was shielded by cadmium is proportional to the gamma exposure, while blackening of that portion which was behind the open window gives some indication of beta exposure. If soft gamma radiation falls on the open window, readings are difficult to interpret because the films are about twenty times more sensitive to soft (<50 kv) than to hard (>200 kv) gamma radiation. When a person works with beta radiation, the film ring (shown in Fig. 6) is worn on a finger. It gives a much better indication of maximum body exposure, since the hands frequently come closer to such sources than the rest of the body.

Fig. 7 shows relative sensitivites of these films to beta and gamma radiation and indicates the useful range of

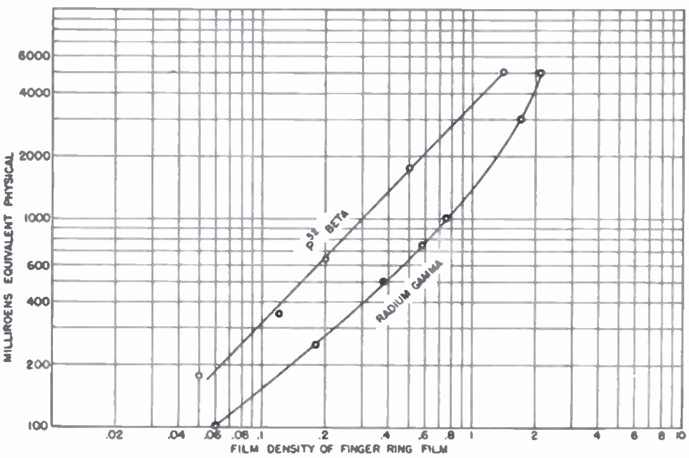


Fig. 7—Relative sensitivities of ring film to beta and gamma radiation.

this film. Persons who work where neutron exposure is possible wear also a fine-grain-particle film. A part of this film is behind cadmium shields and is exposed to fast neutrons. The reaction H^1 (n, p) then produces proton recoil tracks in the film. The film behind the open window is exposed to both fast and thermal neutrons. The thermal neutrons produce proton tracks by a capture reaction N^{14} (n, p) C^{14}. The films are developed and the proton tracks are counted with the aid of a dark-field microscope arrangement, as shown in Fig. 8. Daily tolerance is considered to be 1.3×10^8 neutrons per cm^2 per day, or 0.23 track per field of vision for thermal neutrons, and about 5.8×10^6 fast neutrons per cm^2 per day, or 0.27 track per field of vision of 2.0×10^{-4} cm^2. One of the principal difficulties in the use of the proton-track film is a nonlinear fading rate of the tracks, so that if the films have been worn a long time before developing it is impossible to make the proper fading correction unless the day of exposure is known. Table I shows some of the

TABLE I

FADING OF PROTON TRACKS IN EASTMAN FILMS

Time of Development (Minutes)	Type of Developer Used	Half-Life in Days for	
		NTA Emulsion Batch N4X17	NTA Emulsion Batch N4X18
4	D-19	12	14
5	D-19	16	17
6	D-19	23	20
3	Du Pont	15	15
4	Du Pont	23	20
5	Du Pont	38	23

data obtained by J. S. Cheka of Oak Ridge National Laboratory showing the effect of the type of developer and time of development on two types of Eastman films. When Eastman N4X17 film is developed with Du Pont developer for five minutes, films that have been delayed thirty-eight days between time of exposure and development have lost half of the recognizable tracks. Since the films are developed each week, fading loss causes an average error of about 10 per cent for N4X17 film if developed under optimum conditions.

When a person enters a high-level radiation area, pen-sized fiber electroscopes or dosemeters are worn. These meters (see Fig. 9) are charged with a battery. As an experiment progresses, the fiber drift can be read with the aid of a lens provided in one end of the meter. P. R. Bell of Oak Ridge National Laboratory has developed a very useful portable electronic pocket meter. It uses a VX-32 electrometer tube, and sounds an alarm after it has been exposed to a total dose of 50 mr. This meter, which weighs only 13.6 ounces complete with

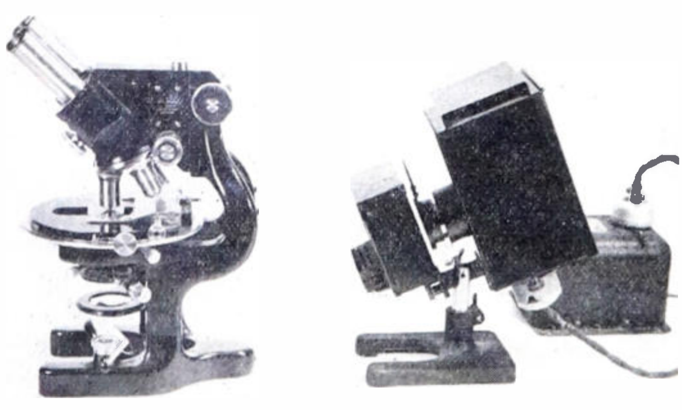


Fig. 8—Dark-field microscope for developing film and counting proton tracks.

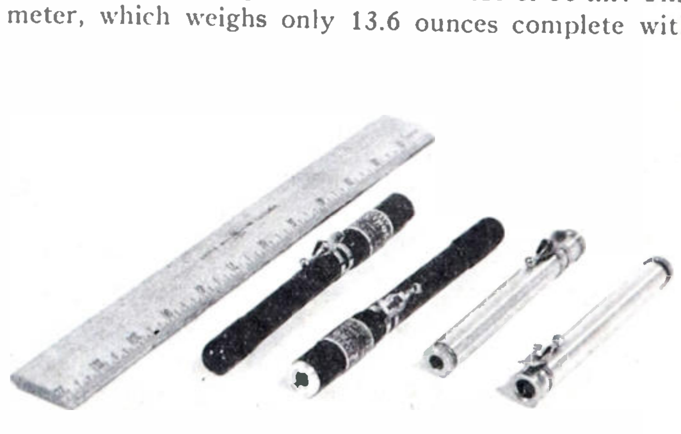


Fig. 9—Pen-sized fiber electroscopes or dosemeters for use in high-level radiation areas.

Fig. 10—Electronic pocket meter complete with batteries.

batteries, is shown in Fig. 10; and its circuit diagram is given in Fig. 11. The drain on the batteries is very low, and the life of the Mallory mercury cell and the three Eveready cells is essentially shelf life.

The foregoing describes personnel-monitoring meters measuring external radiation exposure to which a person is subjected. In addition, there is always the possibility of getting radioactive materials inside the body

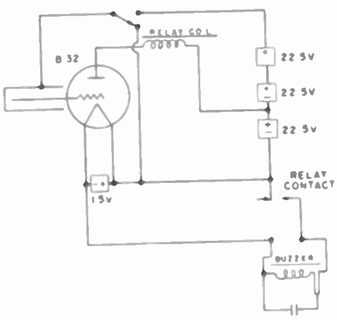


Fig. 11—Circuit diagram of the electronic pocket meter.

by inhalation, ingestion, or through the skin. In order to make certain that persons are not building up undesirable amounts of radioisotopes in the body, those persons who are considered to receive the maximum-body-intake exposures furnish urine samples to the laboratory for analysis on a routine schedule. L. B. Farabee, who developed the chemical method used at Oak Ridge National Laboratory for separating plutonium from urine and of preparing the sample for alpha counting, is shown in Fig. 12 placing one of these samples in an alpha proportional counter.

All protective clothing worn by laboratory personnel is washed in a decontamination laundry. Detergents are used to remove the dirt and a 3 per cent solution of citric acid serves to remove the radioactive products. After the clothes are washed and dried they are checked with special arrangements of Geiger-Muller (G-M) tubes to see that remaining beta and gamma activity is below a tolerance level. Alpha activity is monitored with a proportional counter called "Poppy."

Area Monitoring

Dust samples are collected continuously from the air inside the buildings and from the area in the neighborhood of the plants, to make certain that the air concen-

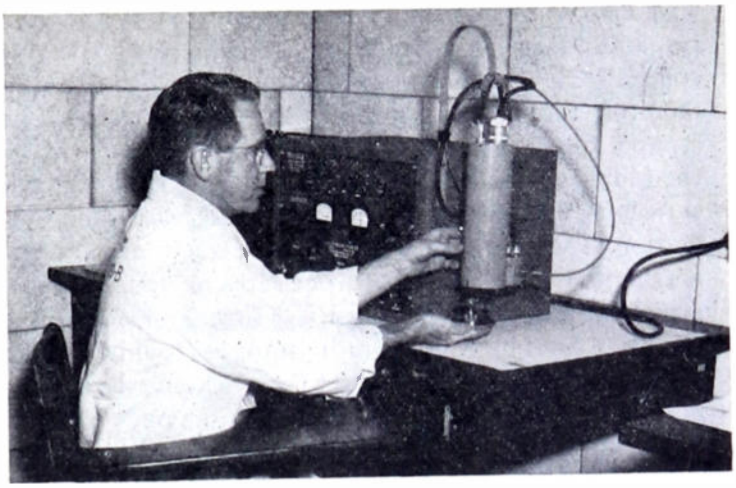


Fig. 12—Alpha proportional counter.

tration of radioisotopes at no time exceeds tolerance levels. Fig. 13 shows three instruments used at Oak

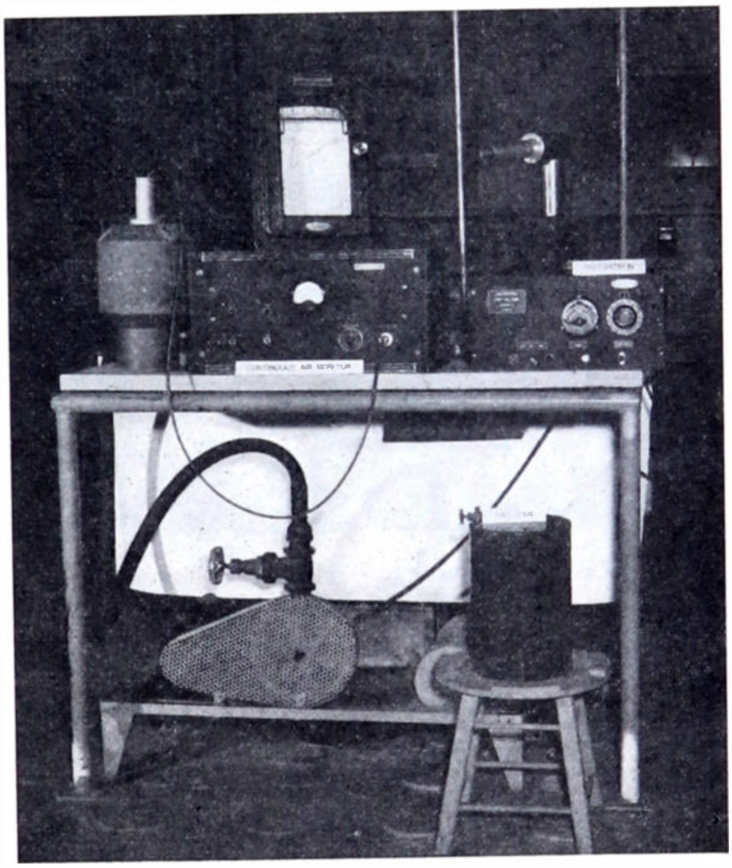


Fig. 13—Equipment for checking activity in the air.

Ridge National Laboratory for checking activity in the air. H. W. Speicher, who has been on loan to Oak Ridge National Laboratory from the Westinghouse Electric Corporation, assisted in the development of the precipitator model. The dust particles are electrically precipitated on a thin aluminum foil. The foil is removed and counted for alpha with a proportional counter and for beta and gamma activity with a G-M counter. If a high level of alpha activity is found, the foil is placed in a range analyzer which automatically plots a graph of energy distribution of the alpha rays emitted by the sample and makes it possible to ascertain the type and concentration of isotopes collected. A continuous air monitor (Fig. 13) consists of a G-M tube surrounded by a paper filter, through which air is drawn and on which the radioactive dust particles are deposited. This unit is surrounded with lead to reduce natural background radiation. The output of the G-M tube is fed into an integrating circuit which in turn operates an Esterline-Angus recorder. This furnishes a written record of the radiation level built up by radioisotopes accumulated on the filter paper, and simulates to some extent the accumulation of radioisotopes in the lungs of a person working in the same area. The evacuated can (Fig. 13) is used to collect samples of radioactive gas such as argon, and the activity of this gas is later determined in a chamber connected to an electrometer.

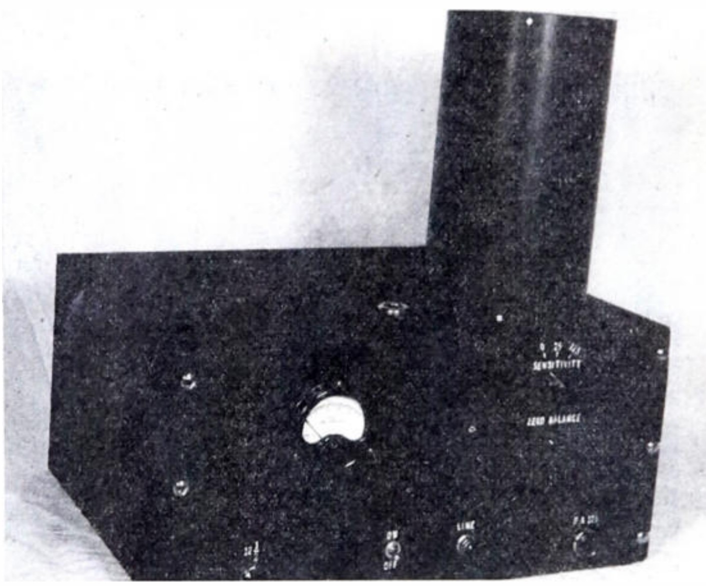


Fig. 14—Monitron, an area-monitoring instrument.

The monitron, developed at Oak Ridge National Laboratory by C. O. Ballou and P. R. Bell, is one of the most serviceable and versatile area-monitoring instruments. It uses an ac amplifier with a vibrating-capacitor ac signal which is modulated by the dc input from the ion-collecting chamber. Some of the ion chambers have an inner lining of boron which makes them sensitive to thermal neutrons. When the boron-lined chamber is surrounded by about six inches of paraffin with an outer shell of cadmium, it becomes an efficient epithermal-neutron monitor. Fig. 14 shows one of these monitors. (See also Figs. 15 and 16.) The instrument sounds an alarm if the radiation exceeds an eight-hour tolerance rate of 12.5 mr per hour.

One of the most satisfactory area-monitoring instruments in the early days of the plutonium projects was a large air ionization chamber or condenser called an X-22 chamber. Its greatest disadvantage was a need to read the instrument several times a day. This type has been replaced at Oak Ridge National Laboratory by G-M tubes and integrating circuits, located in various areas surrounding the Laboratory and maintaining a written record of the air activity on an Esterline-Angus re-

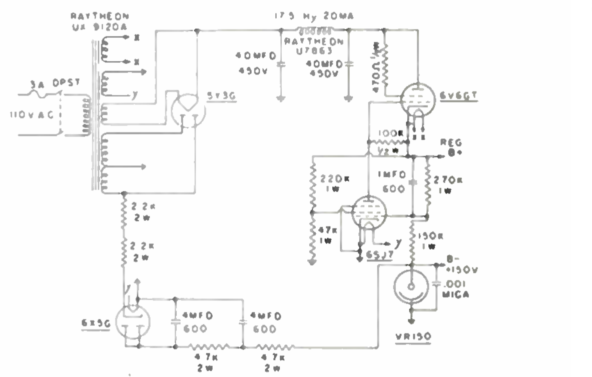


Fig. 16—Power supply for the vibrating-capacitor monitron.

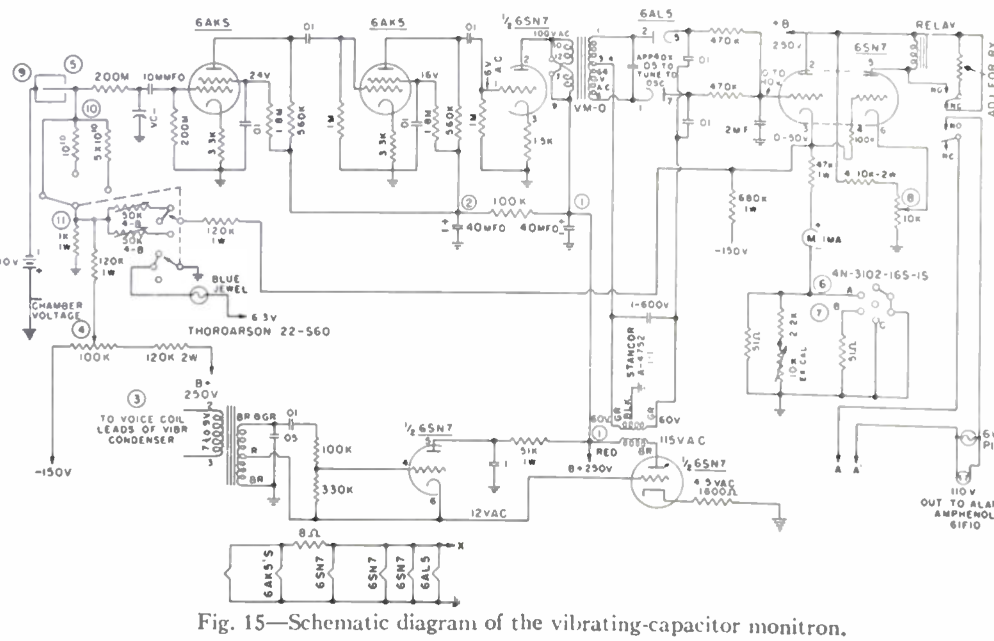


Fig. 15—Schematic diagram of the vibrating-capacitor monitron.

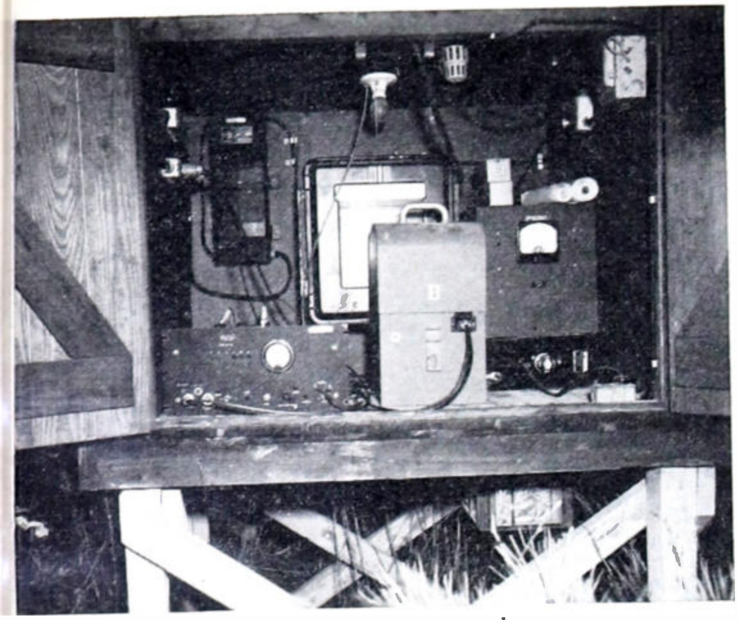


Fig. 17—Esterline-Angus recorder.

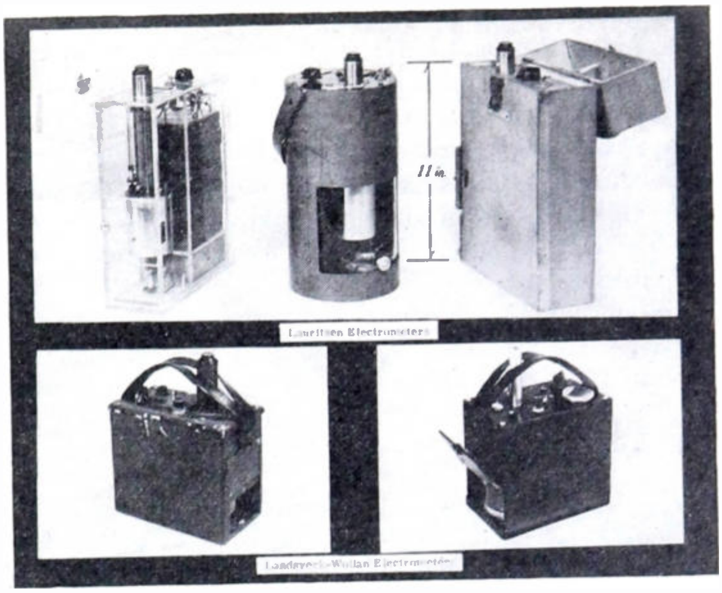


Fig. 18—Lauritsen and Landsverk-Wollan electroscope.

corder. Fig. 17 is a picture of one of these recording stations. In similar manner, G-M tubes submerged in the water of White Oak Lake keep a record of the water activity. In this manner, and by frequent calibration of these instruments, the air and water in the neighborhood of the Laboratory are known to remain within safe tolerance levels of radioactive contamination.

Survey Instruments

If it ever became necessary to choose the most reliable and indispensable of all health-physics instruments, the electroscope would without question be the first choice. Fig. 18 shows several different mountings of the Lauritsen and the Landsverk-Wollan electroscopes. These instruments are very rugged and can be used for years without changing their calibration or requiring any attention other than replacing batteries. Some of the electroscope ionization chambers are furnished with very thin (∼0.001 inch) cellulose-acetate walls to measure alpha and beta radiation. These thin chambers are covered by a bakelite or wood window (of thickness equal to or greater than the range of the gamma secondary electrons) when used for measuring gamma radiation. One of the most useful instruments for measuring thermal neutrons is made by lining the electroscope ion chamber wall with a thin layer of boron. If readings are to be made of a mixture of thermal neutrons and gamma rays, two electroscopes are then used—one lined with boron and the other with graphite. The boron-lined electroscope is converted into an epithermal-neutron meter by surrounding it with a few inches of paraffin, which in turn is surrounded by a sheet of cadmium. The Landsverk-Wollan electroscope is provided with a neon flash timer and is adjusted so that the position of the fiber on the scale when the neon bulb flashes indicates directly the radiation intensity in mr/hr.

Fig. 19 shows Zeus and Cutie Pie, two of the best electronic instruments available for measuring alpha, beta, and gamma radiation. Zeus is a two-tube balanced circuit, developed by F. R. Shonka of the Argonne National Laboratory. (See also Fig. 20.) He redesigned an earlier project instrument called the paint-pail meter

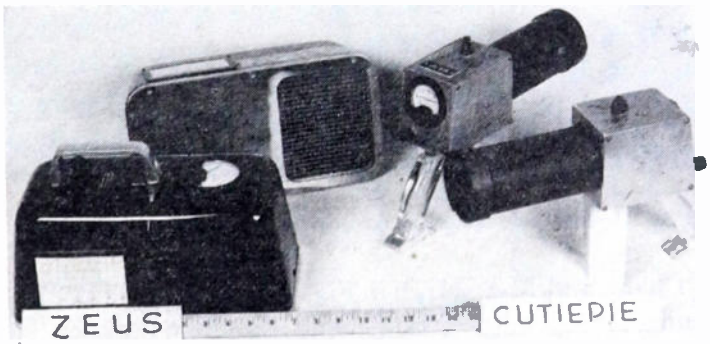


Fig. 19—Zeus and Cutie Pie, electronic instruments for measuring alpha, beta, and gamma radiation.

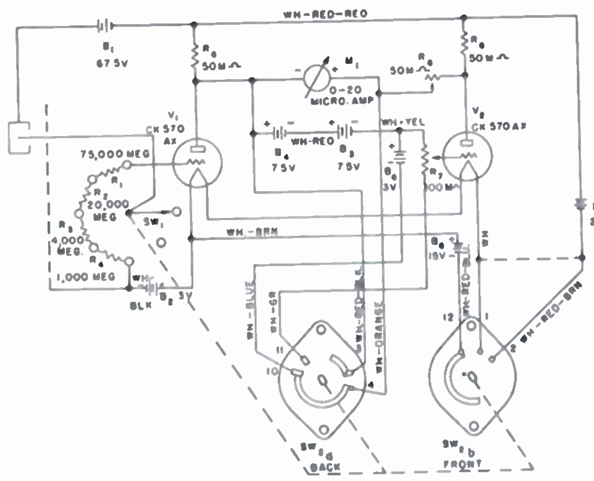


Fig. 20—Zeus meter schematic diagram.

which had been developed by W. P. Overbeck, then of Oak Ridge National Laboratory. A feedback tube was added to this balanced circuit to convert it into another useful meter, called Zeuto. When Zeuto is used for scanning an object like a table top it comes to equilibrium in a mere fraction of the time required by other circuits with high resistance in the input grid circuit. Zeuto resembles Zeus in external appearance. Cutie Pie, which resembles a large pistol, is especially useful for reading radiation behind lead shields in a hood. High levels of radiation are measured with the fish-pole meter, developed by C. O. Ballou at Oak Ridge National Laboratory. The chamber of this meter is fastened on the end of a long rod, as shown in Fig. 21.

Fig. 21—Fish-pole meter for measuring high levels of radiation.

Chang and Eng, shown in Fig. 22, constitute a double ionization-chamber arrangement developed by E. O. Wollan and P. W. Reinhardt, of Oak Ridge National Laboratory, for measuring fast neutrons. One chamber is filled with air or argon, the other with methane. Outputs of the two chambers are of opposite sign and are applied to a Ryerson electrometer. Gas pressures are adjusted for zero drift when placed in a gamma field, and then the drift brought about by fast-neutron hydrogen recoils in the methane chamber is a measure of the fast-neutron flux.

The proportional counter, Poppy, shown in Figs. 23 and 24, was developed by C. J. Borkowski and C. H. Marsh of Oak Ridge National Laboratory. It is one of the most useful alpha survey instruments available. It uses probes of many different sizes ranging from one small enough to be placed in a man's nostril to one constituting a square foot of useful area employed for laundry monitoring. These probes are converted into thermal-neutron detectors by lining them with boron, and into epithermal-neutron instruments by surrounding the boron-lined counters with paraffin and cadmium.

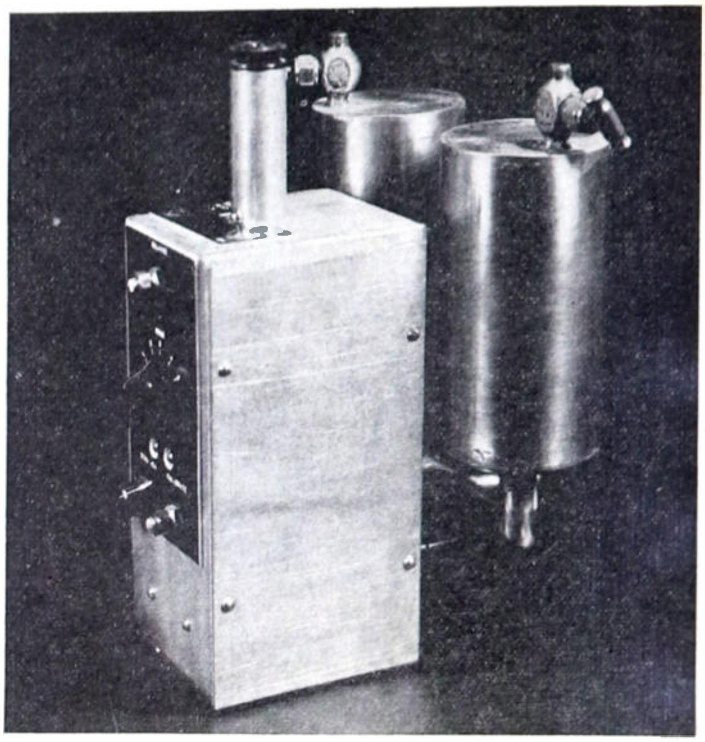


Fig. 22—Chang and Eng, equipment for measuring fast neutrons.

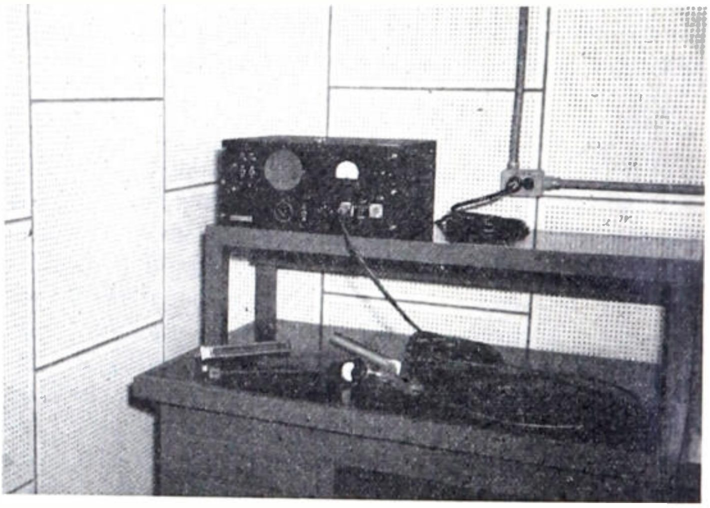


Fig. 23—Proportional counter, Poppy.

The advantages and disadvantages of the G-M tube have been discussed elsewhere,[1] and it should be sufficient to say here that, while it is one of the most sensitive and simplest instruments for beta and gamma measurement, it has many serious limitations. It should be used only as an indicating instrument, and never for quantitative measurements unless calibrated frequently and used by someone familiar with its defects. Fig. 25 shows four types of portable G-M tubes that have been used at Oak Ridge National Laboratory. The small, low-voltage

[1] K. Z. Morgan, *Chemical and Engineering News*, vol. 25, p. 3794; 1947.

Simpson unit, developed by J. A. Simpson of Argonne National Laboratory, proved to be the most reliable of such units at the Bikini tests in 1946. Fig. 26 shows the hand-and-foot counter developed by H. M. Parker of Hanford Works. This unit is one of the most useful applications of the G-M tube. It counts radiation from both sides of hands and feet. The weight of the two hands on microswitches in the slots starts the instrument, and it shuts off automatically after 24 seconds, indicating the radiation level on five separate recorders which are shown at the top of the panel.

Bragg-Gray Principle

Bragg and Gray have stated a principle of radiation measurement which, together with the definition of the roentgen, serves as a basis for the design of all health-physics instruments. According to this principle, the ionization per cubic centimeter of an irradiated medium is equal to the energy measured per cubic centimeter of air in a small air cavity in the medium, multiplied by the

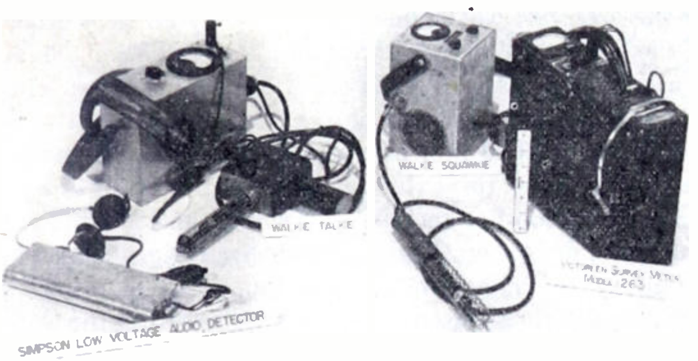


Fig. 25—Portable Geiger-Muller tubes used at Oak Ridge National Laboratory.

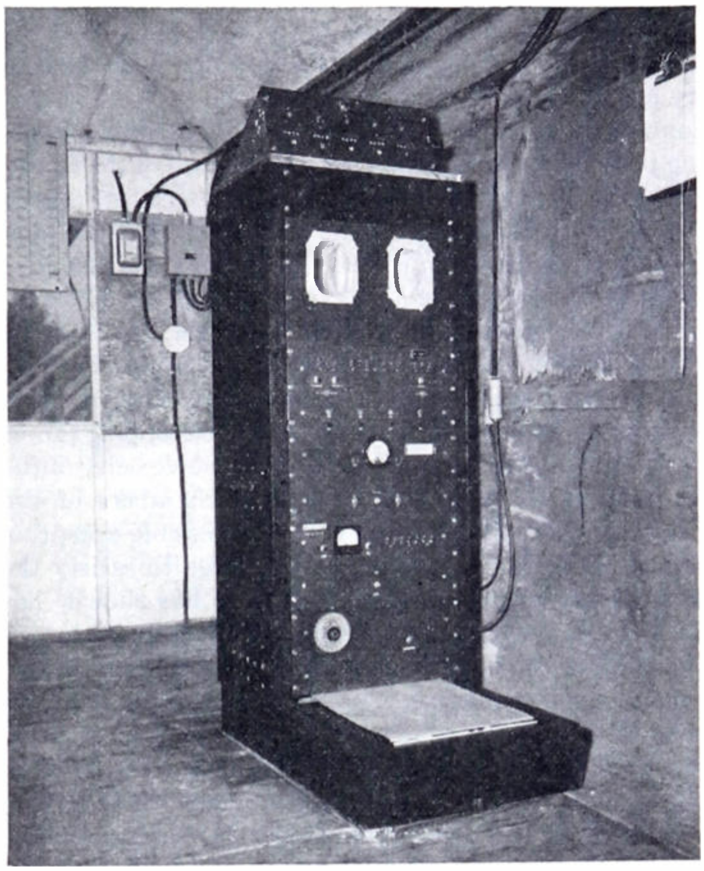


Fig. 26—Hand-and-foot counter.

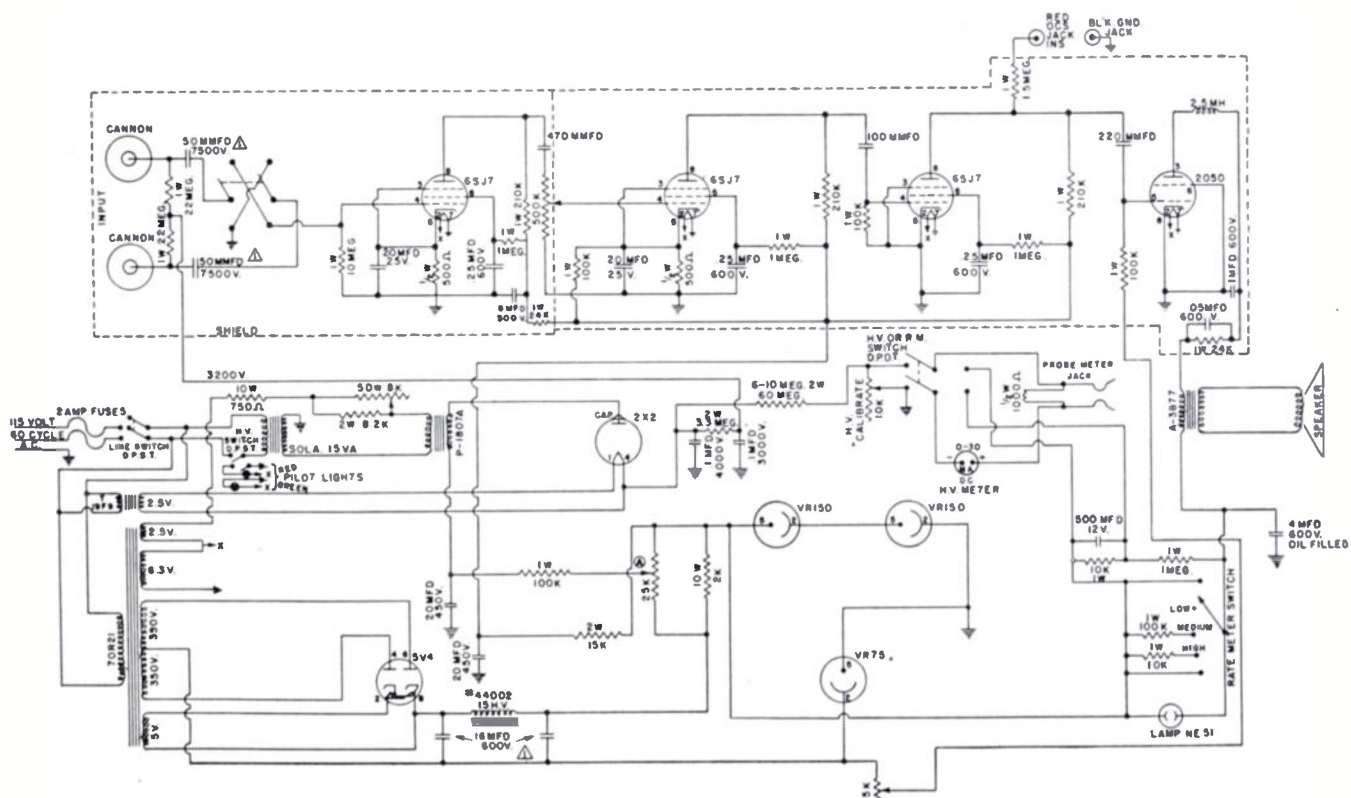


Fig. 24—Poppy amplifier circuit diagram.

relative stopping power of the medium and air for secondaries produced in the medium by the radiation. This principle[2–5] assumes (1) that the cavity is small compared to the range of secondary electrons, so that a negligible fraction of the energy is lost in the cavity; (2) that the intensity and composition of the radiation near the cavity are uniform; (3) that the gas in the cavity does not contribute any appreciable number of secondary ionizing particles—the gamma energy absorbed in the cavity should be negligible—and yet, if the material surrounding the cavity is of low atomic number so that the photoelectric contribution is very small, the Compton electrons generated in the air cavity will not introduce an error in the measurement of the roentgen; (4) that the cavity does not disturb the radiation flux by changing the direction and velocity of the electrons which cross its boundaries; and (5) that the relative stopping power of medium and air are independent of the ve ocity of the secondaries. Most of the ionization chambers of the health-physics instruments, with the possible exception of the pocket meter, are much too large to satisfy the Bragg-Gray principle. However, Gray has shown hat the relative stopping power is practically independent of the state of the medium (the range of grams per sq are centimeter is the same whether the medium is liquid, solid, or a gas). Thus, if the wal' material is air-equivalent (if it has an effective atomic number of air $\overline{Z}=7.7$), the wall in reality becomes continuous with the air in the cavity and the stopping power correction is unity. In any case, the inner air-equivalent wall material should have a thickness at least equal to the range of the secondary photoelectrons, and the gas in the chamber should be air, unless complicating energy-dependence, corrections are to be made. These corrections come about because the photoelectric-absorption coefficient varies as the fifth power of the atomic number of the medium and varies from the 3.5th to the first power of the wavelength as the energy increases from very low to high energy. Therefore, no simple correction can be made for the photoelectric effect unless the energy of the radiation is known (which is not the case for a health-physics instrument), or unless the walls and the gas are made of air. The solution is to use air for the gas and a chamber-wall material of an effective atomic number $\overline{Z}=7.7$, determined for the mixture in accordance with the equation of Fricke and Glasser[6]

$$\overline{Z} = \sqrt[3]{\frac{\sum_i a_i Z_i^4}{\sum_i a_i Z_i}}$$

Where a_i is the fraction by volume, and Z_i is the atomic number of each of the composite elements. The roentgen is defined as that amount of X or gamma radiation whose secondary electrons produce ion pairs in air amounting to 1 esu charge of either sign when the X or gamma radiation is absorbed in 1 cc of air. A good health-physics instrument should have sufficient voltage across its ion chamber to collect the ions produced before much recombination sets in.

Two health-physics instruments essential for calibration of other instruments are the free-air chamber and the extrapolation chamber. They are used to calibrate various types of sources as well as to check energy dependence and other instrument characteristics.

On the plutonium projects, it has been necessary to measure alpha, beta, and neutron radiations, as well as X and gamma. Since the definition of the roentgen applies only to X and gamma radiation, it was found convenient to use the term "roentgen equivalent physical," or the "rep." It was defined originally by H. M. Parker of the Hanford Works as that amount of radiation which is absorbed in tissue at the rate of 83 ergs per gram. The rep is sometimes defined as that amount of radiation which is absorbed in a microscopically small air cavity in the tissue to the extent of 83 ergs per gram of air. Each definition has certain advantages, and for its final name and definition the unit should await the decision of an international committee on units.

From the above discussion, it is apparent that the wall material of neutron chambers should be made of tissue-like material and that special care should be taken to see see that it contains the proper amount of hydrogen and nitrogen (because of the ${}_1H^1$ (n, p), ${}_1H^1$ (n, γ) ${}_1H^2$ and ${}_7N^{14}$ (n, p) ${}_6C^{14}$ reactions).

Conclusion

Not all of the various health physics instruments have been discussed in this paper, but a few of the more important problems have been indicated, with particular emphasis on the part which certain instruments have played in their solution. Few, if any, new principles were involved in the design of these instruments, but the application is new in many cases and the required number of radiation-detection instruments has increased many fold. Most of these instruments are very poor from the standpoint of engineering and design, and the demand for these instruments is far greater than the supply. An effort is being made to interest a larger number of engineering concerns in the development, engineering, design, and production of these instruments. Nevertheless, in spite of the imperfections of some of the instruments, the health physicists have used them successfully. This is attested by the fact that thousands of persons on the plutonium projects at Argonne National Laboratory, Oak Ridge National Laboratory, and Hanford Works have relied upon these instruments while working with millions of curies of radioactive material, and not one has received any radiation damage.

[2] W. Bragg, "Studies in Radioactivity," Macmillan Publishing Co., New York, N. Y., 1912.
[3] L. H. Gray, *Proc. Roy. Soc. A*, vol. 122, p. 647; 1929; and vol. 156, p. 578; 1936.
[4] L. H. Gray, *Brit. Jour. Radiology*, vol. 10, pp. 600, 721; 1937.
[5] L. H. Gray, *Proc. Camb. Phil. Soc.*, vol. 40, p. 72; 1944.
[6] H. Fricke and O. Glasser, *Amer. Jour. of Roentgenology and Radium Therapy*, vol. 13, p. 453, 1925.

High-Frequency-Heating Characteristics of Vegetable Tissues Determined from Electrical-Conductivity Measurements*

T. M. SHAW† AND J. A. GALVIN†

Summary—An investigation was made to determine the high-frequency heating characteristics of certain vegetable tissues by means of electrical-conductivity measurements. The specific conductivities of potato, carrot, apple, and peach at 25°C were found to be of the same order of magnitude and range from about 3×10^{-4} ohm^{-1} cm^{-1} at 10^3 cps to 5×10^{-3} ohm^{-1} cm^{-1} at 4×10^7 cps. Variation in the temperature from -80 to $+30$°C is accompanied by a gradual rise in conductivity with a sharp tenfold increase at the melting point of ice. For temperatures between the melting point and 30°C, the conductivity increases approximately 2 per cent per °C. Investigation of the effect of the water content on the specific conductivity showed, in the case of carrot, that a reduction in the water content from a value of 80 per cent, characteristic of the raw tissue, to a value of 1.5 per cent is accompanied by a ten-thousandfold (at 3×10^7 cps) decrease in specific conductivity. These results are discussed briefly in relation to industrial applications of the high-frequency-heating method in the food and pharmaceutical industries.

INTRODUCTION

THE METHOD of heating poor thermal conductors by means of high-frequency currents has proved to be extremely valuable in industry. Despite the fact, however, that biological materials constitute one of the largest classes of poor heat conductors, the general usefulness of the high-frequency method in the industrial processing of such materials remains largely unproved.

In attempting to evaluate the applicability of the high-frequency method of heating to certain biological materials important in the food and pharmaceutical industries, it was found that few data are recorded pertaining to the high-frequency electrical characteristics of such materials. Although studies of the electrical impedance have been made on a number of important biological systems, these are confined largely to single cells, cell suspensions, and to certain tissues of animal origin. In most cases the frequency range does not include those frequencies of the order of 10^6 to 10^8 cps which are being used in the majority of high-frequency-heating investigations. Other than the very early work of Phillippson[1] on potato, there appear to be no data pertaining to the high-frequency characteristics of biological materials of plant origin.

The present investigation was undertaken to determine how the high-frequency-heating characteristics of biological materials of vegetable origin are related to their temperature and water content, and to the frequency of the heating current. The method chosen was the determination of the ac conductivity; this quantity, as is well known, is a measure of the rate at which energy is dissipated in the material through which current is flowing. The materials studied were potato, carrot, apple, and peach.

For the purposes of this paper the electrical impedance of a sample of a biological material contained between two parallel-plate electrodes of area A (square cm) separated by a distance l (cm) will be represented by a two-terminal network consisting of a capacitance C (farads) and a resistance R (ohms) in parallel. Then the specific conductivity σ and dielectric constant ϵ of the biological material are given by the equations

$$\sigma = l/AR \quad \text{and} \quad \epsilon = 36\pi 10^{11} Cl/A.$$

Both σ and ϵ depend on the frequency, the temperature, and possibly other variables.

If a network of this type is connected to a source of alternating current, the rate at which energy is dissipated in a 1-cm cube of the biological material is given by the equation

$$P = E^2\sigma$$

where P is the power in watts, E is the rms value of the applied alternating emf, and σ is the specific conductivity, as defined above. Since we are especially interested in the high-frequency-heating characteristics of biological materials, attention will be confined primarily to the behavior of σ. The dielectric constant ϵ, which together with σ is sufficient to characterize the electrical properties of a material, is important in the present discussion only insofar as it influences the design of electrodes and the capacitance of the load to which the high-frequency source must be coupled.

EXPERIMENTAL

The specific conductivities of the several vegetable materials were determined by conventional methods. For frequencies between 10^3 and 5×10^6 cps, a bridge-substitution method was used.[2] For frequencies between 10^7 and 4×10^7 cps, a commercial Q meter was used. The procedures described by Bady[3] concerning in-

* Decimal classification: R598×R280. Original manuscript received by the Institute, March 25, 1948; revised manuscript received, June 3, 1948.

† Western Regional Research Laboratory, U. S. Department of Agriculture, Albany, Calif.

[1] M. Phillippson, "Les lois de la résistance électrique des tissus vivants," *Bull. Acad. Roy. Belgique, Cl. Sc.* 7, pp. 387–403; June, 1921.

[2] T. M. Shaw, "The elimination of errors due to electrode polarization in measurements of the dielectric constants of electrolytes," *Jour. Chem. Phys.*, vol. 10, pp. 609–617; October, 1942.

[3] I. Bady, "Notes on Increasing the Accuracy and Usefulness of the Boonton Q Meter, Model 160 A," U. S. Department of Commerce, Office of the Pub. Board, Report No. PB2506.

ductance corrections at the higher frequencies were followed.

The conductivity cell was especially designed to be used with the Q meter at frequencies of the order of 10^7 cps. The need for short, low-inductance leads and small stray capacitance was met by making the cell small and of simple design. As shown in Fig. 1, the cell

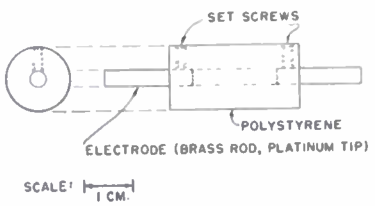


Fig. 1—Sketch of conductivity cell.

accepts a small cylindrical specimen of about 0.3-cm diameter by 1.5-cm length, cut from the material under investigation.

The calculated geometrical cell constant was checked by measuring the conductivity of sodium chloride solutions. The cell constant found in this way agreed within 5 per cent with the geometrical value for the entire frequency range.

Most measurements were made at 25°±1°C in a constant-temperature room. In this way it was possible to avoid the complexities associated with thermostating the conductivity cell in a high-frequency circuit. Similarly, in the few measurements made at temperatures between −80 and +25°C, the need for a thermostated cell was avoided by bringing the conductivity cell to the desired temperature in a constant-temperature bath, and removing it for the few seconds required to establish a preliminary balance of the Q meter. The cell was then returned to the bath and, after temperature equilibrium was again established, the cell was removed again for the final balance.

The principal limitation to the accuracy of the specific conductivities is probably that due to variations between successive samples. The *maximum* variation among a number of samples, varying in length by 100 per cent, was found to be ±20 per cent or less. This result was found by measuring the conductivities of eight samples from two potatoes at 10^3 and 10^5 cps and twelve samples from six apples at 3×10^7 cps. This 20 per cent variation includes, in addition to the variations in the samples, errors in the bridge and Q meter.

Results and Discussion

Fig. 2 shows the relation between σ and frequency for potato. These data are representative of the materials studied in the present investigation. They show the typical sigmoid characteristic first reported for living tissue by Phillippson[1] and which has been found for many biological systems.[4,5] Phillippson's data on potato cover the frequency range from 500 to 3×10^6 cps. The present measurements cover the frequency range from 10^3 to 4×10^7 cps, and are in good agreement with the measurements of Phillippson.

The origin of the sigmoid characteristic, obtained for the relation between the specific conductivity and the frequency for biological tissues, has been discussed by Phillippson[1] and by others. It arises from the fact that the cell membrane is an electrical insulator. Thus, at low frequencies, current flow is confined principally to the intercellular fluid. At higher frequencies, some current flows through the cell membranes into the cell interior. Finally, at sufficiently high frequencies, current penetration through the exterior membrane is complete, and the conductivity levels off in the manner shown in Fig. 2.

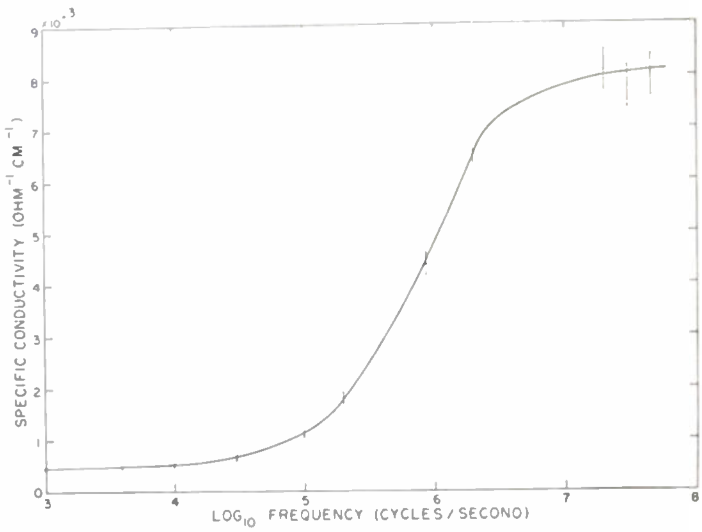


Fig. 2—Frequency variation of specific conductivity of potato at 25°C. The vertical lines through the plotted points indicate an estimated 5 per cent error.

Measured values of the conductivity for all the materials studied in the present investigation are summarized in Table I. From these data it is evident that the specific conductivities of these diversified materials are all of the same order of magnitude. Furthermore, they agree in magnitude with the conductivities of a number

TABLE I

Specific Conductivities of Vegetable Materials at 25°C.
Specific Conductivity, σ(ohm^{-1} cm$^{-1}\times10^{-3}$)

Frequency cps.	Potato (white)	Potato (sweet)	Carrot	Apple	Peach
10^3	0.46	0.64	0.26	0.25	0.52
4×10^3	0.48	0.67	0.28	0.28	0.52
10^4	0.51	0.72	0.30	0.34	0.59
3×10^4	0.65	0.88	0.34	0.56	0.77
10^5	1.1	1.4	0.52	1.0	1.3
2×10^5	1.8	2.0	0.80	1.3	1.9
8×10^5	4.4	3.8	2.7	1.5	—
2×10^6	6.5	5.3	5.2	—	—
2×10^7	8.1	6.8	8.0	1.6	4.4
3×10^7	7.8	6.6	7.9	1.6	4.3
4.4×10^7	8.0	6.7	7.8	1.6	4.5

[4] K. S. Cole, "Electric impedance of suspensions of spheres," *Jour. Gen. Physiol.*, vol. 12, pp. 29–36; September, 1928.

[5] K. S. Cole, "Electric impedance of suspensions of arbacia eggs," *Jour. Gen. Physiol.*, vol. 12, pp. 37–54; September, 1928.

of animal tissues[6] which range from 0.3 to 8×10^{-3} ohm^{-1} cm^{-1} for a frequency of 10^8 cps at 23°C.

The water content of biological materials is an exceedingly variable quantity. Although many vegetable and animal tissues contain from 60 to 90 per cent water, there are specialized tissues which contain less water than others. Because of this variation in the water content, it is of interest to examine the effect of water content on the high-frequency-heating properties. A further interest arises from the possible use of high-frequency heating in the dehydration of biological materials.

Probably the most comprehensive study of the electrical properties of a biological material in various stages of hydration is that of Dunlap and Makower.[7] These authors studied the electrical impedance of a vegetable material (dehydrated carrot) containing from 1.5 to 20 per cent water at frequencies ranging from 1.8×10^4 to 5×10^6 cps. A part of their data, together with our measurements on raw carrot, is presented in Fig. 3, which shows the variation in σ with frequency at

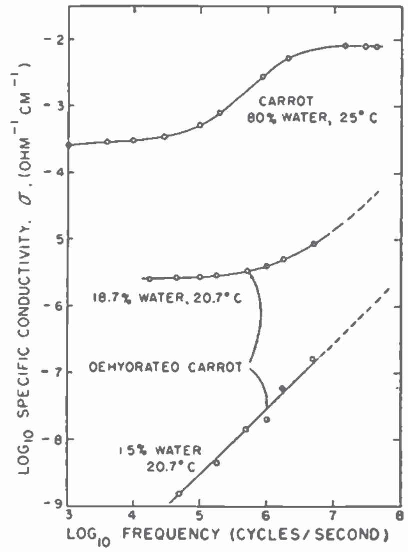


Fig. 3—Frequency variation of specific conductivity of carrot and dehydrated carrot. Data for dehydrated carrot taken from Dunlap and Makower.[7] Conductivities of carrot are for 25°C.; for dehydrated carrot at 20.7°C.

various moisture levels. Since the data on dehydrated carrot extend only to 5×10^6 cps, a short extrapolation was made to higher frequencies of special interest in high-frequency heating.

In comparing the data for raw and dehydrated carrot, it should be noted that the differences shown are not due solely to the effects of moisture. The data on the dehydrated carrot were obtained on a compacted mass of dehydrated carrot particles and, as Dunlap and Makower showed, the values of σ are influenced by the degree of compaction. However, for the present purpose these effects are of secondary importance to the effects of moisture.

It is apparent from the curves in Fig. 3 that an enormous decrease in σ, about 10^4-fold at 3×10^7 cps, accompanies the reduction in moisture content from the 80 per cent level, characteristic of the natural carrot, to that of the dehydrated material. It is also evident that a high-frequency generator designed for dehydration applications must accommodate materials with very wide ranges in conductivity. In tests of the use of high-frequency heating for dehydration an attempt has been made to dry only partially dehydrated materials,[8-10] and thus the necessity of dealing with large changes in σ was partially avoided.

Probably of greater interest than the change in the magnitude of σ, shown in Fig. 3, is the marked change in the form of the frequency dependence of σ which accompanies the reduction in water content. As remarked earlier, natural biological materials containing a large quantity of water show only a relatively small dependence on frequency, especially in the megacycle range. However, when the moisture content is reduced to a very low percentage, the frequency dependence of σ becomes large. In this case σ is approximately proportional to the frequency, as is also the case for many solid dielectrics.[11] Data for other low-moisture biological materials have been published by Sherman,[12] who found that the loss factor ϵ'' of a number of food materials containing a small percentage of water is practically independent of frequency over the range of 5×10^6 to 3×10^7 cps. Since ϵ'' and σ are related according to the equation

$$\sigma = [\epsilon'' f/(1.8 \times 10^{-12})] + \sigma_0$$

(where f = frequency and σ_0 = dc conductivity) it follows that σ is proportional to the frequency. (For nearly all dry materials at the frequencies under discussion here, $\sigma_0 \ll \sigma$).[7]

The conductivity of biological materials shows an interesting dependence on temperature. Since allowance must be made for changes in σ during the course of the high-frequency-heating process in the design of the heating equipment, a knowledge of the temperature dependence is important. Data on temperature dependence are of value in efforts to apply high frequency to the thawing of frozen foods.[13,14]

[6] B. Rajewsky, H. Osken, and H. Schaefer, "Hochfrequenzleitfähigkeit biologischer Gewebe im Wellenängenbereich von 3 bis 1400 Meter," *Naturwissenschaften*, vol. 25, pp. 24–25; January 8, 1937.

[7] W. C. Dunlap, Jr., and B. Makower, "Radio-frequency dielectric properties of dehydrated carrots: application to moisture determination by electrical methods," *Jour. Phys. Chem.*, vol. 49, pp. 601–622; November, 1945.

[8] V. W. Sherman, "Electronic dehydration of foods," *Electronics*, vol. 17, pp. 94–97; February, 1944.

[9] E. Rushton, E. C. Stanley, and A. W. Scott, "Compressed dehydrated vegetable blocks," *Chem. and Ind.*, pp. 274–276; September, 1945.

[10] W. C. Dunlap, Jr., "Vacuum drying of compressed vegetable blocks," *Ind. Eng. Chem., Ind. Ed.*, vol. 38, pp. 1250–1253; December, 1946.

[11] S. O. Morgan and W. A. Yager, "Dielectric properties of organic compounds," *Ind. Eng. Chem., Ind. Ed.*, vol. 32, pp. 1519–1528; November, 1940.

[12] V. W. Sherman, "Electronic heat in the food industries," *Food Ind.*, vol. 18, pp. 506–509, 628, 630; April, 1946.

[13] T. L. Swenson, "Research on agricultural products," *Sci. Monthly*, vol. 62, pp. 525–537; June, 1946.

[14] W. H. Cathcart, "Frozen foods defrosted by electronic heat," *Food Ind.*, vol. 18, pp. 1524–1525; October, 1946.

The variation in σ with temperature of vegetable tissue in the natural state is complex. Fig. 4 shows data obtained with potato for the temperature range -80 to $+25°C$. For temperatures above freezing, the change in σ is small, approximately 2 per cent per °C. Below the freezing point, however, σ changes rapidly until most of the water present in the tissues has been frozen. A similar change in the dielectric constant also occurs and has been demonstrated for plant materials[15] and for flour dough.[16] The very pronounced change in σ near the melting point of ice is of special importance in the use of high-frequency heating to thaw frozen food. This characteristic is responsible for the nonuniform thawing observed in experiments of this nature.

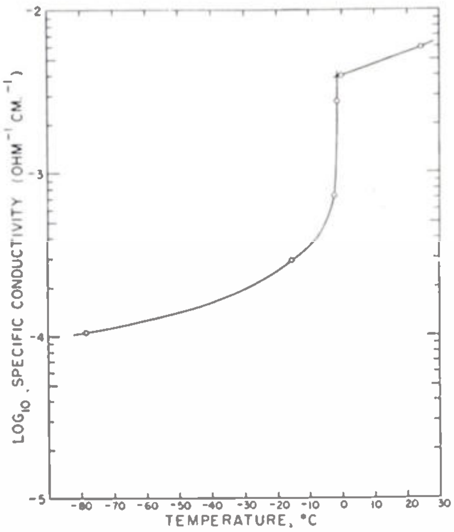


Fig. 4—Variation of specific conductivity of potato with temperature for a frequency of 30 Mc.

If it is assumed that a uniform electric field is applied to the frozen material, all portions of the material must be at the same temperature to assure uniform thawing. If the temperature is not uniform, regions at higher temperatures will heat more rapidly because their electrical conductivity is greater. The final result is that some portions become completely thawed and even overheated, while near-by portions remain frozen. Since it is very difficult, if not impossible, to obtain a uniform temperature throughout a material, it is questionable whether satisfactory results can be obtained by this method, at least at frequencies and for materials for which the conductivity varies in the manner shown in Fig. 4.

The question of proper frequency to employ in high-frequency heating has been raised frequently. According to the well-known Debye theory of polar dielectrics,[17] a maximum energy loss per cycle occurs at a critical frequency corresponding to the relaxation time of the dipoles. Such maxima can be demonstrated experimentally. The amount of heat developed in one second in a dielectric, however, depends on the product of frequency and loss per cycle. It is easily shown that in an ideal Debye case an increase in frequency beyond the critical frequency is sufficient to offset the drop in the loss per cycle, so that the power input to a dielectric never falls off with increasing frequency. For most materials the variation in loss per cycle with frequency is usually much less than is predictable by the Debye theory. In the limiting case of constant loss factor, such as is found experimentally to be approximately true for polystyrene[18] at frequencies as high as 10^{10} cps, a tenfold increase in frequency is accompanied by a tenfold increase in the specific conductivity.

For materials with approximately constant loss factor, the equation $P = E^2\sigma$, given above, indicates that a given power input to the material can be obtained at low voltages by working at higher frequencies where σ is large. This is an important consideration, because with many materials rapid heating at low frequencies requires the use of voltages approaching the breakdown voltage. It has been suggested that the use of frequencies in the microwave range[19] may eliminate this difficulty.

The above situation does not hold for biological materials with large specific conductivities, such as those shown in Table I. As previously stated, σ is practically constant for frequencies in the range 10^6 to 10^8 cps for materials of this nature. Although some increase in σ is to be expected at still higher frequencies, recently published measurements on *cooked* vegetables[20] indicate that such increase in biological tissues is not large. For these materials, σ is about 10^{-2} ohm^{-1} cm^{-1} at 10^9 cps. Although these data are not strictly comparable with the data in Table I, they serve to indicate the probable order of magnitude of σ for natural vegetable tissues in the microwave region. As judged from these data, it appears unlikely that the use of microwave frequencies for heating high-conductivity biological materials will permit any marked reduction in the voltage required, as compared to voltages required at frequencies of the order of 10^6 to 10^8 cps.

[15] L. T. Alexander and T. M. Shaw, "A method for determining ice-water relationships by measurements of dielectric constant changes," *Nature*, vol. 139, pp. 1109–1110; June 26, 1937.

[16] G. E. Vail and C. H. Bailey, "The state of water in colloidal gels: free and bound water in bread doughs," *Cereal Chem.*, vol. 17, pp. 397–417; July, 1940.

[17] P. Debye, "Polar Molecules," Chemical Catalog Company, New York, N. Y. 1929.

[18] C. N. Works, "Resonant cavities for dielectric measurements," *Jour. Appl. Phys.*, vol. 18, pp. 605–612; July, 1947.

[19] T. P. Kinn and J. Marcum, "Possible uses of microwaves for industrial heating," *Product Eng.*, vol. 18, pp. 137–140; January, 1947.

[20] P. W. Morse and H. E. Revercomb, "UHF heating of frozen foods," *Electronics*, vol. 20, pp. 85–89; October, 1947.

Factors Influencing the Perveance of Power-Output Triodes*

GORDON R. PARTRIDGE†, STUDENT, IRE

Summary—**A brief historical survey, discussing Child's formula and the assumptions upon which it was derived, is presented. Perveance is explained as being the factor relating voltage to current in electron tubes, and several formulas for calculating it are listed. The concept of "equivalent diodes" is discussed. Formulas are given for computing cathode current, together with reasons for errors in these formulas. Brief explanations are offered of the effects of grid current, variations in "effective anode area," and variations in μ upon calculations of cathode current.**

I. Introduction

In the design of power-output tubes, the problem arises of obtaining a current which, at a specified plate voltage, will enable the tube to deliver the power required by the designer. The current that may be obtained from a given tube is a complex function of tube geometry and electrode potentials, and no completely correct formulas for computing this current are known. This paper has a twofold purpose: first, to present a summary of known formulas and to recommend certain modifications which, the author believes, will give more satisfactory results in the design of power-output triodes than have heretofore been obtained; and second, to discuss factors that cause the performance of the tube to vary through its operating range, giving results that differ in practice from those predicted by design equations.

The earliest work on tubes involved studies of the unilateral conductivity of current between electrodes in low-pressure atmospheres. Later, in the early part of the present century, Child[1] developed the well-known "3/2-power" law, which is still the basis of studies of current flow between electrodes. Child developed his law from the following assumptions:

(1) The cathode is capable of emitting more electrons than will be drawn to the anode.

(2) The cathode and anode are parallel planes, large enough to render end effects negligible.

(3) Anode and cathode surfaces are equipotential.

(4) Electrons are emitted with zero initial velocity.

On the basis of these assumptions, Child arrived at the following law:

$$i_b = \frac{2.33 \times 10^{-6} A}{d^2} e_b^{3/2} \tag{1}$$

where i_b is the anode current, d is the distance between cathode and anode, A is the area of each electrode, and e_b is the potential in volts applied to the anode when the cathode is considered a surface at zero potential. Child's law may be written with greater generality as

$$i_b = K e_b^{3/2} \tag{2}$$

where the constant K in this case equals

$$\frac{2.33 \times 10^{-6} A}{d^2}.$$

For more complicated tubes where one grid is present, the plate current is given by the Child-Langmuir-Schottky law:

$$i_k = K[e_b + \mu(e_c + E_{cp})]^n \tag{3}$$

where the quantity in brackets is the voltage that will cause the same amount of current flow as would be obtained in an "equivalent diode" with the same potential applied to its anode. In (3), i_k is used to denote the total current from the cathode, e_b is the anode potential, e_c is the voltage applied to the grid, μ is the amplification factor[2,3], and E_{cp} combines the effects of contact potential between cathode and grid, the effects of initial velocity of the emitted electrons, and minor phenomena. The potential E_{cp} is usually of the order of a volt or two, and is negligible compared to other voltages encountered in transmitting triodes. The exponent n is of the order of 3/2 in value, but may be as low as 1.4 or as high as 5/2, the latter being its value in regions of large negative grid potential and very small plate currents. The quantity n is not constant for a given tube, but may vary between the limits indicated, depending upon the region of the tube characteristics where operation is taking place.

II. The Concept of Perveance

The most general statement of the relationship given in (3) is

$$i_k = K(e')^{3/2}. \tag{4}$$

The coefficient K in this equation and in (2) and (3) is called the *perveance* of the tube. This concept was first introduced, and the word coined, by Kusunose.[4] There

* Decimal classification: R334×R130. Original manuscript received by the Institute, February 5, 1948; revised manuscript received, April 19, 1948.

† Yale University, New Haven, Conn.

[1] C. D. Child, "Discharge from hot CaO," *Phys. Rev.*, vol. 32, pp. 492–511; May, 1911.

[2] B. Salzberg, "Formulas for the amplification factor of triodes," Proc. I.R.E., vol. 30, pp. 134–138; March, 1942.

[3] F. B. Vodges and F. R. Elder, "Formulas for the amplification constant of three element tubes," *Phys. Rev.*, vol. 24, pp. 683–689; December, 1924.

[4] Y. Kusunose, "Calculation of the characteristics and the design of triodes," Proc. I.R.E., vol. 17, pp. 1706–1750; October, 1929.

is no argument as to the voltage entering into equations (2) and (4). However, if current i_k for a triode is calculated from (3), the results will often depart greatly from those obtained by direct measurements. Therefore, several modified expressions have been suggested in the past to replace the parenthetic expression in (3), corresponding to e' in (4). One of these is

$$e' = e_b + \mu e_c, \tag{5}$$

which is even less accurate because E_{cp} is neglected, but still is often used because of its simplicity. Two others, for example, are

$$e' = \frac{e_c + \dfrac{e_b}{\mu}}{1 + \dfrac{1}{\mu}} \tag{6}$$ [5]

and

$$e' = \frac{e_c + \dfrac{e_b}{\mu}}{1 + \dfrac{1}{\mu} + \dfrac{4}{3\mu}\dfrac{(t_p - t_g)}{t_g}}. \tag{7}$$ [6]

The quantity t_p in the last expression is the distance between the inner surface of the anode and the centerline of the filament wires in a tube with a filamentary cathode (or the coated surface in a tube with a unipotential cathode). The distance t_g is the distance between the centerline of the filament wires (or the coated surface) and the centerline of the grid winding wires. If the filament, grid, or anode are tapered, the mean distances may be used. These dimensions are shown in Fig. 1(a).

The perveance having been defined, the problem becomes one of determining the most accurate formula for calculating this quantity. To obtain an equation that will be of any practical value whatsoever, a number of factors must be omitted from consideration. These are (1) space charge, and the associated phenomenon of "virtual" cathodes; (2) the initial velocity and angle of emission of the electrons; (3) end effects; (4) variation of potential along the cathode in filamentary cathode tubes; (5) the effects encountered as current ceases to be space-charge-limited and becomes limited by the ability of the cathode to emit electrons; (6) secondary emission; and (7) minor effects, such as ionization of traces of residual gas, accumulation of charge on the walls of glass-envelope tubes, and the Schottky effect[7] at very high plate potentials. All these phenomena have been investigated at various times, but their effects are either insignificant or too complicated to take into account. In practice, empirically derived formulas are normally used for perveance. Several of these will be listed and discussed. For each of the equivalent voltage formulas given above, there is a perveance formula that gives the best results. It has been the practice to use the following set of formulas:

$$i_k = K(e_b + \mu e_c)^{3/2} \tag{8}$$

where

$$K = \frac{2.33 \times 10^{-6} A}{(t_p^{4/3} + \mu t_g^{4/3})^{3/2}} \tag{9}$$

and

$$A = Nlt_e \tag{10}$$

where N is the number of filament strands and l is the "lighted length" of these strands.

$$t_e = 2\left(t_g + \frac{t_p - t_g}{\mu + 1}\right). \tag{11}$$

The formula for A was derived on the basis of the fact that the entire anode is not effective in gathering electrons. Those parts of the anode lying nearest the filament wires collect more electrons than those which are opposite the vacant spaces between wires.

It will be observed that (9) is for plane-parallel electrode configurations, rather than for the concentric cylindrical electrodes employed in the tubes here discussed. The justification for using this formula lies in the fact that power-output triodes normally have electrodes of sufficiently large radii so that, over a small sector of the tube, the electrodes are essentially plane and parallel.

The set of formulas of (8), (9), and (10) give fair results, with errors normally in the neighborhood of 20 to 30 per cent when average measured results for a given tube type are taken as the standard of reference. This paper will recommend certain changes in these formulas which appear to give better results in the majority of cases. However, it must be stated emphatically that no one formula is best in all cases, and, in general, more difficulty is to be expected in determining and accurate value of perveance in cases of positive-grid operation than when the grid is negative.

In the discussion to follow, (7) will be used for calculating e'. Upon comparing this with (5), it will be seen that the value of i_k given by (4) will be approximately $(1/\mu)^{3/2}$ as great as if (5) were substituted in (4). For this reason, the perveance formula (9) must be multiplied by $\mu^{3/2}$, or, what amounts to the same thing, the denominator of (9) must be divided by $\mu^{3/2}$. Such an operation gives the following equation:

$$K = \frac{2.33 \times 10^{-6} A}{\left(\dfrac{t_p^{4/3}}{\mu} + t_g^{4/3}\right)^{3/2}}. \tag{12}$$

[5] E. L. Chaffee, "Theory of Thermionic Vacuum Tubes," 1st. ed. McGraw-Hill Book Co., New York, N. Y., 1933; pp. 185–190.
[6] F. B. Llewellyn and L. C. Peterson, "Vacuum tube networks" Proc. I.R.E., vol. 32, pp. 144–167; March, 1944.
[7] W. Schottky, "Über den Einfluss von Strukturwirkungen, Besonders der Thomsonschen Bildkraft, auf die Elektronenemission der Metalle," *Phys. Zeit.*, vol. 15, pp. 872–878; November, 1914.

This value of perveance was tried in combination with the e' of (7), and gave fairly satisfactory results. It was found, by trial and error, that some additional modification led to a slight improvement. This revision gives

$$K = \frac{2.33 \times 10^{-6} A}{\left(\dfrac{t_p^{4/3}}{\mu + 1} + t_g^{4/3}\right)^{3/2}}. \qquad (13)$$

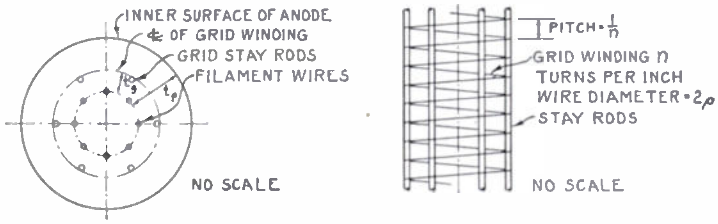
The perveance given by (13) used with the value of e' given by (7) gives accurate results when i_k is calculated for cases of negative grid voltages. The use of (13) is not successful in cases of positive grid voltages.

Equation (13) is essentially a formula for plane-parallel electrode configurations. As the grid voltage becomes positive and cathode current increases, the virtual cathode will move back more nearly to the position of the filament itself. As this happens, the approximation of considering the tube a parallel-plane structure becomes less justifiable, and a perveance formula somewhat more similar to that used for cylindrical tubes is found useful. The following equation is recommended:

$$K = \frac{2.33 \times 10^{-6} A}{t_g^2}. \qquad (14)$$

Its use gives acceptable results only in positive-grid regions of tube operation. It should not be employed when the grid is at zero or negative potentials. Table I summarizes the formulas, and lists conditions under which they are applicable.

There are many physical limitations and factors that cause the formulas of Table I to be in error. For certain tube types, such as the GL-9C24 and GL-889, none of the formulas gave acceptable results with positive grid potentials. This is because the pure-tungsten filaments cannot emit enough electrons to provide the current predicted by the formulas. The same difficulty, although not so pronounced, arises with the GL-862A, which also has a pure-tungsten filament. With these three types, (16), which gives the lowest value of current for a given set of electrode potentials, provides the best results, although its use is not normally indicated for positive-grid regions. In tubes with thoriated-tungsten filaments, where plenty of electron emission is available, (17) gives consistently superior results, with errors of the order of 10 per cent or less, except when the grid potential is near the maximum value it may take without endangering the life of the tube.



(a) (b)

Fig. 1(a)—Plan view of a power-output triode. (Note: In the case of taper, take mean values of t_g and t_p).
Fig. 1(b)—Detail of grid.

Some more or less typical results will now be shown for the tubes for which plate characteristics were calculated. Fig. 2 shows the calculated plate characteristics of the GL-892, as computed by (15), (16) and (17), and also the published data. Note the greater accuracy given by (17) in the positive-grid region. Fig. 3 (for the type GL-889) shows the superiority of (16) in the negative-grid region, and Fig. 4 (for the GL-5513) again shows the better results given by (17) when the grid is positive

TABLE I

Summary of Perveance and Current Formulas

$i_k = Ke'^{3/2}$	Conditions of Applicability	Equation
$i_k = \dfrac{2.33 \times 10^{-6} A}{(t_p^{4/3} + \mu t_g^{4/3})^{3/2}} (e_b + \mu e_c)^{3/2}$	Fair accuracy in negative-grid regions of operation. Errors of the order of 20 to 30 per cent. Of questionable value in positive-grid regions.	(15)
$i_k = \dfrac{2.33 \times 10^{-6} A}{\left[\dfrac{t_p^{4/3}}{\mu+1} + t_g^{4/3}\right]^{3/2}} \left[\dfrac{e_c + \dfrac{e_b}{\mu}}{1 + \dfrac{1}{\mu} + \dfrac{4}{3\mu}\left[\dfrac{t_p - t_g}{t_g}\right]}\right]^{3/2}$	Usually gives better results in negative-grid regions than (15). Of little value in computations when grid is positive.	(16)
$i_k = \dfrac{2.33 \times 10^{-6} A}{t_g^2} \left[\dfrac{e_c + \dfrac{e_b}{\mu}}{1 + \dfrac{1}{\mu} + \dfrac{4}{3\mu}\left(\dfrac{t_p - t_g}{t_g}\right)}\right]^{3/2}$	Gives good results when grid is positive. Errors are of the order of 10 per cent. Not to be used when grid potential is zero or negative. See discussion in text.	(17)

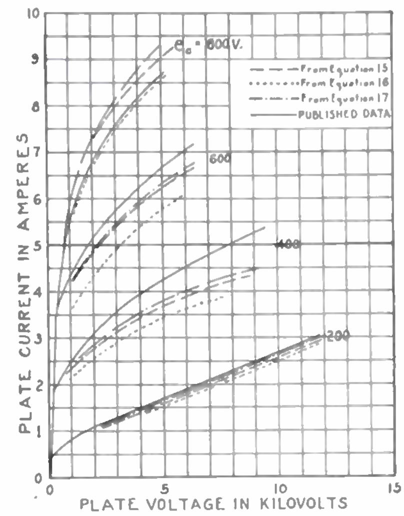


Fig. 2—Plate characteristics of the type GL-892, comparing published and calculated currents and illustrating improved results obtained in regions of positive grid potential by equation (17).

It must be repeated, however, that no one formula gives best results in all cases. The recommended expressions

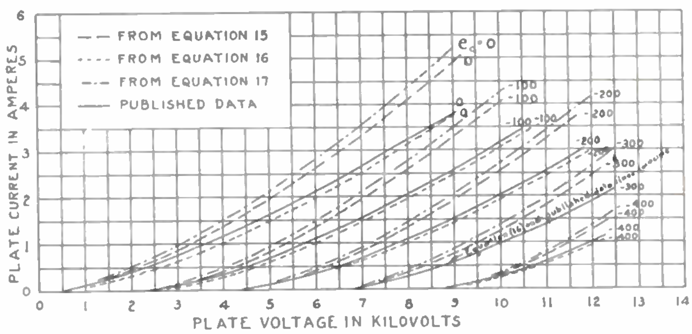


Fig. 3—Plate characteristics of the type GL-889 in the negative-grid region, comparing published data with calculated currents and illustrating improved results obtained by equation (16).

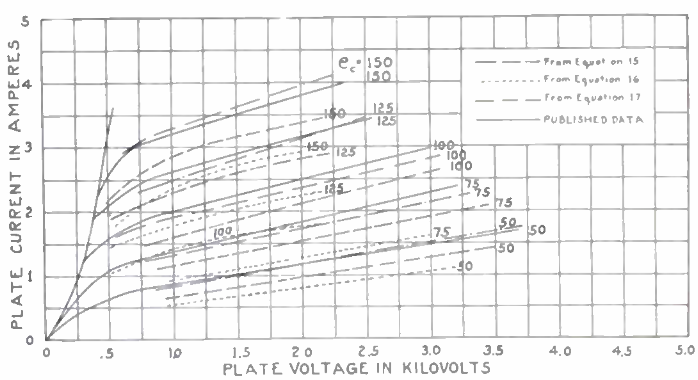


Fig. 4—Plate characteristics of the type GL-5513 comparing published and calculated currents and illustrating improved results obtained by equation (17).

are better in a majority of cases, but since complete calculations were performed on only eight tubes (types GL-862A, GL-880, GL-889, GL-891, GL-892, GL-5513, GL-9C24, and Z-1463, a developmental tube), it would be incorrect to state completely general conclusions.

III. The Problem of Grid Currents

It will be observed that the equations given above are for i_k rather than i_p. This makes no difference when the grid is negative, since in this case $i_k = i_p$ for all practical purposes. But usually the tube designer is interested primarily in plate current, and (15), (16), and (17) may be considerably in error if used to calculate plate current when the grid draws current. The results presented in Figs. 2, 3, and 4 were obtained by subtracting published values of grid current from the calculated values of i_k. In the design of new tube types, grid currents are unknown and must be estimated. Methods for predicting grid current have been presented by various authors, but in general without great success. Jonker and Tellegen[8] have presented several formulas which, while not giving the exact value of grid current, will be of some help in making an estimate of its value. Their formulas, with notation modified to correspond to that used above, are

$$\frac{i_c}{i_k} = \frac{2\rho}{1/n}\sqrt{\frac{e_c}{e'}} \tag{18}$$

where 2ρ is the diameter of the grid-winding wire, n is the number of turns per inch of grid winding, e_c is the potential of the grid, and e' is given by

$$e' = \frac{\mu e_c + e_b}{1 + \mu + \dfrac{t_p - t_g}{t_g}}. \tag{19}$$

Note the great similarity between (19) and (7). Jonker gives an equation for μ, also, as follows:

$$\mu = \frac{2\pi n(t_p - t_g)}{\log_e\left(\dfrac{1}{2\pi n\rho}\right)} \tag{20}$$

This is the formula given by Vodges and Elder[3] for the case where $\rho \to 0$. The subject of formulas for calculating μ will be discussed at a later point in this paper.

Equation (18) is a first approximation, and somewhat greater accuracy is claimed for a modified form also given by Jonker and Tellegen:

$$\frac{i_c}{i_k} = 2n\rho\sqrt{\frac{e_c}{e'}}\left[1 + \frac{e_c - e'}{2e'\log_e\left(\dfrac{1}{2\pi n\rho}\right)}\right]. \tag{21}$$

Grid currents have been calculated for two tubes, the GL-892 and the GL-5513, using (21). The results are presented as Figs. 5 and 6. It will be observed that the accuracy is very low, indeed. At least a part of the error arises from the fact that the formula does not consider

[8] J. L. H. Jonker and B. D. H. Tellegen, "The current to a positive grid in electron tubes," *Philips Res. Rep.*, vol. 1, p. 13, 1945.

secondary emission.[9] The GL-892, which has a tantalum grid, has greater secondary emission than the GL-5513, which has a platinum-clad molybdenum grid. Because of this higher secondary emission, the grid current will drop off faster with rising plate voltage than it does in the GL-5513, a fact confirmed by published data.

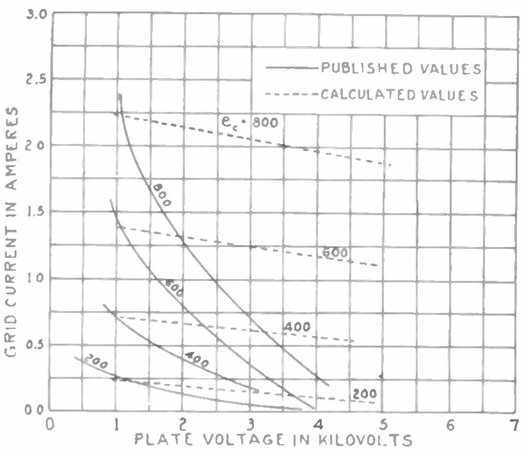


Fig. 5—Comparison between published values of grid currents for the type GL-892, and values computed from the formula of Jonker and Tellegen.

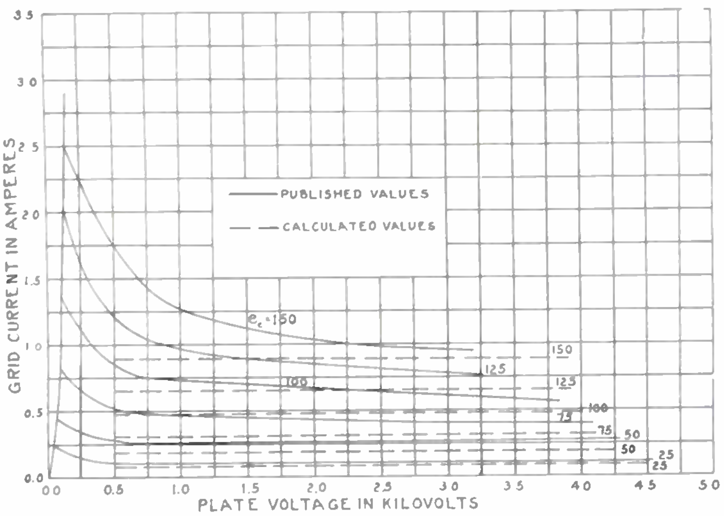


Fig. 6—Comparison between published values of grid current for the type GL-5513 and values computed from the formula of Jonker and Tellegen.

In an effort to determine a more accurate empirical method of calculating grid current, it was found that plotting the ratio of i_c/i_k as a function of e_b/e_c on log-log paper gave a group of points that were, in most cases, scattered in such a way that a straight line could be drawn through the center of the group. This is shown in Fig. 7. The nature of the curve suggests that grid current could be determined by a formula of the type

$$\frac{i_c}{i_k} = m\left(\frac{e_b}{e_c}\right)^b, \tag{22}$$

where m and b are constants for a particular tube, and depend for their values upon the tube geometry and upon secondary-emission properties of the grid material. Table II lists values of m and b for three tubes for which grid current was calculated from (22).

TABLE II

Tube Type	Value of m	Value of b
GL-9C24	0.317	−0.64
GL-880	0.233	−0.82
GL-5513	0.534	−0.305

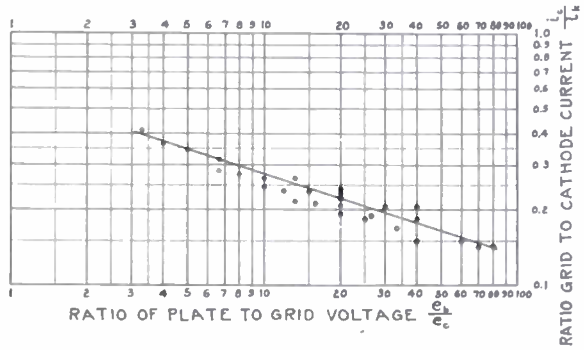


Fig. 7—Ratio of i_c/i_k plotted as a function of the ratio e_b/e_c for the type GL-5513 tube.

The points were so scattered in the case of the GL-9C24 that several straight lines appeared equally plausible, but the method appears to have some degree of validity with the GL-880 and the GL-5513. It is suggested that further study of an equation of the form of (22) would be useful in a more detailed investigation of the effects of grid current upon tube performance.

Formulas (16) and (17) are, it is believed, capable of giving better results, when the conditions of applicability listed in Table I are followed, than other formulas heretofore employed. However, even these formulas are approximations, and the results may be appreciably in error. Some of the causes of these errors will now be discussed.

IV. Effective Anode Area

At high anode potentials, there is a tendency for the electrons crossing from the cathode to the anode to form coarsely defined "beams," especially when the grid is at a negative potential. As the plate voltage becomes higher, the beam becomes sharper, leading to a decrease in the effective (i.e., the electron-collecting) area of the anode. Since the equations of i_k (15, 16, and 17) do not consider changes in the effective area, it follows that the value of A appearing in these equations will be too large when the beam is very narrow, and too small when the beam is wide. Hence, the value calculated for i_k will be either too large at high plate voltages or too small at low plate voltages. If the former is true, the per cent error will increase as plate voltage rises; if the latter is true, the per cent error will be large at low plate voltages, and decrease with an increase of plate voltage. Attempts were made to derive empirical correction fac-

[9] H. Bruining, "Die Secondar-emission Fester Korper," J. Springer, Berlin, 1942; Edwards Bros., Inc., Ann Arbor, Mich., 1944.

tors for the variation in effective anode area. However, while correction factors could be made to fit any one tube, they were of such complexity as to be completely useless in design work. Furthermore, there was no apparent tendency towards generalization. A formula that worked for one tube bore little or no resemblance to one that worked for another. Consequently, attempts to find such correction factors were abandoned in the belief that the dubious usefulness of the results and the lack of generality did not warrant the expenditure of additional time.

The effects of beaming in triodes might conceivably be calculated from the Laplace equation[10]

$$\nabla^2 V = 0, \tag{23}$$

or from Poisson's equation

$$\nabla^2 V = -4\pi\rho, \tag{24}$$

but the labor involved in fitting the boundary conditions for electrodes as complicated as grid and filament wires is prohibitive, if, indeed, the job is possible at all. Furthermore, a knowledge of charge distribution is required for the solution of Poisson's equation, which complicates the problem even further. Other possibilities for determining the beaming effect would be the plotting of electron trajectories from fields obtained in an electrolytic flux-plotting tank, or the beams may be seen by coating the anode of a tube with willemite or other phosphors.

V. Variations in Mu

In addition to variations in effective anode area, the fact that μ is not constant in all parts of the tube characteristics complicates the problem of determining perveance. Mu may take values ranging from slightly greater than the mean value to zero, and may even become slightly negative in some parts of the characteristic curves. Properly speaking, the value of μ varies with electrode potentials and currents. However, for purposes of mathematical treatment it is convenient to think of plate current as a function of μ, and this approach will be used in subsequent discussion. If (17) is written out in full, such constants as A and K being replaced by their values given in (10) and (14), the following expression is obtained:

$$i_k = \left[\frac{2.33 \times 10^{-6}}{t_g^2} \times 2Nl\left(t_g + \frac{t_p - t_g}{\mu + 1}\right)\right]\left[\frac{e_c + e_b/\mu}{1 + \frac{1}{\mu} + \frac{4a}{3\mu}}\right]^{3/2} \tag{25}$$

where

$$a = \frac{t_p - t_g}{t_g}$$

and is introduced solely to simplify writing this and following equations.

Taking the partial derivative of i_k with respect to μ gives

$$\frac{\partial i_k}{\partial \mu} = \frac{3K}{2}\left[\frac{\left\{1 + \frac{1}{\mu} + \frac{4a}{3\mu}\right\}\left\{\frac{-e_b}{\mu^2}\right\} - \left\{e_c + \frac{e_b}{\mu}\right\}\left\{\frac{-1}{\mu^2} - \frac{4a}{3\mu^2}\right\}}{\left\{1 + \frac{1}{\mu} + \frac{4a}{3\mu}\right\}^2}\right]\left[\frac{e_c + \frac{e_b}{\mu}}{1 + \frac{1}{\mu} + \frac{4a}{3\mu}}\right]^{1/2}$$
$$+ \left[\frac{e_c + \frac{e_b}{\mu}}{1 + \frac{1}{\mu} + \frac{4a}{3\mu}}\right]^{3/2}\left[\left\{\frac{2.33 \times 10^{-6}}{t_g^2}\right\} 2Nl(t_p - t_g)\left\{\frac{-1}{(\mu + 1)^2}\right\}\right]. \tag{26}$$

This equation will normally have a negative value, indicating that the plate current increases as the tube is made to operate in regions of small μ, a result observed universally in practice.

The question might be raised whether or not a tube could be designed in which variations in μ would have no effect upon the perveance. This would require that the partial derivative in (26) be equated to zero. Setting up this equality and simplifying the expression as much as possible gives

$$a\left(e_c + \frac{e_b}{\mu}\right) = \frac{3}{2}\left(t_g + \frac{t_p - t_g}{\mu + 1}\right)\left[\frac{-e_b}{\mu^2} - \frac{\left\{e_c + \frac{e_b}{\mu}\right\}\left\{\frac{-1}{\mu^2} - \frac{4a}{3\mu^2}\right\}}{1 + \frac{1}{\mu} + \frac{4a}{3\mu}}\right]. \tag{27}$$

This equation can hold true only for certain special values of electrode voltages and spacings, and so it must be

[10] S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, Inc., New York, N. Y., 1944; p. 64.

concluded that, in general, variations of i_k with μ cannot be eliminated in tube design.

Since, therefore, variations in μ must be considered, the problem next arises as to how to take them into account. To investigate this question, the values of μ were determined for various tubes by measuring the slope of the constant-current curves at different values of plate and grid potentials. These values were plotted, and an attempt was made to set down equations that would give approximately the same sets of curves. The following equation gave fairly good results:

$$\mu = \mu_{\max}(1 - \mathcal{E}^{-k(e_b/e_c)}) \tag{28}$$

Plotting families of curves for μ gave results which were very close to those observed from measuring the slopes of the constant-current curves. A comparison of measured and calculated curves for the type GL-889 is given in Fig. 8. The value of k is 1.1. This varies from tube to tube, and no convenient correlation was found between the value of k and the dimensions of the tube. It should be noted that appreciable variations in the value of μ are observed only in regions of positive grid potential. When the grid is at a negative potential, the value of μ stays very nearly constant, and is equal approximately to the maximum value of μ for the given tube type.

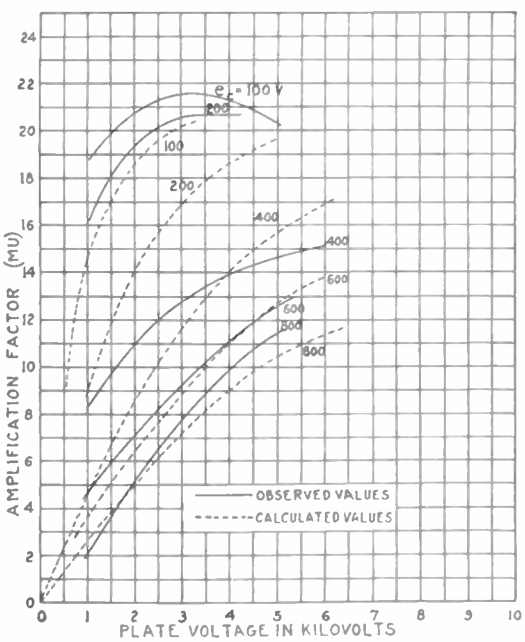


Fig. 8—Comparison between observed and calculated values of μ in tube type GL-889.

When the variable values of μ are introduced into the plate-current equation (17), it is found that the calculated currents are greatly in error, especially when the grid voltage is very positive. The cause of this large error will be explained on the basis of the nature of the plate-current equation.

Assume that the perveance portion of (17) remains constant, and that only the quantity in brackets changes as μ varies in value. This is not strictly true, but leads to a simplified analysis. Actually, as μ decreases in value, the value of A calculated from (10) and (11) will increase, and hence the perveance increases. It will be shown that, as μ becomes smaller, the bracketed quantity in (17) increases in value and becomes a maximum when μ is zero. Taking the partial derivative of this expression gives

$$\frac{\partial}{\partial\mu}\left[\frac{e_c + \dfrac{e_b}{\mu^2}}{1 + \dfrac{1}{\mu} + \dfrac{4a}{3\mu}}\right] = \frac{\left[1 + \dfrac{1}{\mu} + \dfrac{4a}{3\mu}\right]\left[\dfrac{-e_b}{\mu}\right] - \left[e_c + \dfrac{e_b}{\mu}\right]\left[\left(1 + \dfrac{4a}{3}\right)\left(\dfrac{-1}{\mu^2}\right)\right]}{\left[1 + \dfrac{1}{\mu} + \dfrac{4a}{3\mu}\right]^2} = 0 \tag{29}$$

the derivative being equal to zero when the value of the expression being differentiated is a maximum. Simplifying (29) gives

$$\mu = \frac{\mu e_c}{e_b} + \frac{4a\mu e_c}{3e_b}. \tag{30}$$

Equation (30) can be satisfied in two cases:

$$\text{(case 1)} \qquad e_c = \frac{e_b}{1 + \dfrac{4a}{3}} \tag{31}$$

and

$$\text{(case 2)} \qquad \mu = 0. \tag{32}$$

Normally case 2 would be a trivial solution, but it has significance here in that a value of zero for μ locates a point on the tube characteristics where the current is very high, and close to the maximum value that can be obtained in any given tube.

The expression indicated as case 1 gives a relationship between e_b and e_c that locates a point of low μ on the characteristic curves. In general, the value of μ is not zero at this point, presumably because (17), on the basis of which the above discussion is derived, is only an approximation, although the best approximation available at present.

In any event, both case 1 and case 2 define conditions in which the plate current is near the maximum value it can take in any given tube. Moreover, the value of the perveance will be very large when $\mu = 0$, as explained above, so that currents calculated using the variable values of μ will be exceedingly high when μ is small.

These calculated currents are usually several times the values observed in the tubes themselves. Before such high currents could be reached, the inability of the cathode to emit sufficient electrons will limit the current in the tube. Therefore, it is recommended that the mean value (i.e., the nominal value) of μ be used in all calculations, even though a consideration of the variations in μ might be theoretically justifiable.

In the design of new tubes where μ has yet to be determined, the value may be calculated with fair accuracy by the use of Vodges and Elder's formula. Certain charts are also available for determining the value of μ from known dimensions. Vodges and Elder's formula is

$$\mu = \frac{2\pi nb - \log_e \cosh 2\pi n\rho}{\log_e \coth 2\pi n\rho} \tag{33}$$

where

$$b = t_p - t_g. \tag{34}$$

A summary of the results obtained from this formula is given in Table III.

Vodges and Elder's formula is for plane-parallel tube structures, but gives consistently good results with the cylindrical configurations used in large power-output triodes.

In conclusion, the author wishes to state that (16) and (17) appear to be more useful for tube design purposes than equations heretofore used, and to express the hope that other information included in this paper may be of use in guiding future investigators who wish to prepare a more detailed study of the exact causes of variation of perveance in tubes.

TABLE III

COMPARISON OF CALCULATED AND PUBLISHED VALUES OF MU, WHEN VODGES AND ELDER'S FORMULA IS USED

Tube Type	Calculated value	Published value
GL-9C24	18.8	21
GL-862A	43.1	45
GL-880	18.8	20
GL-889	20.3	21
GL-891	3.0	8
GL-892	42	50
GL-5513	104	87
Z-1463	25.8	25

ACKNOWLEDGMENT

The author wishes to express his appreciation to T. A. Elder, K. C. DeWalt, and other members of the staff of the Electronics Department, Tube Division, of the General Electric Company, for valuable advice and assistance during the studies which led to the preparation of this paper. The material discussed herein was prepared at the General Electric Company's Schenectady Works, and first appeared as a report to the General Electric Company Tube Division.

A Coaxial-Line Support for 0 to 4000 Mc*

R. W. CORNES†, VOTING ASSOCIATE, IRE

Summary—**Experimental and theoretical work done in designing a broad-band undercut-bead support is described, and a method is given for calculating the bandwidth. A typical microwave application is discussed which has a VSWR of less than 1.025 for 0 to 4000 Mc.**

INTRODUCTION

RIGID AIR- or gas-filled coaxial lines have been used considerably in recent years as high-frequency transmission lines. In these lines, the center conductor has generally been supported by spaced dielectric beads or by the use of quarter-wave stubs. Both methods limit the usefulness of the lines to a relatively narrow frequency band because of the impedance mismatch and resultant power reflection at frequencies deviating slightly from the design frequency.

One method of reducing the reflection from a dielectric bead in a line is to undercut the line so that the beads have the same characteristic impedance as the rest of the line. Such a bead will produce no reflection at very low frequencies. However, at high frequencies the discontinuity effects due to the undercut are great enough to upset the impedance relations between line and bead. Experimental work was done to determine the changes necessary to take this effect into account, and the analysis based on filter theory is given along with design considerations as a result of this work. It is possible to

* Decimal classification: R117.2. Original manuscript received by the Institute, June 12, 1947; revised manuscript received, August 2, 1948. Presented, 1947 IRE National Convention, March 5, 1947, New York, N. Y.

† Microwave Measurements Laboratory, Sperry Gyroscope Company, Great Neck, L. I., N. Y.

design beads which are good up to the maximum usable frequency of a coaxial line. In the $\frac{7}{8}$-inch-diameter line which is used considerably in radar work, the operating frequencies extend from 0 to 4000 Mc.

Description

The improved undercut bead was first developed for use in test equipment for $\frac{7}{8}$-inch coaxial line. This equipment was being used over the range of 600 to 4000 Mc. When spaced beads were used, it was necessary to employ at least ten different sets of equipment, each with different bead spacing, to cover this band. It was to avoid this duplication of equipment that the work on undercut beads was undertaken.

Fig. 1 shows a section of coaxial line with an undercut bead in place. The dimensions given are for $\frac{7}{8}$-inch

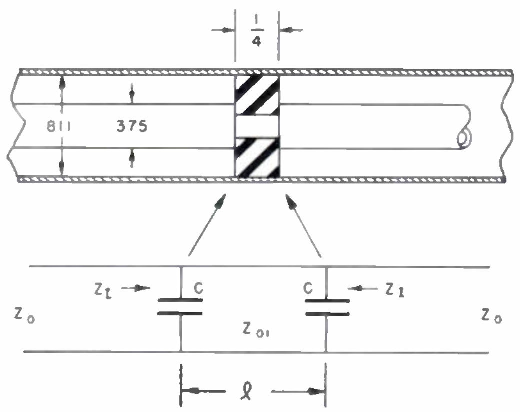


Fig. 1—A coaxial line with an undercut bead, and the equivalent circuit of the bead.

line with a characteristic impedance of 46.3 ohms. The line is similar in size to the Army-Navy RG-44/U line, and has the Army-Navy designation of RG-92/U or CG-462/U when made up in definite lengths.

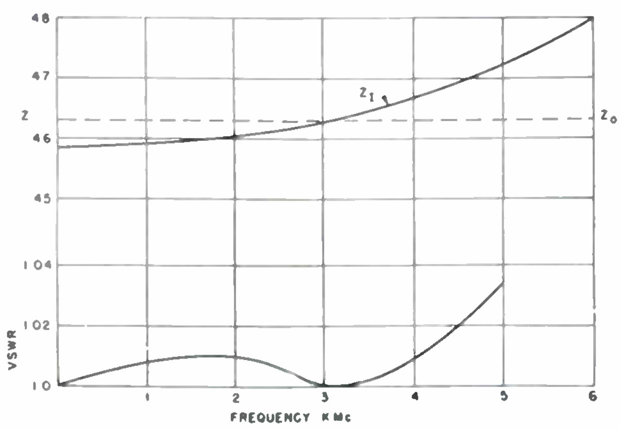


Fig. 2—The image impedance and VSWR of an undercut bead in $\frac{7}{8}$-inch coaxial line.

The lower curve in Fig. 2 shows the calculated voltage-standing-wave ratio, or VSWR, introduced into a line. The measured VSWR is slightly higher, but is still below 1.02 up to about 4000 Mc. This corresponds to a maximum power reflection of 1/10 of 1 per cent.

Because of the low reflection from each bead, multiple beads may be used without adversely affecting the performance of the line at any frequency below the maximum design frequency of several thousand megacycles. However, for best performance on long lines, the beads should not be so spaced (approximately $\frac{1}{2}$-wavelength apart) that the reflections from the individual beads will add at the operating frequency.

Fig. 3 is an illustration of an impedance meter or standing-wave detector for the $\frac{7}{8}$-inch coaxial line which

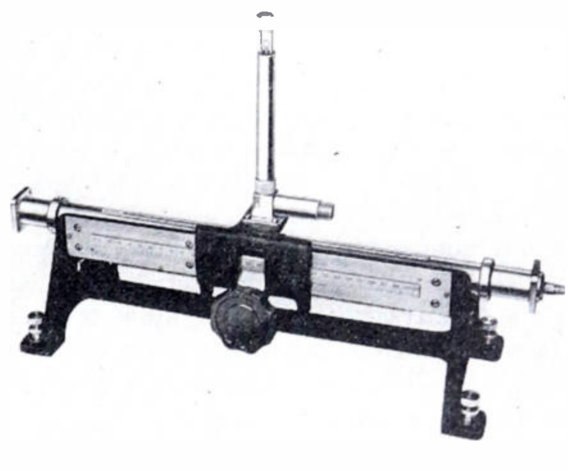


Fig. 3—A coaxial impedance meter or slotted line using undercut beads.

has an undercut bead at each end. The beads are secured in place by dielectric pins through the rings near the couplings.

Theory

A bead in a coaxial line behaves as a low-pass filter in the line, and it is convenient to use filter theory in analyzing the performance of an undercut bead. Fig. 1 shows a bead and its equivalent electrical circuit. This equivalent circuit is similar to a conventional pi-section filter.

The capacitance C is the discontinuity capacitance due to the change in diameter of the center conductor. The diameter change sets up fringing fields and higher modes which are equivalent to a capacitance across the line. Methods of calculating discontinuity capacitances are given by Whinnery, Jamieson, and Robbins.[1,2] Al-

[1] J. R. Whinnery, and H. W. Jamieson, "Equivalent circuits for discontinuities in transmission lines," Proc. I.R.E., vol. 32, pp. 98–115; February, 1944.

[2] J. R. Whinnery, H. W. Jamieson, and T. E. Robbins, "Coaxial line discontinuities," Proc. I.R.E., vol. 32, pp. 695–709; November, 1944.

though the discontinuity capacitance is small—of the order of 0.05 $\mu\mu$f—at high frequencies it represents a moderately large shunt admittance.

Equations describing a filter of this type have been developed from transmission-line equations.[3] The equation for the image impedance Z_I is of particular interest.

$$Z_I = \frac{Z_{01}}{\sqrt{1 + 2\omega C Z_{01} \cot \beta l - \omega^2 C^2 Z_{01}^2}} \tag{1}$$

where

Z_{01} = bead characteristic impedance
C = discontinuity capacitance
β = phase constant in the dielectric
l = length of the bead.

The variation of image impedance with frequency as calculated from (1) is plotted in Fig. 2 for a bead in a $\frac{7}{8}$-inch coaxial line. The image impedance changes very slowly over a very broad frequency range The lower solid curve in Fig. 2 shows the variation of VSWR introduced into a line by a bead whose image impedance varies as is shown in the upper curve.

For minimum reflection from a bead, its image impedance should equal the characteristic impedance of the coaxial line. It is possible to do this at one frequency only. However, with proper design the image impedance can be made to differ only slightly from the line impedance up to some maximum frequency.

In order to make the image impedance equal to the line impedance, it is necessary to design the bead so that its impedance Z_{01} is greater than the line impedance. By doing this, the increased impedance within the bead offsets the discontinuity capacitances. It is this higher value of Z_{01} which gives the improved performance over that obtained with a bead designed to equal the line impedance.

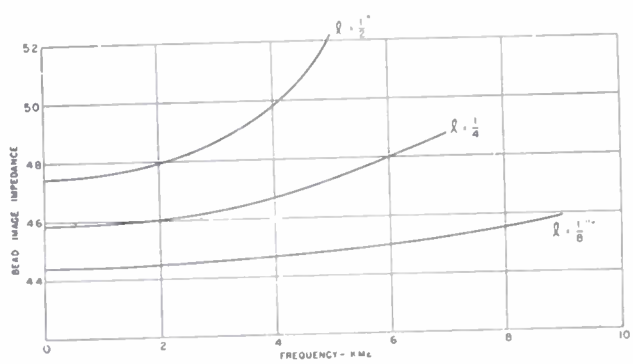


Fig. 4—The image impedance of undercut beads for various bead lengths, $\frac{7}{8}$-inch coaxial line.

The principal factor in determining the average image impedance of a bead over a band is the bead characteristic impedance. This is determined by the depth of the undercut in the center conductor and the dielectric constant. The rate of change of image impedance with frequency is determined largely by the change of cot βl with frequency. If l is large, the change is more rapid than with l small. Fig. 4 shows the variation in image impedance for three beads which are identical except in length.

Design Considerations

In deciding on a bead length, it is necessary to compromise between two opposing factors. Short beads give higher cutoff frequencies, but the length chosen must give sufficient mechanical support to the center conductor.

The dielectric constant of the bead material has a considerable effect on the design. For any given bead dimensions, the discontinuity capacitance is proportional to the dielectric constant of the bead. Also, for a given bead characteristic impedance, the depth of undercut increases with increasing dielectric constant. This causes a further increase in discontinuity capacitance and reduction of bandwidth.

The electrical length of a bead is proportional to the square root of the dielectric constant. Since it is desirable to have a long mechanical length and short electrical length, material with a low dielectric constant is preferable. For these reasons, Teflon with a dielectric constant of about 2 has been used in the $\frac{7}{8}$-inch RG-92/U line.

The characteristic impedance of the bead is the final factor to choose. As was pointed out earlier, this impedance is the most important factor in determining the proper image impedance. To obtain maximum bandwidth, the image impedance should be less than the characteristic impedance of the line at zero frequency, and equal to the characteristic impedance at a point near the maximum operating frequency of the line. It has been found that, for maximum bandwidth, the image impedance of the bead at zero frequency should be approximately 99 per cent of the line impedance. The image impedance at zero frequency can be found from (1) by letting the frequency approach zero. Then

$$Z_I \Big|_{f=0} = \frac{Z_{01}}{\sqrt{1 + \dfrac{2 Z_{01} C v}{l}}} \tag{2}$$

where v is the velocity of propagation within the dielectric.

Since the image impedance is calculated from the bead characteristic impedance, discontinuity capacitance, and bead length, it is necessary to start with some estimated value of bead characteristic impedance. A bead characteristic impedance of 1.08 times the line impedance has been found to give good results.

[3] R. B. Muchmore, "Microwave Filter Design," Report No. 5224.1069, Sperry Gyroscope Company; September, 1946.

With these design considerations in mind, one can proceed to design a bead. The dielectric material for the bead should be chosen, and should be a material with low losses. Losses were not considered in the theoretical analysis, since dielectrics with negligible losses even at the highest operating frequencies of coaxial lines are available.

After the bead length has been decided, the discontinuity capacitances at each end of the bead should be calculated from the formulas of Whinnery and Jamieson.[1,2] In very short beads there is interaction between the two capacitances which tends to decrease the value of each capacitance. This effect should be considered in determining the discontinuity capacitance.

From the bead length, characteristic impedance, and discontinuity capacitance, it is possible to calculate the image impedance from (1) and (2). The impedance at zero frequency should first be calculated. If this comes out close to 99 per cent of the line impedance, the bead dimensions are approximately correct, and the impedance should be calculated for a few frequencies in the desired band. If the calculated image impedance varies symmetrically about the characteristic impedance of the line, then the various values chosen have been correct.[4]

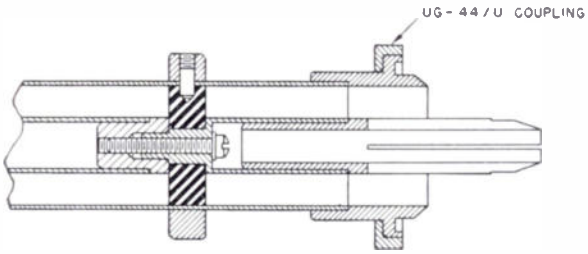


Fig. 5—An undercut bead as used in RG-92/U coaxial line.

One objection to the use of undercut beads over plain spaced beads is the increased difficulty in installing the beads. One method is shown in Fig. 5. Here a bead is shown at one end of a line with a standard coupling in place. Two small holes may be used in the bead to allow the free passage of gas through the line, in case pressurization is desired. This method of mounting has proved quite satisfactory. Another method which could be used is to split the beads to avoid the necessity of parting the center conductor at the undercut.

Method of Test

Since the reflections introduced in a line by a properly designed undercut bead are very small, it is necessary to use special techniques in making measurements of bead characteristics. It is difficult to measure accurately voltage-standing-wave ratios of the order of 1.02 with slotted sections. In making measurements on these beads a method called the "resonance-cavity method" has generally been used.[5] It is possible to make measurements to low VSWR's to an accuracy of one-half per cent or better with this method. It is also possible to calculate the reactance of the bead from such measurements. This method has been used by several laboratories, and provides a relatively rapid method for accurate measurements.

Measurements have been made using the resonance method on beads for $\frac{5}{8}$- and $\frac{7}{8}$-inch coaxial line, and the results have agreed very closely with theoretical predictions.

Conclusion

The feasibility of building coaxial lines which can be used at any frequency up to several thousand megacycles presents new possibilities for broad-banding of equipment. These lines have been used in a large number of test instruments and system components, and have proved very satisfactory.

Acknowledgments

The theoretical analysis of the undercut bead was developed from filter theory by R. B. Muchmore. The author wishes to acknowledge the work done by Mr. Muchmore and J. C. Hupcey in developing a satisfactory bead, and to express his appreciation to many others who assisted in this project.

[4] The reflection from a bead may be calculated from formulas in footnote reference 3 or from formulas 26-48, 26-53, and 26-54 in "Very-High-Frequency Techniques," Radio Research Laboratory, Harvard University, McGraw-Hill Book Co., New York, N.Y.; 1947.

[5] W. H. Pickering, O. W. Hagelbarger, C. V. Meng, S. C. Snowden, "A New Method for the Precision Measurement of Waveguide Discontinuities," NDRC, Division 14, Report No. 317, October, 1944.

Contributors to Waves and Electrons Section

Randolph W. Cornes (S'36–A'38–VA'39) was born in Los Angeles, Calif.,on August 9, 1913. He received the B.A. degree in physics in 1935 from the University of California. After one-half year of graduate study, he joined the service department of the General Electric X-Ray Corporation, in San Francisco, Calif. From November, 1937, until December, 1942, Mr. Cornes was associated with the International Geophysics Company, Inc., in Los Angeles, Calif., and later with the Halliburton Oil Well Cementing Company, in Duncan, Okla., where he was engaged in developing geophysical equipment.

R. W. Cornes

In 1942 Mr. Cornes joined the Mare Island Navy Yard, where he was in charge of the servicing and installation of the magnetic measurement equipment in use. Since March, 1945, he has been employed by the Sperry Gyroscope Company, working on the development and production of microwave measurement equipment and systems components.

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James A. Galvin was born on March 22, 1917, at San Diego, Calif. He received the B.S. degree in mathematics and physics from the University of Nevada in 1938, and did postgraduate work in mathematics and physics at the University of California during 1939 and 1940. Since 1941 he has been employed by the Western Regional Research Laboratory as a physicist. He was previously employed as a school teacher in Nevada and San Francisco, Calif. Mr. Galvin is a member of the American Association for the Advancement of Science.

James A. Galvin

Walter A. Knoop, Jr. (A'42–M'48) was born in Chicago, Ill., in 1919. He was graduated from Rensselaer Polytechnic Institute with the B.E.E. degree, after which he became associated first with the Detroit Edison Company, and then with Allen B. DuMont Laboratories, where he was made technical sales manager. He is now a manufacturer's representative with Gawler-Knoop, Inc., specializing in laboratory test equipment and vacuum tubes.

W. A. Knoop, Jr.

Mr. Knoop is a member of Sigma Xi, Tau Beta Pi, the American Institute of Electrical Engineers, and Instrument Society of America. He holds several FCC licenses, and is a registered professional engineer.

❖

Karl Zeigler Morgan was born near Kannapolis, N. C., in 1907. He received the A.B. degree in 1929 and the M.A. degree in 1930, from the University of North Carolina. He worked with the Westinghouse Electric Corporation in 1931, and received the Ph.D. degree in Physics from Duke University in 1934. From 1934 to 1943 he was head of the physics department of Lenoir Rhyne College, and during this period did research work in co-operation with Duke University in the field of cosmic radiation. He became a member of the Metallurgical Laboratory staff at the University of Chicago in the spring of 1943, and during the fall of 1943 he transferred to Oak Ridge National Laboratory, where he is now director of the Health Physics Division.

Karl Z. Morgan

Dr. Morgan is a member of the American Association for the Advancement of Science, the American Physical Society, the American Association of University Professors, and the American Industrial Hygiene Association.

Gordon R. Partridge (S'45) was born at Denver, Colo., on October 26, 1925. He was graduated from the Yale School of Engineering with the B.E. degree in 1946, and received the M. Eng. degree from Yale in 1947. He served in the U. S. Navy and was commissioned in the Supply Corps in 1946.

G. R. Partridge

During the summers of 1947 and 1948, Mr. Partridge was employed by the General Electric Company, working with power triodes. At present, he is an instructor in electrical engineering at Yale University, and a member of the Yale Graduate School.

He is a member of the AIEE, the Yale Engineering Association, Tau Beta Pi, Sigma Xi, and has been a holder of the General Electric Company Gerard Swope Fellowship.

❖

T. M. Shaw was born on November 9, 1910, at Philadelphia, Pa. He received the B.S. degree in physics in 1936 and the M.A. degree in 1940, both from the George Washington University. He has been employed at the Western Regional Research Laboratory since 1939.

T. M. Shaw

During a leave of absence which was granted in 1942–1943, he was employed at the Radiation Laboratory of the University of California, located at Berkeley, Calif.

Mr. Shaw is a member of the American Physical Society and of Sigma Xi.

Abstracts and References

Prepared by the National Physical Laboratory, Teddington, England, Published by Arrangement with the Department of Scientific and Industrial Research, England, and *Wireless Engineer*, London, England

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534.24 (26.03) 3293
Reflection of Underwater Sound from the Sea Surface—L. N. Liebermann. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 498–503; July, 1948.) Discussion of: (a) the theory of interference due to reflection from a plane sea surface, (b) pulse measurement technique, (c) the dependence of the reflection coefficient on the frequency used.

534.26 3294
The Diffraction of Sound by Circular Disks and Apertures—R. D. Spence. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 380–386; July, 1948.) A theoretical discussion, using oblate spherical co-ordinates. Numerical results for the cross section for scattering and transmission are shown graphically.

534.26 3295
The Diffraction of a Sound Wave by an Infinite Set of Plates—J. W. Miles. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 370–374; July, 1948.) The conditions for plane wave propagation within the plates and for a single reflected wave are given, together with expressions for the reflection and transmission coefficients. See also 2756 and 3504 of 1947 (Carlson and Heins).

534.26 3296
Notes on Sound Diffraction by Rigid Circular Cones—F. M. Wiener. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 367–369; July, 1948.) An experimental determination of the sound pressure at points on the surface of a circular cone exposed to a progressive wave. Results are shown graphically, and are generally similar to those obtained for other obstacles with the same circular face.

534.26:534.321.9 3297
Optical Study of Acoustic Fields near Diffracting Edges—J. C. Hubbard, J. F. Zartman, and C. R. Larkin. (*Jour. Opt. Soc. Amer.*, vol. 37, pp. 832–836; October, 1947.) Diffraction effects in ultrasonic sound fields in air were photographed by schlieren technique and by shadow methods using sparks of extremely short duration. Variations of field intensity were observed in the immediate vicinity of apertures and obstacles, in agreement with theoretical treatment more rigorous than that of Kirchhoff. Typical photographs are reproduced.

534.321.9:534.213.4:546.49 3298
On Ultrasonic Propagation through Mercury in Tubes—H. B. Huntington. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 424–432; July, 1948.) Measurements of the attenuation of 10.6-Mc pulsed ultrasonic waves in Hg as a function of tube diameter and the smoothness of the inner surface of the tube are discussed. Qualitative agreement with the Helmholz theory is obtained for tubes large compared to λ. Sonic propagation in a circular tube is considered theoretically in an appendix.

534.321.9:534.7 3299
Biological and Psychological Effects of Ultrasonics—H. Davis. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 589; July, 1948.) Summary only. Discussion of the principal physiological effects of intense sounds in various parts of the sonic and ultrasonic frequency spectrum.

534.321.9.001.8 3300
Ultrasonics Research and the Properties of Matter—C. Kittel. (*Rep. Progr. Phys.*, vol. 11, pp. 205–242; 1946 and 1947. Bibliography, pp. 242–247.) A study of the application of ultrasonic methods to the invesigation of gases, liquids, and solids.

534.322.3:534.7 3301
Physiology of Noise—A. Moles. (*Radio Franç.*, pp. 5–9; July and August, 1948.)

534.43:621.395.61 3302
A Single Shear Plate Crystal Phonograph Pick-Up—A. M. Wiggins and F. S. Lewis. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 448–450; July, 1948.) The design of phonograph pickups is discussed generally. A system for driving a piezoelectric crystal, using the mechanical transformer principle of coupling the stylus to a single slab crystal, is described as applied to a pickup; it can also be used for microphones, loudspeakers, galvanometers, etc. Advantages claimed are the elimination of even harmonics, very low effective mass, no pickup due to vertical stylus motion, and reduced record wear and needle scratch.

534.612 3303
Sound Pressure Measurement Equipment

for the Range 50 Cycles to 250 kc—F. Massa. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 451–454; July, 1948.) The microphone standard is only ⅛ inch in diameter. The stiffness of the vibrating system is adjusted to give a natural frequency of about 300 kc, so that pressure-wave transients having extremely steep wave fronts can be accurately reproduced. At free-field measurements are claimed to be free from diffraction errors; certain ultrasonic measurements can be made which were formerly very difficult. A preamplifier and power supply unit are included; the equipment is portable.

534.62 3304

Absorbing Media for Underwater Sound Measuring Tanks and Baffles—W. P. Mason and F. H. Hibbard. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 476–482; July, 1948.) Absorbent walls enable small tanks to be used for determining the directional properties of underwater sound instruments. Such walls have an inner lining of rubber, with fine-mesh screen or packed copper wadding in castor oil between the rubber and the outer shell. A 6-inch wall of this type reduces reflections by 20 db.

534.64 3305

Measurement of Acoustic Impedance—O. K. El-Mawardi. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 595; July, 1948.) Summary only. The impedance of material forming one boundary of a shallow cavity is found by measuring the sound pressure produced when a known volume current is injected from a high-impedance source. The volume current is found by observing the pressure when the cavity is terminated rigidly.

534.64 3306

Measurement of Coefficient of Reflection and Phase for Sound Waves—C. Kanta. (*Proc. Nat. Inst. Sci.* (India), vol. 6, pp. 671–693; December 16, 1940.) The reactance and resistance terms of the electrical impedance of a telephone receiver with an air load are determined separately, so that for a given surface both the reflection coefficient α and the phase change ϕ can be calculated. Results are given for various porous materials. No relation between α and ϕ is found. For assemblies of capillary tubes ϕ is $>\pi$ at a frequency of 1200 cps; for most materials it is $<\pi$. The experimental value for the impedance of the capillary tubes is in fair agreement with the value calculated from Rayleigh's formula.

534.64 3307

The Acoustic Feed-Back Method of Measuring Acoustic Impedance—A. London and C. R. Krishnamurthy. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 596; July, 1948.) Summary only. The acoustic impedance is used as a positive feedback network in a circuit containing a continuously variable filter; its frequency variation may thus be determined. Experimental results are given.

534.64 3308

Measurement of Acoustic Impedance by Short-Tube Techniques—P. Chrzanowski. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 595–596; July, 1948.) Summary only.

534.64 3309

Calculations on a Short-Tube Method for Measurement of Impedance—R. K. Cook. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 595; July, 1948.) Summary only. With a source of known volume velocity, the sound pressure within a short tube is dependent on the wall impedance. Impedance can thus be measured for both isotropic and anisotropic materials, as is shown mathematically.

534.64 3310

Acoustic Impedance Measurement of Very Porous Screen—C. M. Harris. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 440–447; July, 1948.) A new and simple method of measuring impedance at different frequencies, applicable even to screens whose impedance is less than 1/100 of that of air. The theory of the method is given, together with typical results for surgical gauze and perforated brass sheet.

534.7 3311

The Spiral of the Human Cochlea—N. R. French. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 591; July, 1948.) Summary only. The equation of the spiral is derived from Békésy's work (*Akus. Zeit.*, vol. 6, September, 1941; and *ibid.*, vol. 8, March, 1943) and the length of the basilar membrane is deduced. Impulse velocities along the cochlear duct are also calculated.

534.7 3312

The Interpretation of Loudness, Pitch, and Masking Phenomena with Regard to the Two-Canal Theory of Cochlea Mechanics—L. A. de Rosa. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 591; July, 1948.) Summary only.

534.76 3313

Stereophonic Sound—A. A. McK. (*Electronics*, vol. 21, pp. 88–89; August, 1948.) A magnetic tape carries 3 simultaneous sound channels. If the loudspeakers excited by these channels are properly arranged, the listener has the impression that the performers are actually present. So far the system has mainly been used for public auditoria but it can be adapted for home use.

534.771 3314

The Loudness of Repeated Short Tones—W. R. Garner. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 513–527; July, 1948.) The relation between loudness and repetition rate was determined experimentally. See also 3762 of 1947.

534.78 3315

Reaction of Small Enclosures on the Human Voice: Parts 1 and 2—C. T. Morrow. (*Jour. Acous. Soc. Amer.*, vol. 19, pp. 645–652; July, 1947; and vol. 20, pp. 487–497; July, 1948. Part 1: Specifications required for satisfactory intelligibility. Part 2: Analyses of vowels.

534.781 3316

Voice Measurements with an Audio Spectrometer—H. W. Rudmose, K. C. Clark, F. D. Carlson, J. C. Eisenstein, and R. A. Walker. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 503–512; July, 1948.)

534.79 3317

The Loudness and Monaural Loudness Matching of Short Tones—W. R. Garner. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 592; July, 1948.) Summary only.

534.79:534.851.86:621.396.665 3318

Loudness Control for Reproducing Systems—Bomberger. (*See* 3380.)

534.844.5 3319

The Time Integral Basic to Optimum Reverberation Time—J. P. Maxfield. (*Jour. Acous. Sci. Amer.*, vol. 20, pp. 483–486; July, 1948.)

534.844.5 3320

An Artificial Reverberation System—G. W. Curran. (*Audio Eng.*, vol. 32, pp. 13–17, 46; May, 1948.) Portions of the incoming signal are fed to two loudspeakers, the first coupled through 3 different lengths of pipe to microphones and the other coupled through a longer pipe to a fourth microphone. The microphone outputs are mixed and amplified. Additional delay is provided by feedback to the second loudspeaker through a differential mixer. Delays of 3 or 4 seconds can thus be obtained.

534.845 3321

Theory of the Absorption of Sound by Compressible Walls with a Non Porous Surface-Layer—C. W. Kosten and C. Zwikker. (*Physica 's Grav.*, vol. 8, pp. 251–272; February, 1941. In English.) For high absorption, the body of the material should have a low specific gravity and low elasticity. The pores should preferably be such that the air inside can, to some extent, vibrate independently. A material with large holes is preferable; it should be fixed some distance from a rigid wall.

534.845 3322

Absorption of Sound by Porous Materials: Parts 1–4—J. van den Eijk and C. Zwikker; C. Zwikker, J. van den Eijk, and C. W. Kosten; C. Zwikker. (*Physica, 's Grav.*, vol. 8, pp. 149–158, 469–476, 1094–1101, and 1102–1106; February, May, and December, 1941. In English, with German summary.)

Part 1. The absorption coefficient a of wood-fiber boards and acoustic plaster is deduced from measured values of porosity and air-flow resistance. The results are compared with direct measurements by an interference method, for normal incidence of the sound, and also by a reverberation method. The agreement between calculated and measured values of a is moderate for the higher frequencies, but for the lower frequencies, the calculated values are considerably below the measured values.

Part 2. The discrepancies noted above can be reduced appreciably by the introduction into the theoretical formulas of a structure factor depending on the directional distribution of the pores and on the number and size of internal cavities.

Part 3. Measurements of the acoustic impedance of sound-absorbing walls of simple structure, consisting of (a) straws, (b) glass tubes, arranged normal to or at an angle of 30° with the wall face, are in excellent agreement with calculated values when the structure factor is taken into account. Discussion shows that the dynamic air-flow resistance is considerably greater than that given by the Helmholtz-Kirchhoff theory.

Part 4. According to Kirchhoff's theory, the losses in absorbing materials are shared about evenly between the viscosity effect and the heat-conduction effect. It is here shown that the losses by conduction are very much less than by viscosity, the ratio being of the order of 1 to 60.

534.845 3323

Extended Theory of the Absorption of Sound by Compressible Wall-Coverings—C. W. Kosten and C. Zwikker. (*Physica, 's Grav.*, vol. 8, pp. 968–978; November, 1941. In English.) Solutions of the appropriate systems of differential equations are obtained for the case of (a) a compressible layer with normal (porous) surface, and (b) a compressible layer with a nonporous film coating. The results are compared with experiment (3324 below).

534.845 3324

Measurements of the Absorption of Sound by Porous Rubber Wall-Covering Layers: Parts 1–3—C. W. Kosten and C. Zwikker. (*Physica, 's Grav.*, vol. 8, pp. 933–967; November, 1941. In English.) Various sponge rubbers, with or without a nonporous surface layer, are found to have excellent absorption properties. Their behavior, in general, is in agreement with theory (3321 and 3323 above). The use of nonhomogeneous layers enables the selective absorption for particular frequencies to be controlled.

534.851:621.395.813:621.317.79 3325

New Circuit Design for Wow Tester—Nicholson. (*See* 3467.)

534.861.1 3326

Sound Measurements in BC [broadcasting] Studios—W. Jack. (*Tele-Tech*, vol. 7, pp. 38–41, 61; June, 1948.) Results of automatic frequency analysis of sustained sounds in various studios are discussed with particular reference to studio design.

621.395.61 3327

A Small High Frequency Microphone—I. Rudnick and H. C. Rothenberg. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 594; July, 1948.) Summary only. A Rochelle-salt crystal probe microphone and its associated two-stage preamplifier are described. The microphone response is relatively flat up to 50 kc in air and up to 100 kc in water.

621.395.61'.62 3328

Design of Laminated Magnetostrictive Longitudinal Oscillators—F. P. Bundy. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 594, July, 1948.) Summary only. The effects of various factors affecting frequency, sharpness of resonance, and efficiency are discussed and the design of various types of oscillator is considered.

621.395.61'.62 3329

Design and Construction of a 100-kc Magnetostrictive Transducer—I. Camp, R. Vincent and F. du Breuil. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 593, July, 1948.) Summary only. Discussion of design principles for a directional array. Performance data are included.

621.395.623.7 3330

Horn-Type Loudspeakers—S. J. White. (*Audio Eng.*, vol. 32, pp. 25–28, 34, May, 1948.) Design factors provide for (a) constant loading of the driver unit, and (b) minimum reflection at the horn mouth for the desired low-frequency cutoff. Reflex horns using conical inner sections are capable of reproducing up to 8000 cps. The characteristics of directional and radial horn loudspeakers are illustrated.

621.395.623.7.015.3 3331

The Transient Response of Loudspeakers—H. C. Hardy and H. H. Hall. (*Jour. Acous. Soc. Amer.*, vol. 20, p. 596; July, 1948.) Summary only. An interrupted varying af signal is applied to a loudspeaker in an anechoic room and the response of a transient-free microphone is displayed on a cro. For this system, the on transient differs from the off transient and cannot be correlated with it. Loudspeaker decay times are often greater than 0.01 second. See also 1313 of 1947 (Shorter).

621.395.625.3:621.395.813 3332

Factors Affecting Frequency Response and Distortion in Magnetic Recording—J. S. Boyers. (*Audio Eng.*, vol. 32, pp. 18–19, 47, May, 1948.) Discussion of methods of improving fidelity.

621.395.623.7 3333

Loudspeakers: The Why and How of Good Reproduction [Book Review]—G. A. Briggs. Wharfdale Wireless Works, Idle, Bradford, 1948, 87 pp., 5s. (*Electronics*, vol. 21, pp. 225–226, August, 1948.) ". . . presents the essentials for intelligent selection and evaluation of loudspeakers."

ANTENNAS AND TRANSMISSION LINES

621.392.029.64+621.396.611 3334

A Method for Calculating the Excitation of Waveguides, Surface and Cavity Resonators—Ya. N. Fel'd. (*Zh. Tekh. Fiz.*, vol. 17, pp. 1471–1482, December, 1947. In Russian.)

621.392.029.64 3335

On the Excitation of Waveguides: Part 2—A. A. Samarski and A. N. Tikhonov. (*Zh. Tekh. Fiz.*, vol. 17, pp. 1431–1440, December, 1947. In Russian.) The electromagnetic field in a waveguide excited by an arbitrarily oriented elementary current is determined. For this it is sufficient to determine the longitudinal components E_z and H_z since from these all other components can be derived. The equation (1) determining the two components and the boundary conditions (2) for the surface of the waveguide are quoted, methods for calculating the various components are deduced. The solutions so obtained can be used for calculating electromagnetic fields excited by linear, surface, and volume currents.

621.392.029.64 3336

A Rigorous Solution of the Problem of the Plane Waveguide with an Open End.—L. A. Weinstein. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 12, pp. 144–165; March and April, 1948. In Russian.) The problem is reduced to the solution of an integral equation. A detailed investigation of this equation and methods for solving it are given under the following headings: reflection coefficient of current; magnetic waves, electric waves, comparison with Kirchhoff's method (Huyghens' principle); phase of reflection coefficient; correction for the open end. The method can also be applied to the solution of analogous problems concerning circular waveguides, coaxial, and 2-wire transmission lines, etc. See also 3337 below.

621.392.029.64 3337

On the Theory of the Diffraction of Two Parallel Semi-Infinite Planes—L. A. Weinstein. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 12, pp. 166–180; March and April, 1948. In Russian.) Continuation of 3336 above. The limiting case when d/λ approaches infinity is investigated, where d is the distance between the two planes forming a plane waveguide with an open end and λ is the free-space wavelength.

621.392.029.64:621.396.67 3338

An Adjustable Waveguide Phase Changer—A. G. Fox. (*Bell Lab. Rec.*, vol. 26, pp. 245–250; June, 1948.) The phase changer consists of three short interconnected sections of cylindrical waveguide, each containing transverse conducting rods. Rotation of the center section advances or retards the phase of the transmitted wave, according to the direction and degree of rotation. An array of radar radiators, with phase-changers fitted in each feeder and driven from a common shaft by suitable gears, may be used to provide continuous scanning of the beam without rotation of the whole antenna system.

621.392.22 3339

Input Impedance of a Non Uniform Transmission Line—P. Gazzana-Priaroggia. (*Alta Frequenza*, vol. 17, pp. 99–109, June, 1948. In Italian, with English, French, and German summaries.) Discusses the hypotheses on which the approximate theory of such lines is based and gives a new method for calculating the input impedance of a line with slight irregularities. Results thus obtained are compared with those of other authors.

621.396.67 3340

Recent Developments on V.H.F. Ground-Communication Aerials for Short Distances—A. H. Brown and H. M. Stanier. (*Jour. IEE* (London), part III A, vol. 94, no. 14, pp. 637–643, 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 131.) The development of four types of antenna is outlined and certain of their characteristics are described. The antennas are: (a) simple dipole, (b) directional antenna embodying a corner reflector, (c) rhombic antenna, and (d) Yagi array, the respective frequency ranges are 100 to 156, 100 to 125, 65 to 85, and 65 to 95 Mc. The Yagi array requires minor adjustments to cover the required frequency range, the other three require no adjustment.

621.396.67 3341

The Performance of Some V.H.F. Aerials used in Naval Communications—A. G. D. Watson, G. Hanson, and J. H. Jones. (*Jour. IEE* (London), part IIIA, vol. 94, no. 11, pp. 666–669, 1947.) Description of a portable 3-element Yagi array for short distance point to point communication, a rotatable broadside array for shore to ship use, a single wire horizontal rhombic antenna for longer distance point-to-point working, and simple dipoles for all round radiation on board ship. Typical performance characteristics are discussed.

621.396.67 3342

Anti-Fading Series-Loaded Mast Radiators—H. Page. (*BBC Quart.*, vol. 2, pp. 165–176; October, 1947.) The radiation characteristics are derived for a mast consisting of two collinear sections separated by an insulator across which a variable reactance is connected. The results are applied to practical cases and the design data are summarized graphically. By varying the reactance, the mast can be used for a range of wavelengths, and the vertical polar diagram can be varied to suit a particular service area.

621.396.67 3343

Antennas for Circular Polarization—W. Sichak and S. Milazzo. (Proc. I.R.E., vol. 36, pp. 997–1001, August, 1948.) IRE 1947 National Convention paper. A formula is obtained for the variation in received voltage when an elliptically polarized antenna is rotated in a plane perpendicular to the direction of propagation of the incident elliptically polarized wave. A circularly polarized antenna will not receive any of the energy transmitted by it and reflected from a highly conducting smooth surface. The conditions for obtaining an omnidirectional circularly polarized pattern are derived. Experimental results are given.

621.396.67 3344

Slot Aerials: Part 2—Ya. N. Fel'd. (*Zh. Tekh. Fiz.*, vol. 17, pp. 1457–1470, December, 1947. In Russian.) Slot antennas in cavity resonators are examined for the case in which the excitation frequency differs little from the natural frequency of the resonator before the slots have been made. Formulas are derived for determining the voltage distribution along the slot and from this the fields inside and outside the antenna can be calculated. The results obtained are applied to the case of cylindrical resonators and some experimental curves are shown.

621.396.67 3345

Field Distributions near a Centre-Fed Half-Wave Radiating Slot—J. L. Putman, B. Russell, and W. Walkinshaw. (*Jour. IEE* (London), part III, vol. 95, pp. 282–289, July, 1948.) A description of experiments at a frequency of about 1000 Mc to map out the electromagnetic fields associated with a narrow $\lambda/2$ slot cut in a large metal sheet. The results are plotted to show the relative magnitudes and phases of the electric vector over a quadrant near the slot, the relative magnitudes of the magnetic vector along the slot axes, and the radiation polar diagram. The results are correlated with theory. See also 1335 of 1947 (Booker) and 3346 below.

621.396.67 3346

Input Impedances of Centre-Fed Slot Aerials near Half-Wave Resonance—J. L. Putman. (*Jour. IEE* (London), part III, vol. 95, pp. 290–294, July, 1948.) Results of measurements, at frequencies between 350 and 650 Mc, on slot radiators of various widths energized from parallel rod transmission lines. The impedances were calculated from the measured standing wave pattern. Two sets of experiments were made: (a) with the slots free to radiate on each side of the sheet, and (b) with resonant cavities to restrict the radiation to one side. The results are shown graphically and compared with those for the corresponding dipole radiator. See also 3345 above.

621.396.67 3347

Wide-Range Dual Band TV Antenna Design—L. L. Libby. (*Communications*, vol. 28, pp. 12–13, 31, June, 1948.) Description of an antenna designed to cover the 54 to 88 Mc and 174 to 216 Mc television bands and providing a substantially uniform directivity pattern and constant input impedance. It can be operated with either balanced or coaxial lines and can be grounded for lightning protection. Its use in directive arrays is also discussed.

621.396.67 3348
Radiation from Short Aerials—R. G. Medhurst. (*Wireless Eng.*, vol. 25, pp. 260–266; August, 1948.) A new trigonometrical approximation is developed and a theorem, based on this approximation and introducing a new concept called "the current center of gravity" of an antenna, is used to show that radiation patterns from linear radiators of length <λ/2 may be combined when the radiators are spaced in any way and carry currents of arbitrary phase differences. Examples of the use of the method are given.

621.396.67 3349
Note on Dielectric Aerials—I. Simon and O. Zinke. (*Onde Élec.*, vol. 28, pp. 278–281; July, 1948.) Simon derives formulas from which the radiation diagram of a dielectric rod antenna can be calculated. It is assumed that the antenna behaves as a purely longitudinal radiator, that it has no standing waves and that its diameter is negligible compared with its length. Diagrams are given for rods of length 2λ and 4λ. These have not been verified experimentally, but satisfactory agreement is found between the theoretical and experimental diagrams for a particular antenna investigated by Mallach (355 of 1947), though Mallach's velocities and critical diameters do not correspond with the theoretical formulas for TM_0 waves.

Zinke points out that Mallach's original work, described in *Ausgewählte Fragen über Theorie und Technik von Antennen*, vol. 2, pp. 132–169, deals much more fully with dielectric antennas than the simple review noted in 355 of 1947, and includes the formula given by Simon for the field intensity, with an added factor to take account of the thickness of the rod. The discrepancies between the theory for TM_0 waves and Mallach's experimental results are explained by the fact that Mallach used a different type of wave, which has been termed Type HE or Type TEM. Calculations for such waves are in good agreement with experiment. It is noted that Mallach's invention of dielectric rod antennas was prior to 1938.

621.396.67 3350
Dielectric-Rod Aerials—D. F. Halliday and D. G. Kiely. (*Jour. IEE* (London), part IIIA, vol. 94, no. 14, pp. 610–618; 1947. Summary *ibid.*, part IIIA, vol. 94, no. 11, p. 114; 1947.) The results of investigations carried out by Mallach and other German scientists and by the authors are combined in a comprehensive survey of the properties of uniform and tapered dielectric-rod radiators of circular cross section. Arrays of rods are studied, and dielectric tubes and horns are briefly treated. Approximate theoretical treatment is used to obtain radiation patterns, and the theoretical and experimental patterns are compared. The influence of such parameters as length and diameter of the rod, amount of taper, and permittivity of the dielectric is studied, and a summary is included of design data for optimum performance. A narrow beam may be obtained, but broad-band matching is difficult to achieve.

621.396.67 3351
Small Aerials in Dielectric Media—R. H. Barfield and R. E. Burgess. (*Wireless Eng.*, vol. 25, pp. 246–253; August, 1948.) Experiments, with theoretical discussion, to investigate the effect of surrounding open antennas and loops with dielectric materials, in order to improve the sensitivity of direction-finding systems. Results indicate that dielectrics exhibit pickup effects analogous to those of conductors. A single formula is derived for the current in an ellipsoidal receiving antenna, which applies generally to conducting or dielectric materials. The sensitivity of Adcock-type direction finders could probably be increased by surrounding them with low-loss dielectrics; but as the dimensions of the dielectric must greatly exceed those of the antenna system, the increase could be achieved more easily by enlarging the antenna system to make full use of the space which would be occupied by the dielectric.

621.396.671 3352
Some Methods for Determining the Power Gain of Microwave Aerials—J. D. Lawson. (*Jour. IEE* (London), part III, vol. 95, pp. 205–209; July, 1948.) Four methods are discussed and their limitations and advantages are indicated. Two require no knowledge of the antenna geometry or polar diagram. The other two require a knowledge of the field across the antenna aperture and can only readily be applied to horns of rectangular cross section, and even then approximations are necessary. A method for comparing antennas with a standard is also described.

621.396.671 3353
Parasitic Beam Patterns—D. C. Cleckner. (*CQ*, vol. 4, pp. 25–29; July, 1948.) Radiation patterns for 3- and 4-element 10-m antenna arrays are obtained from measurements on 1 to 100 scale models, using a wavelength of 10 cm. Front-to-back ratios of over 30 db are possible. Tuning for maximum forward gain will give a front-to-back ratio of about 20 db. This ratio is, in general, less for a 4-element than for a 3-element array, though the forward pattern is somewhat sharper. See also 2011 of 1947.

621.396.674 3354
Properties of Loop Aerials—F. Horner. (*Wireless Eng.*, vol. 25, pp. 254–259; August, 1948.) Nonuniform current distribution in a loop occurs unless the wavelength of excitation is much greater than the perimeter of the loop. Such distribution results in a nonuniform field around the loop; this field may be represented as that from a uniform-current loop, which is a true magnetic dipole, plus a component from an electric dipole, suitably located relative to the loop and suitably excited. Similarly, the field from a compound loop comprising two coplanar, nonuniform loops excited in parallel can be considered as due to that from a uniform loop plus components from two electric dipoles. The appropriate location and excitation of such effective dipoles are discussed for circular and rectangular loops of simple and compound types. The radiation resistance of such loops differs greatly from the value calculated on the assumption of uniform loop-current distribution.

621.392.029.64+621.396.61.029.64 3355
Microwave Transmission Design Data [Book Review]—Moreno. (*See* 3545.)

621.392.029.64 (083.72) 3356
British Standard 204:1943. Supplement No. 1, Glossary of Terms used in Waveguide Technique [Book Notice]—British Standards Institution, London, 1948, 2s. (*Brit. Stand. Inst. Mon. Inform. Sheet*, p. 1; May, 1948.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.012.2:621.392.5 3357
Circle Diagrams of Impedance or Admittance for Four-Terminal Networks—J. Rybner. (*Jour. IEE* (London), part III, vol. 95, pp. 243–252; July, 1948. Summary, *ibid.*, part I, vol. 95, p. 280; June, 1948.) A general theory is given which applies to both dissipative and asymmetrical networks, transmission-line diagrams being considered as special cases. The theory is based on the concept of the iterative impedances and the iterative transfer constant of the 4-terminal network, and leads to circle diagrams with the iterative impedances or admittances as double points. Such diagrams are actually graphical representations of the complex hyperbolic tangent function. For nondissipative networks, special forms of the diagrams correspond to the cutoff frequency, the stop band, or the pass band. These diagrams are studied by means of projective geometry. Inversion of the ordinary circle diagram leads to the Smith diagram, which is described both for real and complex iterative impedances. For a transmission line of length <λ/4, the error resulting from considering the characteristic impedances as purely resistive is of the same order of magnitude as that which would result from neglecting the damping along the line.

621.3.076.12:621.385.1:621.396.822 3358
Methods of Compensating the Various Actions of the Shot Effect in Valves and Connected Circuits—M. J. O. Strutt and A. van der Ziel. (*Physica 's Grav.*, vol. 8, pp. 1–22; January, 1941. In German, with English summary.) Spontaneous fluctuations in electrical circuits may, under certain specified assumptions, be considered as single-wave alternating currents and voltages. The different kinds of such fluctuations are discussed and methods for their suppression are proposed which all involve a special type of feedback. The compensating circuits are analyzed with special reference to the signal-to-noise ratio; in some circuits the signal-to-noise ratio may not be increased, though the fluctuations themselves are greatly reduced in some parts of the circuits. An exact analysis of one circuit is given in an appendix, showing the order of magnitude of the terms neglected in the previous approximations.

621.314.3† 3359
The Theory of Magnetic Amplifiers and Some Recent Developments—E. H. Frost Smith. (*Jour. Sci. Instr.*, vol. 25, pp. 268–272; August, 1948.) Defects of these amplifiers, such as nonlinear amplification and long time-constant, are explained. Performance may be materially improved by injecting into the core a constant biasing flux from permanent magnets.

621.314.5+621.396.622]:621.385.2 3360
The Diode as Converter and as Detector—J. Haantjes and B. D. H. Tellegen. (*Philips Res. Rep.*, vol. 2, pp. 401–419; December, 1947.) The current through a diode, to which a small extra voltage v is applied in addition to an ac and a dc voltage, can be developed in a series of powers of v with Fourier-series coefficients. The magnitude of these coefficients is calculated for a diode that has a linear characteristic in the pass direction. Using only the terms of the power series linear in v, quadripole equations and equivalent circuits can be set up for the diode as converter and as detector. From these equivalent circuits, various properties and quantities can be deduced. The fluctuations of the diode as converter can also be represented with the help of the equivalent circuit. For both conversion and detection, the diode should have a small internal resistance.

621.316.86:546.281.26 3361
Silicon-Carbide Non-Ohmic Resistors—F. Ashworth, W. Needham, and R. W. Sillars. (*Jour. IEE* (London), part I, vol. 95, pp. 244–246; June, 1948.) Discussion on 141 of 1947.

621.316.86:621.396.813 3362
Presence of Odd Harmonics in Alternating Current traversing Certain Nonmetallic Resistors—P. Sevin. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 183–185; July 19, 1948.) Measurements of the distortion of a sinusoidal 1000-cps current in passage through homogeneous resistor rods of various values showed the presence of an appreciable third harmonic, in some cases as high as 0.85 per cent. For $\frac{1}{4}$-w resistors having a steatite body with a thin conducting coating, distortion is difficult to measure and the third harmonic is in all cases <0.1 per cent. It thus appears that the resistors commonly used in radio technique may produce appreciable distortion. The dc resistance of these resistors diminishes when the applied voltage is increased; this diminution is greater for the homogeneous type of resistor than for the coated-steatite type.

621.385.1 3363
Applications and Mounting Details of Rimlock and Mazda-Medium Valves—Giniaux. (*See* 3555.)

621.392 3364
On some Properties of Electrical Networks—W. Nijenhuis and F. L. Stumpers. (*Physica 's Grav.*, vol. 8, pp. 289–307; February, 1941. In English, with German summary.) The relations between the real or imaginary part of an impedance function, or their logarithms, and the function itself, are discussed, with particular reference to transfer impedances. These relations enable a constant-modulus type of network and a constant-phase type to be distinguished. The design of such networks is considered and some general phase characteristics of driving-point and transfer functions are described.

621.392:518.61 3365
Network Analysis by the Chain-Relaxation Method—L. Tasny-Tschiassny. (*Jour. IEE* (London), part III, vol. 95, pp. 177–182; May, 1948. Summary, *ibid.*, part I, vol. 95, pp. 360–361; August 1948.) A certain number of branch currents are selected and considered as unknowns to which arbitrary values are assigned. All other branch currents, node potentials, generator voltages and currents, are expressed in terms of the selected branch currents. Additional "residual" external node currents are provided in order to make the resulting voltage and current distribution physically possible. The resulting simultaneous linear equations for the initially selected branch currents are solved. The special technique is based on the superposition theorem, obviates unnecessary repetition of algebraic symbols and reduces the possibility of errors. The method leads, in general, to simultaneous equations with a smaller number of unknowns than the usual methods of network analysis; in many cases, only two unknowns are required.

621.392:621.317.715 3366
The Moving Coil Galvanometer as a Circuit Element—N. F. Astbury. (*Proc. Phys. Soc.*, vol. 60, pp. 590–596; June 1, 1948. Discussion, pp. 596–597.) "An account is given of a simplified calculus applicable to a moving-coil galvanometer in which use is made of the concept of motional impedance. The fundamental galvanometer equations are presented in a novel and very simple form. Examples of the use of the formulas are given, notably with reference to the ultimate 'noise level' of a galvanometer."

621.392.5 3367
Theoretical Analysis of the Mercury Delay Line—H. J. McSkimin. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 418–424; July, 1948.) A formula for the pressure distribution in the line is derived. Many waves having slightly different phase velocities can exist; under certain circumstances these waves may interfere with each other and produce distortion. The voltage developed by the piezoelectric pickup crystal and the distortion that could be caused by carrier phase modulation are also considered.

621.396.611+621.392.029.64 3368
A Method for Calculating the Excitation of Waveguides, Surface, and Cavity Resonators—Ya.N.Fel'd. (*Zh. Tekh. Fiz.*, vol. 17, pp. 1471–1482; December, 1947. In Russian.)

621.396.611 3369
On the Frequency Stability of Certain Resonators—K. F. Niessen. (*Physica, 's Grav.*, vol. 8, pp. 1077–1093; December, 1941. In German.) The variations of the electromagnetic fundamental natural frequency of a hollow metal cube and of a hollow metal sphere are calculated (a) for constant length in one direction, with equal reductions in the other two directions, and (b) for an increase in one direction with equal decreases in the other two directions such that the total surface area remains constant. In case (a) the frequency variation for the cube is 8 times that for the sphere, but in case (b) the variation for the sphere is 3.6 times that for the cube. See also 3208 to 3210 of 1946.

621.396.611.1 3370
A New Approach to Tunable Resonant Circuits for the 300- to 3000-Mc Frequency Range—F. C. Isely. (PROC. I.R.E., vol. 36, pp. 1017–1022; August, 1948.) Discussion of the use of distributed constants and of discontinuities, in lines of fixed length. The advantages and disadvantages of various types of such circuits are discussed, with particular reference to those proposed by Karplus (3260 of 1945), Everett (2520 of 1946), and Huggins (330 of March and 643 and 644 of April).

621.396.611.1 3371
Rieke Diagrams for Oscillator Design—L. S. Schwartz. (*Tele-Tech*, vol. 7, pp. 33–36, 65; June, 1948.) If changes in power output and frequency of an oscillator, corresponding to load changes, are plotted on a Smith impedance diagram, contours of constant frequency and output can be drawn [see Radiation Laboratory Report 62-2, August 24, 1942 (Rieke and Evans)]. Such diagrams are useful for analysis of oscillator performance and assist materially in determining the optimum coupling. Apparatus and methods for obtaining the data necessary for the construction of these diagrams for microwave oscillators are described; examples are discussed.

621.396.611.3 3372
Cathode-Coupled Negative-Resistance Circuit—P. G. Sulzer. (PROC. I.R.E., vol. 36, pp. 1034–1039; August, 1948.) A complete investigation at low-frequency, medium-frequency, and high-frequency of a circuit discussed by Sziklai and Schroeder (3811 of 1945), Pullen (2507 of 1946), and Crosby (2157 of 1946). The effects of supply voltage variations are considered. The more common types of dual triode can develop a negative resistance of the order of 1000Ω and in this circuit they can be used as oscillators even at vhf. It is assumed that amplitudes are sufficiently small to permit the use of linear tube parameters; this may not be true unless some means of amplitude control is provided.

621.396.615 3373
Oscillator Circuits for Wide-Range Tuning—R. J. Ballantine and E. G. James. (*Jour. IEE* (London), part IIIA, vol. 94, no. 14, pp. 596–602; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 114; 1947.) Triodes of the disk-seal type are described briefly and a full account is given of various types of external circuit to cover frequency ranges between 300 and 3750 Mc. Coaxial-line common-grid circuits are described; a wide over-all frequency range may be covered by operation in different modes. Mechanical details of sliding-contact bridges are given and internal and external capacitive feedback is discussed. Performance data are included for the CV90 and CV273 tubes. Noncontact tuning circuits are also considered, including the butterfly circuit and its derivatives and a cylindrical resonator. Practical designs of symmetrical noncontact circuits are given, with performance data.

621.396.615 3374
High-Stability LC Oscillator—T. Roddam. (*Wireless World*, vol. 54, pp. 286–288; August, 1948.) The design of a bridge-stabilized oscillator is described, the performance of which does not depend on the nonlinearity of the tube characteristics. The selective feedback network uses a 4.5-v tungsten-filament lamp and a series-resonant *LC* circuit in opposite arms of a bridge network to maintain the oscillations at a chosen level. The short-period stability is within about 1 part in 10^6.

621.396.645 3375
Distributed Amplification—E. L. Ginzton, W. R. Hewlett, J. H. Jasberg, and J. D. Noe. (PROC. I.R.E., vol. 36, pp. 956–969; August, 1948.) IRE 1948 National Convention paper. By an appropriate distribution of ordinary tubes along artificial transmission lines, amplification can be obtained over much greater bandwidths than those possible with ordinary circuits. The concept of "maximum bandwidth-gain product" does not apply to such "distributed" amplifiers, for which the high-frequency limit appears to be determined by grid-loading effects. Band-pass amplifiers, and low-pass amplifiers with uniform frequency response between zero and several hundred megacycles per second, can thus be designed, using commercially available tubes; practical amplifiers have been built which have verified the theoretical predictions. Experimental work will be described later: theoretical discussion is here included of (a) the effect of improper termination of transmission lines, (b) methods for controlling the frequency response and phase characteristic, (c) the design which provides the required gain with the minimum number of tubes, (d) high-frequency limitations, (e) noise-factor evaluation.

621.396.645.029.3 3376
An Amplifier for Very Low Frequencies—S. P. Pivovarov. (*Zh. Tekh. Fiz.*, vol. 18, pp. 799–804; June, 1948. In Russian.) A 6-stage amplifier, with high input impedance, passing rectangular impulses at frequencies from 5 to 10,000 cps. The gain is 150,000.

621.396.645.371 3377
Improvements in Negative-Feedback Amplifiers—J. Polonsky. (*Ann. RadioÉlec.*, vol. 3, pp. 240–251; July, 1948.) A local reaction path is provided in one part of the amplifier and gives positive feedback within the pass band and negative outside the band. This increases the signal-to-noise ratio considerably (20 to 30 db) without affecting the amplifier stability. The gain is not appreciably altered, but linear and nonlinear distortion are both reduced. The theoretical conclusions are confirmed by tests carried out on a 1-kw and a 20-kw broadcasting transmitter. Other applications are suggested.

621.396.662.21 3378
Two Simple Methods of Detecting Short-Circuited Turns in Coils—E. B. Brown. (*Jour. Sci. Instr.*, vol. 25, p. 282; August, 1948.) In both methods, the coil to be tested is threaded on the core of a transformer with an open magnetic circuit. The secondary voltage of this transformer has previously been balanced against that of an exactly similar transformer. If the coil has short-circuited turns, a current flows in the detector. The lay-out of apparatus is shown in detail.

621.396.662.21 3379
Short-Circuited Turns—K. R. Sturley. (*Wireless Eng.*, vol. 25, pp. 240–245; August, 1948.) The common practice of short-circuiting turns to alter the inductance of a coil at rf is very unsuitable at audio and power frequencies. This is due not to any fundamental differences between rf and af short-circuits but to differences in coupling coefficients, *Q*-factors, and inductance ratios of the main coil to the short-circuited section.

621.396.665:534.79:534.851.86 3380
Loudness Control for Reproducing Systems—D. C. Bomberger. (*Audio Eng.*, vol. 32, pp. 11–12, 38; May, 1948.) A switch inserts up to 10 identical attenuator sections; these are designed on an image-impedance basis and have gain/frequency characteristics which compensate for the intensity/loudness relation of the human ear. A practical design is given, with component values and measured characteristics.

621.392+621.396.645.37 3381
Network Analysis and Feedback Amplifier Design [Book Review]—H. W. Bode. D. Van Nostrand, New York and Macmillan, London, 551 pp., 42s. (*Wireless Eng.*, vol. 25, p. 268; August, 1948.) "This book will undoubtedly become a standard textbook." It is for the specialist, not for the occasional designer. It is well packed with solid theory. See also 583 and 2169 of 1946.

GENERAL PHYSICS

53.081+621.3.081 3382
Universal Conversion Table of Electrical Units—Kaufmann. (*See* 3451.)

534.133 3383
Consideration of a Linear Non-Homogeneity of the Applied Alternating Field in the Excitation of a Quartz Bar—K. F. Niessen. (*Physica, 's Grav.*, vol. 8, pp. 695–702; July, 1941. In German.) The odd harmonics of a quartz bar can be produced by the application of a homogeneous alternating field of suitable frequency. The resonance curves in the neighborhood of these harmonic frequencies have been calculated by von Laue. The possibility of exciting the even harmonics by applying a nonhomogeneous alternating field is examined and resonance curves near these new frequencies are calculated. An electrode system suitable for excitation of the second harmonic is described briefly.

534.21 3384
Thickness-Shear Vibrations of Thin Anisotropic Plates—G. Hok. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 406–417; July, 1948.)

536.48 3385
Low Temperature and Some of Its Effects upon the Behavior of Matter—S. C. Collins. (*Science*, vol. 107, pp. 327–333; April 2, 1948.) Discussion of: (a) progress in low-temperature research, (b) industrial methods of producing low temperatures, (c) the effect of magnetic fields on superconducting substances, (d) various theories concerning the behavior of HeI, HeII, and the isotope He_3. See also 3847 of 1947 (Mendoza).

537.315.6:621.315.59 3386
Use of Electron Mirror to Display the Potential Distribution on Metallic and Semiconductor Surfaces—R. Orthuber. (*Z. Angew Phys.*, vol. 1, pp. 79–89; March, 1948.) Description of a method which has applications in infrared technique.

537.52 3387
Dynamic Probe Characteristics—T. A. Anderson. (*Phil Mag.*, vol 38, pp. 179–185; March, 1947.) Discussion of errors occasioned by using the static characteristics of probes in gas discharge investigations. Apparatus for obtaining the dynamic characteristics is described and preliminary results are given.

537.521.7 3388
Certain Aspects of the Mechanism of Spark Discharge—L. B. Loeb. (*Proc. Phys. Soc.*, vol. 60, pp 561–573, June 1, 1948.)

537.523:538.67:621.316.98 3389
Lightning Arrester Discharge Currents—R. C. Cuffe. (*Nature* (London), vol. 161, pp. 885–886; June 5, 1948.) The earth lead in each 38-kv single-phase arrester was equipped with a pair of magnetic links. Results indicate that these had not been magnetized by a unidirectional current discharge. The magnitude of the current reversal is evaluated and various possible explanations are given.

537.525.029.64 3390
Breakdown of a Gas at Microwave Frequencies—M. A. Herlin and S. C. Brown. (*Phys. Rev.*, vol. 74, pp. 291–296; August 1, 1948.) The criterion for breakdown of a low-pressure gas at microwave frequencies is that ionization by collision of electrons with neutral gas molecules replaces loss by diffusion to the walls of the discharge tube. A new ionization coefficient is introduced appropriate to the high-frequency discharge condition and its relation to the dc Townsend coefficient is explained. The energy transfer from field to electrons, for a given ratio of field strength to pressure, is most efficient when the pressure is high enough or the frequency low enough to result in many collisions, in each cycle, between electrons and gas molecules. When the pressure is lower or the frequency is higher, energy transfer is lower because the electrons have an out-of-phase component of motion.

537.533 3391
Thermionic Emission from Metals Covered by a Thin Semiconducting Layer—S. I. Pekar and O. F. Tomasevich. (*Zh. Tekh. Fiz.*, vol. 17, pp. 1393–1396, December, 1947. In Russian.) Formulas are derived for determining the thermionic current. They are consistent with Richardson's formula (16) in which use is made of the work function of the metal and not of the semiconductor.

537.533.8 3392
Secondary Emission: Parts 1 and 2—J. R. Koller. (*Gen. Elec. Rev.*, vol 51, pp 33–40 and 50–52; April and June, 1948.) Discussion of the various factors governing secondary emission, with a table giving the maximum secondary yield and corresponding voltage for many metals, and a short account of various applications.

537.58:621.385.1 3393
Space Charge Current Theory and the Mechanical Impulse of the Electrons—P. Selényi. (*Physica, 's Grav.*, vol 8, pp 885–902; August, 1941). In English, with German summary.) It is shown that the electrodes of a tube are subjected to pressures due to the impulse of the electrons. The value of the field strength at the surface of the cathode can be deduced from the mechanical equilibrium conditions, in accordance with the kinetic theory. If the electrodes of a tube are capable of movement, the inertia of the electrons can be demonstrated in many ways. A method of measurement of the pressure exerted by an electron beam is described. A preliminary report of this work was noted in 1804 of 1941.

538.12 3394
On Magnetic Field Theories—G. H. Livens. (*Phil Mag.*, vol 38, pp 453–479, July, 1947.) The field theory of Hertz, based on an assumed linear law of induction, is criticized. It is suggested that no theory has proved superior to the original Poisson-Kelvin theory which assumes nothing but a polar constitution for all magnetism. The work of Cohn and Guggenheim in developing the Hertz theory is discussed and a number of inconsistencies are pointed out. Arguments are given for the adoption of B as the fundamental field vector, and the consequent mathematical modifications of the Poisson Kelvin theory are indicated.

538.221 3395
Ferromagnetism—E. C. Stoner. (*Rep. Progr. Phys.*, vol. 11, pp. 43–109; 1946 and 1947. Bibliography, pp. 110–112.) The first part of this review summarizes the development of the basic ideas up to about 1934. The remainder is devoted to a survey of later theoretical and experimental work on the fundamental problem of intrinsic magnetization and its changes with field and temperature. A third part to be published next year will deal with magnetization curves.

538.551.21.029.63 3396
On the Significance of the Terms Current, Potential, Resistance, and Quadripole for Decimetre Waves—H. H. Meinke. (*Z. Angew. Phys.*, vol. 1, pp. 90–98; March, 1948.) A quantitative treatment of vhf devices is only considered possible if the terms current, potential, and resistance are critically analyzed and re-defined with the help of so-called induction-free and displacement-current-free surfaces. The simplest examples of such surfaces are the planes normal to homogeneous conductors. The general properties of these surfaces are discussed and the case of nonhomogeneous concentric conductors, with surfaces of greater complexity, is considered especially.

538.569.4.029.64:546.171.1 3397
Collision Broadening of the Inversion Spectrum of Ammonia: Part 3—The Collision Cross Sections for Self-Broadening and for Mixtures with Non-Polar Gases—B. Bleaney and R. P. Penrose. (*Proc. Phys. Soc.*, vol. 60, pp. 540–549; June 1, 1948.) For pure NH_3 the width variation of lines near 1 cm^{-1} is consistent with a simple dipole-dipole interaction for the collision mechanism. The collision diameters for mixtures of NH_3 with nonpolar gases are about the same as those obtained from the kinetic theory, whereas in pure NH_3 they are 2 to 4 times greater. Part 2: 1916 of August.

538.569.4.029.64:546.21 3398
Atmospheric Absorption of Microwaves—H. R. L. Lamont. (*Phys. Rev.*, vol. 74, p. 353; August 1, 1948.) The absorption of electromagnetic radiation was measured for λ between 6.34 mm and 4.48 mm, over distances ranging from 0.12 km to 2.2 km. The mean corrected values for the attenuation were found to be in good agreement with Van Vleck's theoretical curve (3098 of 1947).

549.211:537.533.9:539.16.08 3399
Remarks on Diamond Crystal Counters—K. Lonsdale. (*Phys. Rev.*, vol. 73, p. 1467; June 15, 1948.) "Type-I" diamonds which are opaque to ultraviolet radiation are said to be nearly perfect and neither of mosaic structure nor laminated. The "type-II" diamonds which make the best γ-ray counters, however, are apparently laminated. See also 2575 of October and 2757 of November.

549.514.51:535.417 3400
Application of Multiple-Beam Interferometry to the Study of Oscillating Quartz Crystals—S. Tolansky and W. Bardsley. (*Nature* (London), vol. 161, p. 925; June 12, 1948.) A method giving sharp fringes which become extremely narrow along nodal lines. Similar methods were used by Dye (*Rep. Nat. Phys. Lab.*, pp. 113–115; 1928).

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5:551.510.535 3401
The Theory of Meteor Ionization—N. Herlofson. (*Rep. Progr. Phys.*, vol. 11, pp. 444–453; 1946 and 1947. Bibliography, pp. 453–454.) Discussion of the salient points of the principal theories.

523.5:551.510.535 3402
Meteoric Ionization and Ionospheric Abnormalities—A. C. B. Lovell. (*Rep. Progr. Phys.*, vol. 11, pp. 415–442; 1946 and 1947. Bibliography, pp. 442–444. Part 1 summarizes the history of the correlation of irregularities in *E*-region behavior with meteoric activity. Part 2 gives a more detailed review of contemporary work on transient echoes from meteors. This includes the determination of the characteristics of the meteor showers themselves and also of the clouds of ionization from which echoes are received.

523.53:621.396.82 3403
Meteor Whistles—H. V. Griffiths, S. E. Martingell, and R. W. Bayliff. (*Nature* (London), vol. 161, pp. 478–479; March 27, 1948.) A summary of the results of measurements taken over a long period at the BBC receiving station at Tatsfield. Generally, the pitch of the whistle decreases to zero from about 2 kc, but for 1 to 2 per cent of the total

number observed, the pitch increases again up to 2 kc. The Appleton-Naismith theory for the change in pitch (3874 of 1947) would explain these "doublet" whistles.

523.7:538.12 3404

On the Sun's General Magnetic Field—T. G. Cowling. (*Mon. Not. R. Astr. Soc.*, vol. 105, no. 3, pp. 166–174; 1945.) Discussion of various theories, none of which seems entirely adequate; considerations are advanced which may make clearer the conditions which an adequate theory must satisfy.

523.746 3405

Magneto-Hydrodynamic Waves and Sunspots: Parts 1 and 2—H. Alfvén. (*Mon. Not. R. Astr. Soc.*, vol. 105, nos. 1 and 6, pp. 3–16 and 382–394; 1945.) A new theory of sunspots is advanced, according to which the magnetic field of the spot is the primary phenomenon. The other properties of a spot (e.g., the low temperature) are found to be due to the effect of the magnetic field on the solar atmosphere. The field is supposed to originate near the sun's center and to be transmitted outwards along the magnetic lines of force of the sun's general magnetic field in the form of a new type of wave, called the magneto-hydrodynamic wave. The progression of sunspot zones is discussed and a progression curve is derived from sunspot observations. Comparison with theory enables conclusions to be drawn regarding the sun's general magnetic field.

Part 2 deals with the shape and orientation of the magneto-hydrodynamic whirl rings. Such rings, created in the solar core, proceed outwards along the magnetic lines of force; when a whirl ring intersects the solar surface, a bipolar spot is produced. From the life history of a bipolar spot, the shape of the whirl ring when in the solar core can be found. The properties of whirls are considered and their propagation in the sun is discussed. The sun cannot be divided into a convective and a nonconvective part, as suggested in some solar models, since convection in one part of the sun must cause magneto-hydrodynamic waves which give rise to convection in other parts also. Part 3: 3104 of December.

538.12:521.15 3406

Comments on the Theories Interpreting the Magnetism of Celestial Bodies—M. Forró. (*Phys. Rev.*, vol. 74, pp. 218–219; July 15, 1948.) In the theory of Babcock (3891 of 1947) and Blackett (3112 of 1947), the consideration that with every mass an electric charge should be associated, must also involve the converse that with every moving (rotating) charge a momentum must be associated. This second inference can be shown to be untrue by a simple laboratory experiment. The difficulty in interpreting Wilson's relation, $e{\sim}M(G)^{1/2}$, is not so much in determining a field theory to account for the proportionality between mass and charge, but in explaining the irreversibility of this relationship. Barnóthy's theory (3407 above) will account for Wilson's formula and its irreversibility, but cannot explain the large static potentials arising and also give the correct values of magnetic momenta.

538.12:521.15:539.15 3407

Elementary Particles and the Geomagnetic Field—T. G. Cowling. (*Nature* (London), vol. 160, p. 847; December 13, 1947.) Short account of a theory proposed by J. Barnóthy (Hungarian Institute for Meteorology and Terrestrial Magnetism, Papers on Terrestrial Magnetism, no. 2, 1947), which explains the elementary particles, such as protons and electrons, in terms of serial universes, each enclosed in one of higher order. Quantitative application of the theory leads to values of fundamental constants in good agreement with experimental values. Barnóthy claims that the theory explains the proportionality between magnetic moment and angular momentum of sun and earth. He asserts that protons and neutrons possess, besides their real mass, an imaginary mass, 643 times the electron-mass, which behaves like an electric charge. The rotation of such a charge would produce terrestrial and solar magnetic fields of the observed order of magnitude. Such a charge would, however, produce a large electrostatic field, which Barnóthy declines to consider for the present.

550.384 3408

Audio-Frequency Magnetic Fluctuations—H. F. Willis. (*Nature* (London), vol. 161, pp. 887–888; June 5, 1948.) Magnetic fluctuations at frequencies in the range 5 to 1000 cps have been observed at various sites remote from man-made sources of electrical disturbance. These fluctuations are considered to be of natural origin, possibly arising from distant current movements in the atmosphere. See also 1628 of June (Menzel and Salisbury).

550.384.3 3409

The Main Geomagnetic Field—S. Chapman. (*Nature* (London), vol. 161, pp. 462–464; March 27, 1948.) A geophysical discussion held at the Royal Astronomical Society. A new theory suggested that the secular magnetic variation may be due to an electric current, produced by electromagnetic induction, flowing in the mass of an eddy near the surface of the earth's liquid core. Measurements of the magnetic anomalies in tholeiite dykes in Northern England, and the results of a survey over Eire of the vertical magnetic force (V), showing further anomalies, were also discussed. Some results of measurements of the variation with depth below ground of H and V were given, and discussed in relation to Blackett's theory noted in 3112 of 1947.

551.510.535:621.396.11 3410

On the Localization of the Sporadic-E Ionized Region of the Upper Atmosphere—P. Revirieux and P. Lejay. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 79–81; July 5, 1948.) In June, 1948, good 2-way low-power communication at frequencies near 60 Mc was maintained for several hours between amateur stations near Paris and others in (a) Czechoslovakia (June 27), (b) South Norway (June 28), (c) South Sweden (June 29). Such long-distance communication could not be established with any other stations during the periods concerned. On June 7 long-distance communication in the 60-Mc band could only be established with Algeria. These results indicate the existence of sharply localized ionization zones which move slowly and are situated roughly midway between Paris and the places with which radio contact on these frequencies is possible. Movement of such an ionization zone was noted on June 4 between 1630 and 2000 G.M.T., when communication was carried out successively between Paris and (a) northern countries, particularly Sweden, (b) Denmark, (c) Czechoslovakia, (d) Switzerland and Italy, (e) French stations near the Mediterranean, (f) Algeria.

Lejay states that particularly strong sporadic-E was recorded at Bagneux in two periods which coincide precisely with those noted above. The sporadic-E upper frequency limits, on the afternoon of June 4 and the morning of June 5, exceeded the highest frequency (13 Mc) available for the Bagneux transmitter. Between June 24 and 29, the upper limits were not so high, but many multiple echoes were noted, particularly on June 26 and 27.

The amateur observations supply valuable information as to the situation and extent of a phenomenon whose cause is at present uncertain; such information is not given by vertical soundings of the ionosphere. See also 3117 of December (Ferrell).

551.593.9 3411

Excitation Processes of the Night Sky Spectrum—S. N. Ghosh. (*Proc. Nat. Inst. Sci.* (India), vol. 9, pp. 301–310; December 29, 1943.) Discussion of various features of Mitra's hypothesis (1109 of 1944).

551.594.6:621.396.821 3412

Variation, with Wavelength of the Range of Atmospherics and of the Impulsive Flux per Metre Corresponding to the Threshold of Operation of Receiver-Recorders of the Mean Level—Carbenay. (*See* 3504.)

LOCATION AND AIDS TO NAVIGATION

621.396.93 3413

The Measurement of Errors in Radiogoniometers at High and Very High Frequencies—B. G. Pressey. (*Jour. IEE* (London), part III, vol. 95, pp. 221–228; July, 1948. Summary *ibid.*, part I, vol. 95, p. 279; June, 1948.) Description of three methods of error measurement which all involve the application of two signal-frequency voltages of known ratio r, one to each field coil of the goniometer. The goniometer reading for minimum search-coil output is compared with the true angle $\tan^{-1}r$. The accuracy of the methods is discussed and details of the construction and calibration of the apparatus are given. Typical error measurements made on various goniometers are shown.

621.396.93 3414

Radiogoniometers for High- and Very-High-Frequency Direction-Finding—B. G. Pressey. (*Jour. IEE* (London), part III, vol. 95, pp. 210–220; July, 1948. Summary, *ibid.*, part I, vol. 95, p. 278; June, 1948.) General design principles are discussed; the important factors are the coupling law, the electrical symmetry of the field coils, and the coupling factor. A distributed search-coil winding for reducing the coupling error at medium frequency and high frequency is described and its application to four high-frequency instruments is considered. At vhf a compound-wound search coil is used; its windings are in two sections whose planes are set at an angle θ. The coupling error depends upon θ and the field configuration. A vhf instrument based on these principles is described. The operation of the inductive goniometer as a phase shifter is analyzed.

621.396.932:621.396.96 3415

Radar Navigational Aid for Liverpool—(*Engineering* (London), vol. 165, p. 534; June 4, 1948.) Tests with experimental equipment showed that a discrimination better than 1° in bearing and about 40 yards in range was necessary to meet requirements for the long and narrow approach channel. The permanent equipment will include a 15-foot parabolic antenna for λ 3 cm, rotating at 12 rpm, and 6 plan-position indicators mounted in a U-shaped console. The first of these gives a small-scale picture of the whole channel up to a distance of either 13 or 20 miles; 4 give a larger-scale composite picture and the sixth enables any area within 20 miles, not shown on the composite picture, to be scanned. Information will be passed to pilots or masters of vessels by radio telephone. For other accounts see *Wireless World*, vol. 54, pp. 317–320; September, 1948, and *Engineer* (London), vol. 186, pp. 130–132; August 6, 1948.

621.396.933 3416

Surveillance Radar Deficiencies and How They Can Be Overcome—J. W. Leas. (Proc. I.R.E., vol. 36, pp. 1015–1017; August, 1948.) Surveillance or primary radar can be used by itself only as a monitor in civil operations. Its defects could be largely overcome, but only by major design alterations of existing equipment. Primary radar assisted by co-operating airborne transponders, however, can become a primary aid in traffic control.

621.396.96:621.396.65.029.64 3417

The Transmission of Radiolocation Displays by Means of Microwave Linkage Systems—Germany and Lawson. (*See* 3519).

621.396.933.2 3418
Radar Beacons [Book Review]—A. Roberts (Ed.). McGraw-Hill, New York, N. Y., 1947, 474 pp., $6.00. (Proc. I.R.E., vol. 36, p. 1010; August, 1948.) No. 3 of the MIT Radiation Laboratory series reviewing war-time work. "The theory and practice underlying the use of radar beacons is thoroughly covered." For another review see *Nature* (London), vol. 162, pp. 633–635; October 23, 1948.

621.396.96 3419
Principles of Radar [Book Review]—D. Taylor and C. H. Westcott. Cambridge University Press, London, 141 pp., 12s. 6d. (*Wireless Eng.*, vol. 25, p. 267; August, 1948.) One of the series on "Modern Radio Technique": others were noted in 350 of March (Huxley) and 1048 of May (Smith). General principles peculiar to radar are stressed; the book is mainly concerned with primary radar involving reflection from the target. The properties of the target are discussed with special reference to absorbing, scattering, and echoing areas.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.37 3420
The Luminescence of Simple Oxide Phosphors—R. B. Head. (*Electronic Eng.*, vol. 20, pp. 219–226; July, 1948.) Discussion of qualitative rather than quantitative experimental data. The phosphors examined are inefficient, and very insensitive to heat treatment and the addition of fluxes, but they provide data from which reliable conclusions can be drawn. Color diagrams are included. CaO is much the most important source of phosphors and CaO(Ce) gives the shortest afterglow.

535.37 3421
The Electron Trap Mechanism of Luminescence in Sulphide and Silicate Phosphors—G. F. J. Garlick and A. F. Gibson. (*Proc. Phys. Soc.*, vol. 60, pp. 574–590; June 1, 1948.)

535.37 3422
The Relation between Efficiency and Exciting Intensity for Zinc-Sulphide Phosphors—H. A. Klasens, W. Ramsden, and Chow Quantie. (*Jour. Opt. Soc. Amer.*, vol. 38, pp. 60–65; January, 1948.)

535.37 3423
The Decay of the Luminescence of Zinc Sulphide Phosphors Excited by X Rays—W. de Groot. (*Physica, 's Grav.*, vol. 8, pp. 789–795; July, 1941. In English, with German summary.)

535.37 3424
The Absorption Spectra of Zinc Sulphide and Willemite—J. H. Gisolf, W. de Groot, and F. A. Kröger. (*Physica, 's Grav.*, vol. 8, pp. 805–809; July, 1941. In English.) A survey of published data.

535.37:546.655.3 3425
Luminescence of Cerium Compounds—F. A. Kröger and J. Bakker. (*Physica, 's Grav.*, vol. 8, pp. 628–646; July, 1941. In English, with German summary.) Compounds of trivalent Ce, both in the solid state and in solution, have an absorption spectrum made up of several broad overlapping bands. Irradiation in this absorption region causes the emission of a similar set of broad double bands lying partly in the ultraviolet and partly in the short-wave portion of the visible spectrum. The doublet separation of about 1900 cm^{-1} is in satisfactory agreement with the value of 2253 cm^{-1} found for the free ion.

535.37:621.385.832:535.65 3426
Spectral Power Distriubtion of Cathode-Ray Phosphors—R. M. Bowie and A. E. Martin. (Proc. I.R.E., vol. 36, pp. 1023–1029; August, 1948.) Discussion of the principal existing colorimetric methods used for investigating screen materials. Methods for standardizing colorimetric measuring equipment are suggested.

538.213 3427
The Determining Factors of Permeability—J. L. Snoek. (*Physica, 's Grav.*, vol. 8, pp. 344–346; March, 1941. In English, with German summary.) The observed magnetic properties of cold-worked Ni-Fe alloys cannot be explained solely by internal strains and crystal anisotropy; at least one other factor, which is not as yet fully known, must be taken into account. See also 3428 below.

538.213 3428
Magnetic Anisotropy Phenomena in Cold-Rolled Nickel-Iron—G. W. Rathenau and J. L. Snoek. (*Physica, 's Grav.*, vol. 8, pp. 555–575; June, 1941. In English, with German summary. An exhaustive experimental investigation confirming the conclusions in 3427 above.

538.213:546.74 3429
Metastable Stages of Nickel Characterized by a High Initial Permeability—J. L. Snoek and J. F. Fast. (*Nature* (London), vol. 161, p. 887; June 5, 1948.) For pure, well annealed Ni, the initial permeability μ_0 is not a unique function of the temperature. The value at room temperature, when obtained from the results at decreasing temperatures, is nearly double that observed after demagnetizing. Temperature agitation alone does not bring about the more stable condition, but small mechanical shocks or demagnetizing treatment are also required.

538.221 3430
The Properties of Ferromagnetic Materials in Alternating Fields—K. M. Polivanov. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 12, pp. 98–115; March and April, 1948. In Russian.) A theoretical discussion of the relationship between the ratio $\dot{E}_0/\dot{H}_0$ on the surface of a sample in an alternating electric or magnetic field, and the following factors: frequency, complex permeability, specific conductivity, and one of the linear dimensions. Methods for calculating the performance of a sample are also indicated.

538.221:621.314.3† 3431
Improved Material for Magnetic Amplifiers—(*Electronics*, vol. 21, pp. 128, 166; August, 1948.) A co-ordinated summary of the papers presented at a symposium on magnetic materials held in Washington, D. C., June 15, 1948. Methods of producing materials having substantially rectangular hysteresis loops with a sharp knee are discussed, with special reference to the production of permenorm 5000-Z, an alloy consisting of equal proportions of Ni and Fe, by a process combining cold reduction and annealing. The permenorm is finally rolled to a thickness of 0.0012 or 0.002 inch and wound into spiral cores. These cores should not have airgaps or be mechanically strained, and the magnetic field must be sensibly uniform over the entire cross section. Applications of these cores are discussed, namely: (a) magnetic amplifiers for industrial control equipment needing little maintenance, and (b) saturable reactors for improving the timing of commutation in efficient high-current mechanical rectifiers.

538.23 3432
The Influence of Eddy Currents on the Apparent Hysteresis Loop of Ferromagnetic Bars—J. L. Snoek. (*Physica, 's Grav.*, vol. 8, pp. 426–438; April, 1941. In English, with German summary.) If the magnetizing current is varied so rapidly that the magnetization of the interior of a bar lags behind that of the exterior, the measured mean induction differs considerably from that found after a relatively slow variation of the field. With ring-shaped samples this effect does not occur. With bars, the measured value of the coercive force may be in error by a factor of 2 or more and the error in the total hysteresis loss may be still larger. The effect explains various inconsistencies and anomalies in measurement results for bars and wires.

549.514.51 3433
The Laboratory Growing of Quartz—D. R. Hale. (*Science*, vol. 107, pp. 393–394; April 16, 1948.) A report of experimental work carried out by the Brush Development Company.

549.623.5:621.315.616 3434
Magnetic Susceptibility of Mica—J. T. Kendall and D. Yeo. (*Nature* (London), vol. 161, pp. 476–477; March 27, 1948.) The high paramagnetism of synthetic mica prepared by the method of Eitel and Dietzel is due to inclusions of magnetite.

620.193.2 3435
Electrical Contacts: The Effect of Atmospheric Corrosion—U. R. Evans. (*Metal Ind.* (London), vol. 73, pp. 10–13; July 2, 1948.) A discussion of present knowledge and the results of experiments conducted on "tarnish films." Two entirely different types of corrosion are possible: (a) oxide films produced when the air is not contaminated, and (b) corrosion at high humidities in air contaminated by sulphur in various forms. The need for further research is stressed and the conditions which should be observed in practice are indicated. A bibliography of 31 items is included.

621.3.015.5.029.63/.64:546.217 3436
Experiments on the Electric Strength of Air at Centimetre Wavelengths—R. Cooper. (*Jour. IEE* (London), part III, vol. 95, p. 312; July, 1948.) Discussion on 750 of April.

621.315.5:666.11 3437
Electrically Conducting Glasses—R. I. Green and K. B. Blodgett. (*Jour. Amer. Ceram. Soc.*, vol. 31, pp. 89–100; April 1, 1948.) Glasses containing oxides of Pb, Bi, or Sb, or combinations of these, become conducting after several hours' reduction in H. The electronic surface conductivity is both stable and reproducible. The magnitude of the conductivity depends on the nature and amounts of the reducible oxides, the temperature at which reduction takes place, and the electrical influence of the unreduced portions of the glasses.

621.315.59 3438
Modulation of Conductance of Thin Films of Semi-Conductors by Surface Charges—W. Shockley and G. L. Pearson. (*Phys. Rev.*, vol. 74, pp. 232–233; July 15, 1948.)

621.315.59:546.289 3439
Nature of the Forward Current in Germanium Point Contacts—W. H. Brattain and J. Bardeen. (*Phys. Rev.*, vol. 74, pp. 231–232; July 15, 1948.)

621.315.612 3440
Ceramic Dielectric Materials—B. H. Marks. (*Electronics*, vol. 21, pp. 116–120; August, 1948.)

621.315.612.011.5 3441
Properties of Barium-Magnesium Titanate Dielectrics—G. R. Shelton, A. S. Creamer, and E. N. Bunting. (*Jour. Amer. Ceram. Soc.*, vol. 31, pp. 205–212; July 1, 1948.) Continuation of 3551 of 1947. Results of measurements of dielectric constant and power factor reciprocal are tabulated for frequencies of 50 to 20,000 cps together with some measurements at 3000 Mc. The effects of composition, various heat treatments, aging, and temperatures in the range −60°C to +85°C are discussed and tabulated.

621.315.616:534.213:534.321.9 3442
Measurement of Ultrasonic Bulk-Wave Propagation in High Polymers—A. W. Nolle and S. C. Mowry. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 432–439; July, 1948.) The velocity and attenuation of longitudinal waves of dilatation in solid samples of high polymers are measured at frequencies between 10 and 30 Mc by an acoustic pulse technique. Results for various materials and temperatures are discussed.

621.315.616:679.5 3443
Electrical Properties of Plastics—A. J. Warner (*ASTM Bull.*, pp. 60–64, August, 1948. Discussion, p. 64.) Results are given for the loss factor of polytetrafluorethylene (a) at 24°C from 60 cps to 3000 Mc, (b) at 60 cps, for heat cycles between +30°C and −160°C. The effect of various plasticizers on the high-frequency power factor of polyethylene and polytetrafluorethylene at high temperatures is also discussed.

621.385.032.21:[546.86+546.36 3444
Optical and Photoelectrical Properties of Antimony-Caesium Cathodes—N. D. Morgulis, P. G. Borzyak, and B. I. Dyatlovitskaya. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 12, pp. 126–143; March and April, 1948. In Russian.)

666.115 3445
A Note on Very Soft Glasses and Some of Their Electrical Applications—A. E. Dale and J. E. Stanworth. (*Jour. Soc. Glass Tech.*, vol. 32, pp. 147–153; June, 1948.) Preliminary results of the experimental production of "solder glasses" which bear the same relationship to normal glasses as solder bears to normal metals. (Pb, Zn)-borate glasses appear to be the most promising.

669.71:620.193.2 3446
Protective Films—F. A. Champion. (*Metal Ind.* (London,) vol. 72, pp. 440–442, 444 and 463–464; May 28 and June 4, 1948.) The film of alumina formed on Al and its alloys under corrosive conditions is responsible for the high corrosion resistance of these metals and controls the corrosion versus time curve. Results of research on these curves are discussed.

538.221 3447
New Developments in Ferromagnetic Materials [Book Review]—J. L. Snoek. Elsevier Publishing Co., New York and Amsterdam. British distributors: Cleaver-Hume Press, London, 1947, 136 pp., 13s 6d. (*Metal Ind.*, (London), vol. 72, p. 465, June 4, 1948.) Based on work carried out at the Natuurkundig Laboratorium, Eindhoven, Holland, since 1933 and during the last war. In part 2, the dynamics of ferromagnetism are discussed; a new approach is made by the introduction of a time factor, whose influence in various processes is considered. See also *Nature* (London), vol. 161, p. 666; May 1, 1948.)

MATHEMATICS

518.5 3448
An Ultrasonic Memory Unit for the EDSAC—M. V. Wilkes and W. Renwick. (*Electronic Eng.*, vol. 20, pp. 208–213; July, 1948.) An introductory description of the general principles and mechanical construction of the memory unit of the EDSAC, now being built at the University Mathematical Laboratory, Cambridge, pointing out its advantages over the ENIAC developed in America. Theoretical circuit and block schematic diagrams and photographs are included. A similar type of memory is used in the EDVAC (2828 of 1947)

518.61:621.392 3449
Network Analysis by the Chain-Relaxation Method—Tasny-Tschiassny. (*See* 3365.)

517.942.82 (083.5) 3450
Tabellen zur Laplace-Transformation und Anleitung zum Gebrauch. (Tables for the Laplace Transformation and Introduction to its Use.) [Book Review]—G. Doetsch. Springer, Berlin, and Göttingen, 1947, 185 pp. (*Phys. Blätter*, vol. 4, no. 3, p. 129; 1948.) The book is limited to the classical one-dimensional Laplace transformation. A further work on two- and multi-dimensional transformations, which are important for dealing with systems of partial differential equations, is announced in the preface.

MEASUREMENTS AND TEST GEAR

53.081+621.3.081 3451
Universal Conversion Table of Electrical Units—A. Kaufmann. (*Radio Franç*, pp. 9–11; July and August, 1948.) A useful table relating the cgs electrostatic and electromotive systems to the practical, Giorgi, and rationalized Giorgi systems of units.

621.3.018.4 (083.74) 3452
Emission of Standard Frequencies—(*Phys. Blätter*, vol. 4, no. 3, p. 127; 1948) From May 1, 1948 standard frequencies of 440 and 1000 cps, of quartz-clock accuracy, will be radiated between 10.30 and 10.40 each Monday by every transmitter of the North-West Germany broadcasting station. The following transmission frequencies will be used: 904 kc (331.9 m), 1330 kc (225.6 m), and 6115 kc (49.08 m).

621.317.332 3453
Absolute Measurement of the Time Constant of Resistors—J. W. L. Köhler and C. G. Koops. (*Philips Res. Rep.*, vol. 2, pp. 454–467; December, 1947.) A new method in which standard capacitors with negligible losses are used as reference standards.

621.317.361.029.64+621.317.763.029.64 3454
Contribution to the Study of Methods and Apparatus for Measurements in the Centimetre Wave Band: Part 2—M. Denis and R. Liot. (*Ann. RadioÉlec.*, vol. 3, pp. 189–213; July, 1948.) A detailed account of methods for the measurement of frequencies and frequency variations, with particular reference to wavemeters of the cavity-resonator type and their best operating conditions. The determination of the modulation characteristics of cm-λ generators is discussed; this illustrates the qualities which precision wavemeters should possess and also the limits of their usefulness. Part 1: 1076 of May.

621.317.44 3455
The Development of a Magnetic Testing Apparatus to Determine Iron Loss at High Flux Densities—W. Cormack. (*Trans. S. Afr. Inst. Elec. Engrs.*, vol. 38, part 10, pp. 257–299; October, 1947.) Discussion, pp. 299–300.) Limitations of routine test methods at flux densities approaching 20,000 lines/cm² are discussed. The main problem is found to be that of producing a voltage whose harmonic content can be varied so as to produce a sinusoidal flux wave in the test specimen under all conditions. The design of suitable equipment is considered. Using a cro as a phase indicator, values of B_{max} can be measured to an accuracy within 1 per cent and losses within 5 per cent.

621.317.715 3456
Moving-Coil Galvanometers of Short Period and their Amplification—A. V. Hill. (*Jour. Sci. Instr.*, vol. 25, p. 284; August, 1948.) Discussion of performance records of one of the galvanometers considered in 3174 of December (Downing) and 3175 of December (Hill).

621.317.715 3457
Investigation and Improvement of the Vibration Galvanometer—G. von Mayrhauser. (*Z. Angew. Phys.*, vol. 1, pp. 68–75; March, 1948.) The use of "koerzit" steel for the magnetic needle of a galvanometer of the frequency-independent type developed by Meissner and Adelsberger (1930 abstracts, *Wireless Eng.*, p. 405) gives a considerable increase of sensitivity and broadens the frequency-independent zone. A sharp minimum in the current versus sensitivity curve due to mechanical coupling is investigated theoretically; calculations are in good agreement with observations.

621.317.723:621.385.5 3458
H. F. Pentodes in Electrometer Circuits—Crawford. (*See* 3557.)

621.317.729 3459
An Electrolytic Tank for Exploring Potential Field Distributions—R. Makar, A. R. Boothroyd, and E. C. Cherry. (*Nature* (London), vol. 161, pp. 845–846; May 29, 1948.) When using ac, polarization and corrosion of the electrodes cause difficulties which are greatly reduced when the operating frequency is raised. Accurate results are obtained at 1000 cps with electrodes 0.5 mm in diameter, provided the electrode current density is not greater than 0.3 ma/mm². Copper-plated steel sewing needles make suitable electrodes.

621.317.73 3460
Radio-Frequency Resistance Meter—H. W. Kline. (*Gen. Elec. Rev.*, vol. 51, pp. 25–30; June, 1948.) An instrument for determining rf resistance from the scale readings of a meter rather than as the ratio of reactance to resistance. A circuit diagram, photographs, and an abac for quick calculations are given. Applications are discussed.

621.317.73 3461
New Measuring Circuit for Conductance Meter—W. A. McCool. (*Tele-Tech*, vol. 7, pp. 30–31, 48; June, 1948.) Full circuit details of an instrument for measurement at 1 Mc of rf losses in high-quality insulating materials. A diode conductance circuit is used, with a sensitive differential voltmeter which enables the resonance voltage to be maintained accurately constant during a measurement.

621.317.733 3462
A Schering Bridge for Testing Insulating Materials—R. J. Stanley. (*Muirhead Technique*, vol. 2, pp. 20–23; July, 1948.) Details of the Type D-98-A bridge and Type D-99-A Wagner earth. Optimum performance is at frequencies of the order of 1 kc but good results can be obtained from 200 cps to 10 kc. Possible errors are within 5 per cent for both power factor and dielectric constant.

621.317.761 3463
Comparison of Neighbouring Frequencies by Counting Beats—R. Sewig. (*Z. Angew. Phys.*, vol. 1, pp. 49–50; March, 1948.) Description of a "phase filter," with which two frequencies in the range 50 cps to 100 kc can be accurately compared. Provision is made for recording the beat frequency and indicating the direction of drift of one frequency no matter how often the beat frequency becomes zero.

621.317.763 3464
An Absorption Wavemeter for the Decimetre Region—D. G. Reid and J. K. Garlick. (*Jour. IEE* (London), part IIIA, vol. 94, no. 14, pp. 603–604; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 106; 1947.) A coaxial-line instrument, with two fixed conductors and a sliding conductor of intermediate diameter. Resonance is indicated by a microammeter operated from a crystal rectifier. Construction details are given for a model for the frequency range 470 to 500 Mc, reading to ±0.5 Mc. Silver plating increases the Q-factor considerably.

621.317.763 3465
A New Wavemeter for Centimetre Waves—N. N. Malov. (*Zh. Tekh. Fiz.*, vol. 18, pp. 793–798; June, 1948. In Russian.) A high-accuracy narrow-band wavemeter consisting of a waveguide of variable cross section and filled with two different dielectrics.

621.317.784.088 3466
Some Sources of Error in Microwave Milliwattmeters—G. F. Gainsborough. (*Jour. IEE* (London), part III, vol. 95, pp. 229–238; July, 1948.) An approximate analysis of the factors that influence the intrinsic accuracy of a heated filament used for measuring power. Instruments incorporating such filaments are usually subject to significant errors if the filaments are longer than about λ/10. The errors are usually smaller in resistance milliwattmeters than in thermocouple instruments used

similarly. Additional serious errors can be caused by the inefficiency of the transformer used to match the power source to the filament of the instrument. Filament instruments cannot, in their present state of development, be relied on as standard transfer instruments for measuring microwave power in terms of low-frequency calibration.

621.317.79:534.851:621.395.813 3467
New Circuit Design for Wow Tester—M. G. Nicholson. (*Tele-Tech*, vol. 7, pp. 26–29, 63; June, 1948.) Equipment for the measurement of frequency variations of 0.5 to 40 cps, and also amplitude variations up to 25 per cent at any variation rate from 0.5 to 40 cps. Such amplitude variations may be caused in magnetic wire reproducers if the wire is not continuously in contact with the reproducing head. Recording milliammeters and cro attachments enable graphs and oscillograms of the quantities in question to be obtained.

621.317.79:621.392.029.64 3468
The Design of Precision Standing-Wave Indicators for Measurements in Waveguides—D. Hirst and R. W. Hogg. (*Jour. IEE* (London), part IIIA, vol. 94, no. 14, pp. 589–595; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 114; 1947.) Slotted-section indicators for H_{10} waves in rectangular waveguides are considered. Requirements for precision measurement are summarized, and methods of obtaining the necessary mechanical accuracy are discussed. Mechanical rigidity and first-class workmanship are essential. The design of a detector-matching system using a coaxial line is described, and the various available types of detector are compared. Methods are given for checking the accuracy of construction and for calibrating the detector. A detailed illustrated description is given of typical equipment for the 10-cm and 3-cm wavelength bands.

621.317.79:621.396.615:621.397.62.001.4 3469
Picture-Modulated Television Signal Generator—A. Easton. (*Electronics*, vol. 21, pp. 110–115; August, 1948.) For production testing of receivers at points remote from television transmitting stations. Circuit and performance details are given. A mixing pad permits the combination of picture, sound, and noise signals.

621.317.79:621.396.621.54.001.4 3470
The Testing of Communication-Type Radio Receivers—W. J. Bray and W. R. H. Lowry. (*Jour. IEE* (London), part III, vol. 95, pp. 271–276; July, 1948.) Discussion on 2298 of September.

621.317.79:621.396.645 3471
A Practical Gain Set—C. G. McProud. (*Audio Eng.*, vol. 32, pp. 20–23; May, 1948.) Design and construction details of a simple instrument.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.321.9.001.8 3472
Ultrasonics Research and the Properties of Matter—Kittel. (*See* 3300.)

535.61-15 3473
Developments in the Infra-Red Region of the Spectrum—G. B. B. M. Sutherland and E. Lee. (*Rep. Progr. Phys.*, vol. 11, pp. 144–175; 1946 and 1947. Bibliography, pp. 175–177.) Developments in detectors are first reviewed, the frequency-selective types (photographic, phosphor, photoemissive, photoconductive, and photovoltaic) and nonselective types (thermoelectric, bolometric, pneumatic, radiometric, evaporographic, and convective) being considered separately. Improvements in filters and in dispersive and window materials are described. Recent applications and possible future developments, particularly spectroscopic, are discussed.

539.16.08 3474
Studies on Helium-Filled Geiger Müller Counters—H. R. Sarna, P. L. Kapur, and Charanjit. (*Proc. Nat. Inst. Sci.* (India), vol. 8, pp. 277–287; August 18, 1942.) Experiments with tubes filled with a mixture of He and alcohol vapor showed that methyl and ethyl alcohols are the most suitable. Optimum results are obtained for alcohol partial pressures of 2.5 to 3.0 cm Hg.

539.16.08:537.533.9:549.211 3475
Remarks on Diamond Crystal Counters—Lonsdale. (*See* 3399.)

550.837.7:621.3.091:553.57 3476
The Attenuation of Ultra-High-Frequency Electromagnetic Radiation by Rocks—R. I. B. Cooper. (*Proc. Phys. Soc.*, vol. 61, pp. 40–47; July 1, 1948.) The attenuation of 200-Mc signals in dry sandstone was found to be 3 to 4 db/foot.

621.318.572:518.5 3477
Megacycle Stepping Counter—C. B. Leslie. PROC. I.R.E., vol. 36, pp. 1030–1034; August, 1948.) Development and general construction are described, with special reference to the input-pulse commutator and the method of interstage coupling. Application to electronic digital computers is also considered.

621.38.001.8:522.6 3478
Electronics in Astronomy—G. E. Kron. (*Electronics*, vol. 21, pp. 98–103; August, 1948.) Discussion of accurate time-keeping devices, automatic aids to the precise tracking of telescopes, and photoelectric photometers.

621.38.001.8:771.36.001.4 3479
Testing Photographic Shutters—S. H. Duffield and L. R. Lankes. (*Electronics*, vol. 21, pp. 82–87; August, 1948.) A review of the basic circuits used in electronic testers.

621.384.6 3480
Some High-Frequency Measurements on a Cyclotron Model—G. Lindström. (*Ark. Mat. Astr. Fys.*, vol. 35, part 2, section A, 10 pp; June 18, 1948. In English.) An account of measurements of dee-voltage variations for a 1 to 10 scale model of the complete high-frequency system of the cyclotron under construction at the Nobel Institute of Physics, Stockholm; the frequency was correspondingly increased.

621.384.6 3481
A Note on the Shape of the Polepieces of a Synchrotron Magnet—N. Davy. (*Proc. Phys. Soc.*, vol. 60, pp. 598–599; June 1, 1948.) The shape of polepieces giving the required field variation is deduced by treating the problem as two-dimensional and using conformal transformations.

621.384.6 3482
A 16-MeV Betatron—K. J. R. Wilkinson, J. L. Tuck, and R. S. Rettie. (*Nature* (London), vol. 161, pp. 472–473; March 27, 1948.) The betatron weighs 0.85 ton, operates at 50 cps, and requires 4 kw input power. The limited orbit space requires high uniformity in the field around the orbit.

621.384.6:621.319.339 3483
Electrostatic Generators for the Acceleration of Charged Particles—R. J. Van de Graaff, J. G. Trump, and W. W. Buechner. (*Rep. Progr. Phys.*, vol. 11, pp. 1–17; 1946 and 1947. Bibliography, pp. 17–18.) Discussion of the endless-belt type of generator and the methods of applying the potentials to accelerating tubes. Particular attention is paid to the design of the belt system, insulation problems, and the construction of multiple-electrode tubes. Voltage measurement and stabilization are considered briefly.

621.385.833 3484
Electron Optics—L. de Broglie. (*Bull. Soc. Franç. Élec.*, vol. 8, pp. 292–300; June, 1948.) Theory of the formation of images in ordinary optical and in electron instruments is reviewed; the resolving power of electron microscopes is discussed in relation to the theories of Heisenberg.

621.386.1 3485
On a New Type of Rotating-Anode X-Ray Tube—A. Taylor. (*Proc. Phys. Soc.*, vol. 61, pp. 86–94; July 1, 1948.) Description of the "Peristron," a water-cooled tube without rotary vacuum seals.

629.13.053.2 3486
Sonic True Air Speed and Mach Number Indicator—V. B. Corey. (*Jour. Acous. Soc. Amer.*, vol. 20, pp. 583–584; July, 1948.) Summary only. A practical instrument in which the extension of a servo-controlled movable beam, carrying an acoustic transmitter, indicates the true air-speed. The basic principles of the instrument could be used to measure the speed of a body moving in any fluid of low viscosity. See also 2883 of November (Hoather).

PROPAGATION OF WAVES

538.566 3487
The Application of a Variational Method to the Calculation of Radio Wave Propagation Curves for an Arbitrary Refractive Index Profile in the Atmosphere—G. G. Macfarlane. (*Proc. Phys. Soc.*, vol. 61, pp. 48–59; July 1, 1948.) An analysis is given of the propagation of radio waves through an atmosphere in which the variation of the modified refractive index with height can be represented by a linear term together with one or more exponential terms. Two examples are given, corresponding to a surface duct and an elevated duct respectively, and a practical height-gain curve measured for λ 3 m is shown to be in good agreement with that derived theoretically from the refractive-index gradient. See also 2892 of 1947 (Booker and Walkinshaw).

621.396.11+538.566 3488
The Velocity of Propagation of Electromagnetic Waves derived from the Resonant Frequencies of a Cylindrical Cavity Resonator—L. Essen and A. C. Gordon-Smith. (*Proc. Roy. Soc. A*, vol. 194, pp. 348–361; September 2, 1948.) Full account of the work noted in 3249 of 1947. Final measurements gave the velocity as 299,792 km/second. The estimated maximum error in this result is 9 km/second.

621.396.11 3489
Propagation of a Direct Wave Around the Earth Taking into Account Diffraction and Refraction—V. A. Fock. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 12, pp. 81–97; March and April, 1948. In Russian.) Assuming uniformity of the surface of the earth, the propagation of radio waves is determined by diffraction around the curved surface of the earth, refraction in the lower layers of the atmosphere and reflection from the ionosphere. For ranges over 1000 km, the third factor predominates, but it is still possible, under certain conditions, to separate and observe the direct (diffraction) wave. Its study is of great importance for distance ranging by interference methods.

A theory of the direct wave is given and a solution of the Maxwell equations for the Hertzian vector is found in which diffraction and refraction are both taken into account. The conception of an equivalent of the earth is examined and conditions under which its use is justified are established. It is shown that this conception can be used for regions of shadow and semishadow, where the methods of geometrical optics are not applicable.

621.396.11 3490
Radio Shadow Effects Produced in the Atmosphere by [temperature] Inversions—W. L. Price. (*Proc. Phys. Soc.*, vol. 61, pp. 59–78; July 1, 1948.) The meteorological conditions

described by Booker (516 of 1947) enable radar sets to detect objects at very large ranges. The possibility that these long ranges are associated with "shadow" zones in which an aircraft would not be detected is here investigated theoretically by ray tracing methods. Controlled experiments, in which measurements of all the relevant physical conditions were made, show that these shadow zones do exist above certain temperature inversions, and that their position and extent agree well with theory. See also 2892 of 1947 (Booker and Walkinshaw).

621.396.11 3491
The Elements of Wave Propagation using the Impedance Concept—II—G. Booker. (*Jour. IEE* (London), part III, vol. 95, pp. 203–204; May, 1948.) Discussion on 3250 of 1947.

621.396.11:535.3 3492
Intensity-Distance Law of Radiation—H. R. L. Lamont and J. A. Saxton. (*Wireless Eng.*, vol. 25, p. 269; August, 1948.) A value $d^2/8\lambda$ is suggested for the critical distance parameter r_0 dividing the optical case from the radio case; this is much smaller than Bell's value $d^2/0.52\lambda$ (2592 of October). The value $d^2/8\lambda$ is based on practical results obtained for λ 3 cm and λ 9 cm; the inverse square law is found to hold at ranges $>4r_0$ when the path difference between rays arriving from the edge and center of the aperture is not greater than $\lambda/4$, if the phase of the radiation is uniform over the aperture.

621.396.11:551.510.535 3493
Equivalent Path and Absorption for Oblique Incidence on a Curved Ionospheric Region—J. C. Jaeger. (*Proc. Phys. Soc.*, vol. 61, pp. 78–86; July 1, 1948.) A Chapman distribution of ionization density is assumed. The absorption, the equivalent path, and the projection of the path on the earth's surface may all be expressed by equivalence theorems of the plane-earth type with suitably modified parameters and functions. Numerical values of the functions involved are given. See also 3252 of 1947.

621.396.11:551.510.535 3494
On the Localization of the Sporadic-E Ionized Region of the Upper Atmosphere—Revirieux and Lejay. (*See* 3410.)

621.396.11.029.58 3495
Investigations of High-Frequency Echoes—H. A. Hess. (Proc. I.R.E. vol. 36, pp. 981–992; August, 1948.) See also 3214 to 3216 of December and 3496 below.

621.396.11.029.58 3496
Short-Wave Echoes—H. A. Hess. (*Funk und Ton*, vol. 2, pp. 334–344; July, 1948.) Continuation of 2053 of August. The results of observations for frequencies of 10 to 20 Mc during 1941 to 1945 are summarized and discussed. The time t for a complete circuit of the earth varied between 0.13760 and 0.13805 second, the mean value being 0.13778 second. No dependence of t on either frequency, time of day, or season of year was found, but the frequency range for echo signals increased toward the period of sunspot maximum, for which it was 15 to 30 Mc. Comparison of the amplitudes of direct and reverse signals, and single- and multi-circuit signals, showed on the average a field-strength decreasing linearly with distance. Multipath effects were noted for the main signal of high-power transmitters when the transmission path was <1000 km; the time delays of the interfering signals may be several milliseconds. Split signals and the Doppler effect are discussed and a possible explanation is suggested for the fading-out of the reverse signal which is mainly observed during summer nights. The results, in general, favor a tangential propagation path above the E layer. See also 3214 to 3216 of December and 3495 above.

621.396.11.029.62:621.396.932 3497
V.H.F. Cross-Channel Communication—N. Levin, A. G. D. Watson, G. Hanson, G. W. Parks, and D. A. Cobb. (*Jour. IEE* (London), part IIIA, vol. 94, no. 14, pp. 663–665; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 131; 1947.) Propagation conditions between the Isle of Wight and Normandy at frequencies between 85 and 95 Mc had to be determined for designing reliable communication equipment for D day. Transmitter height would be about 700 feet, receiver height 50 feet, the maximum range required was 83 sea miles and the optical range was about 36 sea miles. Crystal-controlled transmitters and receivers were to be used, the transmitted power being 100 w and the receiver sensitivity 6.0μv input for 20 db signal-to-noise ratio. Propagation tests were undertaken over paths off the Cornish and Welsh coasts. The results are discussed; it was predicted that a 20-db signal-to-noise ratio could be expected over the operational path for 60 per cent of the time. The operational network which actually provided almost continuous service is described.

621.396.11 3498
Elementary Manual of Radio Propagation [Book Review]—D. H. Menzel. Prentice-Hall, New York, N. Y., 1948, 220 pp., $7.65. (Proc. I.R.E., vol. 36, p. 1009; August, 1948.) Graphical methods of calculating field strength are included for various conditions of distance, time, frequency, geographical location, and topography. The treatment of the frequency range 3 to 30 Mc is particularly comprehensive.

RECEPTION

621.396.621 3499
On the Selectivity of Broadcast Receivers—W. Kleinsteuber and G. Riese. (*Funk und Ton*, vol. 2, pp. 327–333; July, 1948.) Consideration of the possibility of cutting out the reception of a powerful station, particularly a local station. Curves are given showing the frequency limits within which this is practicable for single and 2-stage circuits. The selectivity of two simple circuits coupled by a tube is compared with that of a band-pass filter.

621.396.621:621.396.619.11 3500
The Synchrodyne—"Cathode Ray." (*Wireless World*, vol. 54, pp. 277–281; August, 1948.) A simple explanation of its basic principles. See also 1139 of May and back references.

621.396.621.54 3501
The Theory of the Super-Regenerative Receiver Operated in the Linear Mode—G. G. Macfarlane and J. R. Whitehead. (*Jour. IEE* (London), part III, vol. 95, pp. 143–157; May, 1948. Summary, *ibid.*, part I, vol. 95, p. 233; May, 1948.) Formulas are derived for the properties of the receiver in two states of operation corresponding roughly to sinusoidal and rectangular wave quench. It is predicted and confirmed by experiment that the frequency response and the envelope of the output oscillations of such a receiver will each have the shape of a Gaussian error curve. A theory is given according to which the whole of the noise energy collected by the rf acceptance band of the receiver goes to produce noise in a bandwidth equal to half the quench frequency in the receiver output. The effect of further decreasing the post-detector bandwidth is also considered.

621.396.621.54 3502
Tracking in the Superheterodyne—K. Pfeil. (*Funk und Ton*, vol. 2, pp. 358–370; July, 1948.) Design formulas are derived for the values of L and C for the various circuits of a superheterodyne, with numerical examples and also curves and tables of correction terms which greatly simplify calculation.

621.396.662:621.396.62 3503
Tuning Devices for Broadcast Receivers—R. C. G. Williams. (*Jour. IEE* (London), part III, vol. 95, pp. 240–241; July, 1948.) Discussion on 1194 of 1947.

621.396.821:551.594.6 3504
Variation with Wavelength, of the Range of Atmospherics and of the Impulsive Flux per Metre corresponding to the Threshold of Operation of Receiver-Recorders of the Mean Level—F. Carbenay. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 51–52; July 5, 1948.) Measurements with the recorders at the Laboratoire National de Radioélectricité, for λ 2 to 25 km, indicate that the impulsive flux varies approximately as the square of the range, which increases from about 1000 km for λ 2 km to over 3000 km for λ 25 km. See also 2902 of December.

621.396.822:621.3.015.33:621.396.619.13 3505
On the Calculation of Impulse-Noise Transients in Frequency-Modulation Receivers—F. L. H. M. Stumpers. (*Philips Res. Rep.*, vol. 2, pp. 468–474; December, 1947.) The effect of such transients is calculated by means of a series expansion of the phase, the general term of which contains $[A(t)]^n$ when the amplitude $A(t)$ of the disturbance is smaller than that of the signal, and $[A(t)]^{-n}$ in the contrary case. The Laplace transform is used to calculate the effect in the filters. The large effect of phase opposition during the capture time is explained.

STATIONS AND COMMUNICATIONS SYSTEMS

621.396.41 3506
A Centimetre-Wavelength Beam Telephone—D. G. Reid and J. K. Garlick. (*Jour. IEE* (London), part IIIA, vol. 94, no. 14, pp. 605–609; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 114; 1947.) The principles of operation and some construction details of lightweight 3-cm equipment for optical paths up to about 10 miles. It can be used for bridging a gap in a telephone line. The transmitter tube also acts as the local oscillator for the receiver, making send/receive switching unnecessary. The antenna system is highly directional; this reduces the risk of interference or jamming. The input power of 100 w is obtained either from ac mains or from a 12-v battery.

621.396.41:621.396.619.16 3507
Methods and Equipment used in Pulse Multiplex Communication Systems—G. Potier. (*Funk und Ton*, vol. 2, pp. 273–284, 345–357, and 396–406; June to August, 1948.) Translation into German of paper abstracted in 4031 of 1947.

621.396.41:621.396.65 3508
Multiplex Radiotelephony Link between the Mainland and Corsica—P. Rivère. (*Ann. Radioélec.*, vol. 3, pp. 221–239; July, 1948.) The characteristics which a radio-telephone link should possess are reviewed and the particular advantages of FM carrier-current systems are enumerated. Theoretical calculations of the performance of the link between Grasse and Calenzana (Corsica) are compared with the results obtained in practice. Favorable refraction conditions, combined with a lower noise figure than that expected on theoretical grounds, assure uninterrupted communication in spite of the fact that 55 km of the total distance of 205 km is beyond the optical range. A detailed description of the 12-channel equipment is given. The transmission frequency toward Corsica is 97 Mc and toward the mainland 107 Mc. Transmitter power at the base of the feeder is 100 w.

621.396.41.029.62:621.396.611.21 3509
Reference-Crystal-Controlled V.H.F. Equipment—D. M. Heller and L. C. Stenning. (*Jour. IEE* (London), part III, vol. 95, pp. 157–160; May, 1948.) Discussion on 2616 of October.

621.396.5 3510
Modern Single-Sideband Equipment of the Netherlands Postals Telephone and Telegraph—C. T. F. van der Wyck. (Proc. I.R.E., vol. 36, pp. 970–980; August, 1948.) For another account see 1158 of May.

621.396.61/.62 3511
Citizens Band Transceivers: Part 4—W. B. Lurie. (*Electronics*, vol. 21, pp. 76–81; August, 1948.) Details of the modifications required to enable an IFF [identification, friend or foe] transponder, type BC-645, to provide 2-way communication at 465 Mc for mobile and fixed stations. Field tests carried out over various types of terrain are discussed. For earlier parts see 2456 of October (Rowland) and back references.

621.396.619.13 3512
Frequency Modulation—K. R. Sturley. (*Jour. IEE* (London), part III, vol. 95, p. 142; May, 1948.) Discussion on 4047 of 1945.

621.396.619.13:621.396.41 3513
The Application of Frequency Modulation to V.H.F. Multi-Channel Radiotelephony—J. H. H. Merriman and R. W. White. (*Jour. IEE* (London), part IIIA, vol. 94, no. 14, pp. 649–658; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 131; 1947. Discussion, *ibid.*, part III, vol. 95, pp. 277–281; July, 1948.) The use of FM radio links to form part of a trunk telephone network is discussed. Up to twelve channels are provided, each of 4 kc bandwidth, in the range 60 to 108 kc; the carrier frequency is between 45 and 80 Mc, with ±300 kc deviation. In general, on optical paths the radiated power is about 10 w, whereas 100 w would be required for AM. The factors determining signal-to-noise ratio, distortion, and interchannel crosstalk are discussed and results of trials carried out over optical and long nonoptical paths are given.

621.396.65 3514
Ultra-High-Frequency Techniques applied to Mobile and Fixed Communication Services—J. Thomson, J. D. Denly, I. J. Richmond, F. Pugliese, and H. Borg. (*Jour IEE* (London), part IIIA, vol. 94, no. 14, pp. 630–632; 1947.) Discussion on 2081 of August.

621.396.65 3515
Résumé of V.H.F. Point-to-Point Communication—F. Hollinghurst and C. W. Sowton. (*Jour. IEE* (London), part IIIA, vol. 94, no. 14, pp. 669–672; 1947.) Discussion on 2080 of August.

621.396.65 3516
G.E.S-T [studio-transmitter] Link Equipment—D. J. Nigg. (*FM and Telev.*, vol. 8, pp. 32–35, 47; June, 1948.) Detailed description of Type BL-2-A FM broadcast equipment, designed specifically for relaying high-fidelity programs, and consisting of transmitter, receiver, two identical high-gain antennas, and an external rack-mounted pre-emphasis unit. The transmitter output is 10 w and the carrier frequency range 920 to 960 Mc. See also 3250 of December (DeWitt).

621.396.65:622.86 3517
Applicability of Radio to Emergency Mine Communications—E. W. Felegy and E. J. Coggeshall. (*United States Bureau of Mines*, Report R.I. 4294, 56 pp; May, 1948.) An account of experiments carried out in several different mines, using frequencies in the range 33 to 220 kc. Methods tried included (a) communication through the ground, (b) carrier-current communication over the power distribution system, and (c) inductive communication making use of telephone, signal, and power wires, or track lines, without direct connection. The results are summarized and discussed. See also 2367 of September.

621.396.65.029.62 3518
Choice of Frequency for V.H.F. Radio Links—D. A. Bell. (*Jour. IEE* (London), part IIIA, vol. 94, no. 14, pp. 633–636; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 131; 1947.) The main factors affecting the choice are: (a) rate of diffraction attenuation beyond the horizon, (b) acceptable amount of fading, (c) directional gain of antennas (d) effective height of receiving antennas, and (e) freedom from mutual interference with distant channels using the same frequency.

621.396.65.029.64:621.396.96 3519
The Transmission of Radiolocation Displays by Means of Microwave Linkage Systems—L. W. Germany and D. I. Lawson. (*Jour. IEE* (London), part IIIA, vol. 94, no. 14, pp. 619–629; 1947.) An account of the development of two systems by which information can be transmitted by cm-λ links to remote PPI and height range displays. The first, which was operated over a distance of 16 miles is fully described, with block diagrams. Signal-to-noise ratio was about 8 to 1. The second was not completed, but an account of the principles of the system is given. Four relay stations were intended to convey information over a total distance of about 120 miles.

621.396.676:621.396.93 3520
Common-Aerial Working for V.H.F. Communication—A. G. D. Watson, J. H. Jones, and D. L. Owen. (*Jour. IEE* (London), part IIIA, vol. 94, no. 14, pp. 644–648; 1947. Summary, *ibid.*, part IIIA, vol. 94, no. 11, p. 131; 1947.) To economize in the number of antennas required on board ship, a number of transmitters or receivers tuned to different frequencies use the same antenna. Each set is connected to the antenna through a filter which acts as a short-circuit at unwanted frequencies. To prevent the main feeder from being completely short-circuited, connectors approximately λ/4 long are inserted between each filter and the branch point. A brief theoretical treatment is given.

621.396.712 3521
Large [British s.w.] Broadcasting Station—(*Elec. Rev.* (London), vol. 142, pp. 993–997; June 25, 1948!) Short illustrated account of the BBC overseas station at Skelton, Cumberland. The two sets of buildings are about 1½ miles apart, at opposite ends of an oval site covering 750 acres. The 51 antenna arrays are supported by 31 masts of heights from 200 to 350 feet. The equipment of one section, OSE8, was provided by the English Electric group of companies and includes six 100-kw Marconi transmitters. The equipment of OSE9 is similar to that of OSE8, but the transmitters were made by Standard Telephones and Cables. Transmissions are beamed to different parts of the world by the aid of 6 antenna-switching towers, which are remotely controlled. A general description is given of a transmitter unit and its associated power supplies. At slightly reduced power, the combined maximum demand of the two sections is 5400 kw, with a load factor of about 75 per cent. With all transmitters operating, 1500 kw is delivered to the antenna arrays.

621.396.712.2 3522
The Trend of Design of Broadcasting Control Rooms—F. Williams. (*BBC Quart.*, vol. 2 pp. 184–192; October, 1947.) A description of the pre-war system and the design for future BBC installations. The continuity-suite system described by Wynn (4033 of 1947) was used extensively during the war and will be developed to provide the operator with up to 200 program sources with 5-channel mixing. The present trend is for decentralization, leaving the control room proper to handle only miscellaneous programs.

621.396.72:621.3.029.62 3523
Hifam. E.H.F. [88-Mc/s] Amplitude-Modulated Broadcasting in U.S.A.—S. Tarzian. (*Wireless World*, vol. 54, pp. 297–298; August, 1948.) An account of results with an experimental AM transmitter. With an antenna height of 800 feet, power gain of 10, output of 200 w and using vertical polarization, the service area extended to a radius of 25 miles. Simple frequency converters can be used with standard broadcast-band receivers. Frequency stability of the local oscillator was obtained by using invar coils and capacitors with zero temperature-coefficient. Receivers are simpler and cheaper than for FM and a narrower band of frequencies is used.

621.396.932:621.396.11.029.62 3524
V.H.F. Cross-Channel Communication—Levin, Watson, Hanson, Parks, and Cobb. (See 3497.)

621.396.97 3525
The War-Time Activities of the Engineering Division of the BBC—H. Bishop. (*Jour. IEE* (London), part III, vol. 95, pp. 301–302; July, 1948) Discussion on 2086 of August.

621.396.619.13 3526
FM Transmission and Reception [Book Review]—J. F. Rider and S. D. Uslan. J. F. Rider, New York, N. Y., 1948, 409 pp., $1.80. (*Electronics*, vol. 21, pp. 222, 224; August, 1948.) Most of the commercial broadcast types of FM transmitter are included, as well as some used in amateur and police radio work. Receivers, reactance tubes, limiters, discriminators, locked oscillators and ratio detectors are described in detail. Antenna systems and servicing are also considered.

SUBSIDIARY APPARATUS

621.314.12:621.394/.396].66 3527
Amplifying Dynamos. The Amplidyne—A. Valentin. (*Bull. Soc. Franç. Élec.*, vol. 8, pp. 304–328; June, 1948.) A comprehensive discussion of basic principles, different types of machine, and practical applications.

621.396.68 3528
Vibrator Power Packs—D. A. Bell. (*Wireless World*, vol. 54, pp. 272–276; August, 1948.) The principles of design are considered, with reference to (a) the choice of the correct value for the "timing" capacitance connected across the transformer secondary, (b) the operating conditions of the transformer iron and copper with square-wave currents, (c) the voltage regulation, and (d) the suppression of radio interference.

621.396.682:621.316.722.1 3529
A D.C. Stabilized Power Supply of Low Impedance—V. H. Attree (*Jour Sci. Instr.*, vol. 25, pp. 263–268; August, 1948.) The unit gives an output of 400 v for a load current of 0 to 90 ma, with a voltage variation of 200 mv. It consists of a full-wave rectifier with a single-stage II filter followed by a series power tube, the grid of which is driven from a single-tube dc amplifier coupled to the 400-v output. Design and operation details are given. The output impedance is <2Ω at any frequency up to 10 kc.

TELEVISION AND PHOTOTELEGRAPHY

621.397.3:621.385.832 3530
Flying Spot Designed for Television Studio Scanning—(See 3563.)

621.397.6:621.385.832 3531
New Tubes for Color Television—H. Gilloux. (*Radio Franç.*, pp. 22–24; July and August, 1948.) An account of the construction and principles of the chromoscope. See also 2937 of November (Bronwell).

621.397.61 3532
TV Transmitter Design: Parts 2 and 3—G. E. Hamilton. (*Communications*, vol. 28, pp. 20–22, 29, and 10–13; June and July, 1948.) Discussion of modulated amplifiers, adjustment of video signal amplitude and of Class-B linear amplifiers. To be continued. Part 1: 2948 of November.

621.397.62+621.397.82 3533
The TV Receiver: Its Operation and Common Forms of Interference—W. Brown. (*CQ*, vol. 4, pp. 31–35, 91; July, 1948.) Discussion of the general circuit arrangements of television

receivers, and particularly of the diagnosis of interference by studying the distortion of the screen picture.

621.397.62.001.4:621.317.79:621.396.615 3534
Picture-Modulated Television Signal Generator—(*See* 3469.)

621.397.621 3535
Frame Deflector-Coil Efficiency—W. T. Cocking. (*Wireless World*, vol. 54, pp. 289–292; August, 1948.) Some of the problems involved in the design of *RC*-fed deflector coils are discussed. Although the frame scan is mainly dependent on the coil resistance, both resistance *R* and inductance *L* are of importance during the fly-back period. The power supplied to the coil depends on I^2L and R/L. There is a direct relation between the input power from the tube and the coil power, provided the ratio of the coupling resistance R_a to the coil resistance has its optimum value. In practice, the fly-back requirements set a minimum value to R_a and the optimum relation cannot always be realized. The relative merits of pentode and triode circuits for feeding the coil are discussed.

621.397.743 3536
Land-Line Technique for Television Outside Broadcasts—H. B. Rantzen, A.R.A. Rendall, and S. N. Watson. (*BBC Quart.*, vol. 2, pp. 177–183; October, 1947.) By using transportable repeater stations at intervals of the order of 1 mile, ordinary telephone lines may be used to link up with exchanges in Central London where special "television cable" is available. For comparatively long distances (>8 miles) this cable is usually coaxial, frequency translation being necessary to eliminate the effect of low-frequency noise. For short distances, twin cable is used, without frequency translation, although this makes equalization more difficult.

621.397.828 3537
You *can* Live with Television!—H. M. Bach, Jr. (*CQ*, vol. 4, pp. 33–37, 87 and 36–38, 90; June and July, 1948.) General discussion of television interference due to amateur transmissions, and of methods of eliminating such interference. See also 3538 below.

621.397.828 3538
TVI [television interference] Corrective Measures—W. M. Scherer. (*CQ*, vol. 4, pp. 34–37, 90; August, 1948.) Practical methods for application to amateur transmitters. See also 3537 above.

TRANSMISSION

621.3.09 3539
Reducing Transmission Bandwidth—R. S. Bailey and H. E. Singleton. (*Electronics*, vol. 21, pp. 107–109; August, 1948.) Pulse trains for two channels are superimposed and transmitted as one train at half the original pulse recurrence frequency: the two trains are separated at the receiver. The bandwidth reduction thus obtained is compared with that theoretically possible.

621.396.61:621.396.97 3540
The Design and Operation of High-Power Broadcast Transmitter Units with their Outputs combined in Parallel—T. C. Macnamara, A. B. Howe, and P. A. T. Bevan. (*Jour. IEE* (London), part III, vol. 95, pp. 183–198; May, 1948. Discussion, pp. 198–202; summary, *ibid.*, part I, vol. 95, pp. 269–272; June, 1948.) The method of ensuring equality of the rf output voltage and modulation amplitude of the separate transmitter units is described. The circuits for combining and matching a variable number of transmitters to a common load are detailed and special design features are discussed relating to the parallel operation of transmitters, including the principle of "drive suppression" for protection purposes. The method, developed by the BBC, is used for the Droitwich medium-wave transmitter, the antenna input power of 400 kw being provided by two 200-kw units in parallel. An antenna power of 800 kw for the Ottringham long-wave station of the BBC European service was obtained by means of four similar 200-kw units in parallel. The design of the modulated-amplifier output circuit, and of a rf impedance monitor for fault protection, are discussed in appendices.

621.396.619.13 3541
Energy Distribution in the Spectrum of a Frequency Modulated Wave: Part 2—A. S. Gladwin. (*Phil. Mag.*, vol. 38, pp. 229–251; April, 1947.) Continuation of 2182 of 1945. The energy distribution can be calculated approximately when the energy spectrum and the statistical time distribution of amplitude of the modulating wave are known. The modulating wave is replaced by a synthetic wave form with the same energy spectrum and time distribution of amplitude. The sideband spectrum can thus be calculated; its asymptotic nature for very small or very large deviation ratios is demonstrated. For very small deviation ratios, the shape of the sideband spectrum is determined wholly by the energy spectrum of the modulating wave, and for very large deviation ratios wholly by the time distribution of amplitude of the modulating wave. Examples of spectra for modulation by telephonic signals are given.

621.396.619.23 3542
Some Aspects of the Design of Balanced Rectifier Modulators for Precision Applications—D. G. Tucker. (*Jour. IEE* (London), part III, vol. 95, pp. 161–172; May, 1948.) The performance of ring and Cowan types of modulator is shown to depend on rectifier characteristics, the circuit impedance in which the modulator operates, the resistance of the carrier generator, and the carrier voltage. The main features discussed are efficiency, stability, production of unwanted modulation products, impedance, and carrier leak. The design of a ring modulator is described in which the input impedance remains relatively constant over the cycle of carrier voltage.

621.396.619.23:621.317.35 3543
The Effects of an Unwanted Signal Mixed with the Carrier Supply of Ring and Cowan Modulators—D. G. Tucker. (*Jour. IEE* (London), part III, vol. 95, pp. 173–176; May, 1948.) A frequency analysis of the modulating function is used to investigate modulator performance. It is concluded that, for both ring and Cowan modulators, the primary modulating effect is, in general, largely independent of the resistance of the circuit supplying the carrier. When the frequency of the input signal is equal to the carrier frequency, the output of the difference frequency between the carrier and unwanted signal is zero for an ideal modulator and increases as the resistance of the carrier-supply circuit is decreased.

621.397.61 3544
TV Transmitter Design: Parts 2 and 3—Hamilton. (*See* 3532.)

621.396.61.029.64+621.392.029.64 3545
Microwave Transmission Design Data [Book Review]—T. Moreno. McGraw-Hill, New York, N. Y., 1948, 248 pp., $4.00. (*Electronics*, vol. 21, pp. 216, 218; August, 1948.) "The general topics covered are: transmission-line theory as applied to microwave components; coaxial lines, and flexible cables; waveguides, giving practical design data for structures, bends, tees, transformers, obstacles, windows, and couplings; waveguides filled with dielectric material both completely and partially; and cavity resonators."

VACUUM TUBES AND THERMIONICS

537.58:621.385.1 3546
Space Charge Current Theory and the Mechanical Impulse of the Electrons—Selényi. (*See* 3393.)

621.314.653 3547
The Ignitron Valve — Notes on Operation under Experimental Conditions—H. de B. Knight. (*Jour. Sci. Instr.*, vol. 25, pp. 273–275; August, 1948.) Data for Type BK22 and Type BK24 ignitrons.

621.385.029.63/.64 3548
Effect of Passive Modes in Traveling-Wave Tubes—J. R. Pierce. (Proc. I.R.E., vol. 36, pp. 993–997; August, 1948.) As the beam current is increased, the local fields due to the bunched beam become appreciable in comparison with the fields propagated longitudinally. The effect is to reduce gain, to increase the electron speed for optimum gain, to introduce a lower limit to the range of electron speeds for which gain is obtained, and to change the initial loss.

621.385.029.63/.64 3549
Field Theory of Traveling-Wave Tubes—L. J. Chu and J. D. Jackson. (Proc. I.R.E., vol. 36, pp. 853–863; July, 1948.) The problem of a helix-type traveling-wave amplifier is solved, under certain simplifying assumptions, as a boundary-value problem. The beam causes the normal mode to break up into three modes with different propagation characteristics. Over a finite range of electron velocities, one of these waves has a negative attenuation; outside this range, the propagation constants of all three waves are purely imaginary. Numerical examples for a specific tube show the effect of beam current and beam radius. The initial conditions, signal level, and limiting efficiency are also investigated.

621.385.032.21:[546.86+546.36 3550
Optical and Photoelectrical Properties of Antimony-Caesium Cathodes—N. D. Morgulis, P. G. Borzyak, and B. I. Dyatlovitskaya., (*Bull. Acad. Sci.* (U.R.S.S.) *sér. phys.*, vol. 12 pp. 126–143; March and April, 1948. In Russian.)

621.385.032.216:537.533.8 3551
Electron Emission from Oxide-Coated Cathodes under Electron Bombardment—T. J. Jones. (*Nature* (London), vol. 161, p. 846; May 29, 1948.) Results of experiments designed to throw light on the disagreement between the conclusions of Johnson (1482 of 1945, 281 and 282 of 1947) and those of Pomerantz (3107, 3467, and 3468 of 1946). The secondary-to-primary ratio δ_m was usually found to change in the lower part of the temperature range 20° to 850°C, but at the higher end all targets gave values between 5 and 10 independent of temperature; this is in general agreement with Johnson. The variations at the lower temperatures were found to be due to the resistance of the oxide coatings. No enhanced thermionic emission was observed and a possible explanation of this effect noted by the other workers is given.

621.385.1 3552
Applications and Mounting Details of Rimlock and Mazda-Medium Valves—G. Giniaux. (*TSF Pour Tous*, vol. 24, pp. 89–91; April, 1948.) Suitable output circuits are given and also a circuit diagram and response curve for a complete ac/dc receiver using these tubes. See also 2664 of October.

621.385.1:621.396.822:621.3.076.12 3553
Methods of Compensating the Various Actions of the Shot Effect in Valves and Connected Circuits—Strutt and van der Ziel. (*See* 3358.)

621.385.2 3554
Total Emission Damping—J. Thomson. (*Nature* (London), vol. 161, p. 847; May 29, 1948.) A mathematical analysis of the damping due to a nonconducting diode connected across

a tuned circuit. It is assumed that all electrons are emitted with a constant speed at right angles to the cathode and are acted upon by a constant retarding field superimposed on the small alternating field. The results are in agreement with the measurements of other workers noted in 3109 of 1946 (Smyth), 3343 of 1947 (van der Ziel and Vershel), and 3344 of 1947 (van der Ziel).

621.385.3 3555

Positive-Grid Characteristics of a Triode—G. W. Wood. (Proc. I.R.E., vol. 36, pp. 804–808; June, 1948.) The Jaffé equation (see 2577 of 1944) for reduced current density in a diode is applicable to a triode with a positive grid. Although experimental conditions and electrode geometry differed from Jaffé's theoretical assumptions, the characteristics found for receiving triodes were in reasonable agreement with the theory.

621.385.3:621.396.813 3556

Microphonism in a Subminiature Triode—V. W. Cohen and A. Bloom. (Proc. I.R.E., vol. 36, pp. 1039–1048; August, 1948.) The simple theory of the symmetrical plane triode is applied to the calculation of the change in anode current as a function of motion of the grid and cathode. Experimental investigations to determine the mechanical origin of several different forms of microphonyare also discussed.

621.385.5:621.317.723 3557

H. F. Pentodes in Electrometer Circuits—K. D. E. Crawford. (*Electronic Eng.*, vol. 20, pp. 227–231; July, 1948.) Experimental investigation of the suitability of certain British tubes. When the grid potential is made sufficiently negative to prevent electrons from actually reaching the grid, a current of the order of 10^{-10} amp exists in the opposite direction. This current produces noticeable effects when grid leaks of 100 MΩ or more are used; its causes were investigated by Metcalf and Thompson (1931 abstracts, *Wireless Eng.*, p. 98). Low anode voltage, anode current, and heater voltage help to reduce this grid current. Special circuit arrangements to widen the limits of useful operation of the tubes are discussed. For similar American work, see 1673 of July (Nielsen).

621.385.83.032.29 3558

Electron Optics and Space Charge in Simple Emission Systems with Circular Symmetry—O. Klemperer and B. J. Mayo. (*Jour. IEE* (London), part III, vol. 95, pp. 135–141; May, 1948. Summary, *ibid.*, part I, vol. 95, p. 273; June, 1948.) The properties of such systems depend largely on the spacings between the electrodes. For any grid-anode spacing, there is a certain grid-cathode spacing for which the current density in the focused beam is a maximum. For this spacing, both the divergence and the spherical aberration of the emitted electron beam have minimum values. Discussion shows that a direct application of first-order optical laws does not explain the crossover in an electron gun.

621.385.831:621.318.25 3559

Demagnetising Valves—W. Grey Walter, H. W. Shipton, and W. J. Warren. (*Electronic Eng.*, vol. 20, p. 235; July, 1948.) When the input stages of high-gain balanced low-frequency amplifiers, such as are used in biological research, are heated with raw ac, there is a residual ripple. The magnetic component of this ripple can be eliminated by first demagnetizing the tubes themselves in a decreasing ac field. Reduction of the electrostatic component was discussed in 3478 of 1947 (Grey, Walter, and Brooks).

621.385.832 3560

Repeller Storage Tube—H. Klemperer and J. T. deBettencourt. (*Electronics*, vol. 21, pp. 104–106; August, 1948.) The tube possesses internal "memory" and can discriminate between periodically recurring and new information, but charge leakage during the interval between scans makes this discrimination imperfect. The beam velocity is chosen to bring as many electrons to the bombarded surface as leave it on account of secondary emission. The potential of the beam trace on the storage surface is then governed by the potential of the collector grid. Output is considerably improved by applying a high negative bias to the repeller electrode, which is attached to the back of the insulating storage plate.

621.385.832 3561

More Cathode-Ray Tube Data—D. W. Thomasson. (*Wireless World*, vol. 54, pp. 296–297; August, 1948.) Extension of the list noted in 1835 of July.

621.385.832:621.386 3562

New X-Ray Tube Magnifies 500 Times—(*Electronic Ind.*, vol. 2, p. 11; June, 1948.) The X-ray image on a fluorescent screen causes electrons to be emitted from an adjacent photoelectric surface. These electrons are accelerated and focused on a phosphor screen, giving an image ½ of the original size but 500 times as bright. An optical system restores the image to its initial size.

621.385.832:621.397.3 3563

Flying Spot Designed for Television Studio Scanning—(*Tele-Tech*, vol. 7, pp. 42–43, 67; June, 1948.) Short account of the RCA Type 5W(P15) cathode-ray tube and its use for the television of test patterns, films, or announcements.

621.396.615.141.2 3564

The Cavity Magnetron—H. A. H. Boot and J. T. Randall. (*Jour. IEE* (London), part III, vol. 95, pp. 130–134; May, 1948.) Discussion on 890 of April.

621.396.615.141.2 3565

The High-Power Pulsed Magnetron: A Review of Early Developments—E. C. S. Megaw. (*Jour. IEE* (London), part III, vol. 95, pp. 130–134; May, 1948.) Discussion on 891 of April.

621.396.615.141.2 3566

The High-Power Pulsed Magnetron. Development and Design for Radar Applications—W. E. Willshaw, L. Rushforth, A. G. Stainsby, R. Latham, A. W. Balls, and A. H. King. (*Jour. IEE* (London), part III, vol. 95, pp. 130–134; May, 1948.) Discussion on 892 of April.

621.396.615.142.2 3567

An Analysis of Klystron Reflector Performance—H. Motz. (*Jour. IEE* (London), part III, vol. 95, pp. 295–301; July, 1948. Summary *ibid.*, part I, vol. 95, pp. 362–363; August, 1948.) The electrostatic potential distribution in the reflector space is calculated for three typical reflector designs, and the transit time for electrons moving along the tube axis is calculated as a function of the high-frequency voltage picked up in the resonator gap. The bunching of electrons and the energy exchange between electrons and resonator are discussed in detail and it is shown that the tube characteristics may be correlated with reflector design. Off-axis electron paths are computed and the effect on the tube impedance of focusing by the reflector is considered.

621.396.615.142.2:537.291:537.525.92 3568

Influence of Space Charge on the Phase Focusing of Electron Beams—J. Labus. (*Z. Naturf.*, vol. 3a, pp. 52–61; January, 1948.) An investigation as to whether the occurrence of charge concentration, or electron bunching, in tubes of the klystron type, can be hindered by space charge effects and, consequently, by the repulsive forces between the electrons. Such forces are found to have no effect on the bunching. Because of the lengthened electron path time, the bunching is favored by the use of a lower degree of modulation. If the space charge exceeds a critical value, neither charge nor current peaks can occur at the end of the drift space, however great the degree of modulation; the peaks recede into the drift space. Below this value of space charge, the efficiency is 44 per cent. With an infinitely short control space, the theoretical efficiency may reach 58 per cent, but energy exchange within the drift space results in a value between 58 and 44 per cent, depending on the space charge.

621.385.1:621.396.97 3569

Daten, Kennlinien und Schaltungen der deutschen Rundfunkröhren und ausführliche Anwendungsbeispiele. (Data, Characteristics and Connections of German Broadcasting Valves, and Detailed Application Examples.) [Book Review]—F. Kunze. Funkschau-Vertrieb Wilhelm Wolf, Potsdam, 1947. (*Phys. Blätter*, vol. 4, no. 3, p. 131; 1948.)

621.385.3.029.64+621.396.615.142.2 3570

Klystron and Microwave Triodes [Book Review]—D. R. Hamilton, J. K. Knipp, and J. B. H. Kuper (Eds). McGraw-Hill, New York, N. Y., 1947, 533 pp., $7.50. (*Electronics*, vol. 21, pp. 214, 216; August, 1948.) Volume 7 of the MIT Radiation Laboratory series. An advanced theoretical analysis, including second-order effects and large-signal conditions as well as the usual small-signal first-order theories. Descriptive material is perforce excluded.

MISCELLANEOUS

058:621.001 3571

Almanach des Sciences, 1948—L. de Broglie (Ed.). This publication, noted in 2674 of October, can be obtained from: Éditions de Flore, 10, rue Jean-du-Bellay, Paris (IVe).

06.064 Paris:621.396/.397 3572

Radio and Television Salon at the Paris Fair (1st–17th May 1948)—(*Onde Élec.*, vol. 28, pp. 282–286; July, 1948.) A general account, discussing the special features of the receivers on view, tropicalization, models for export, miniature sets, electroacoustics, dielectric heating equipment, radar, marine depth-sounding recorder, and measurement apparatus.

06.064 Paris:621.396 3573

Technical Novelties at the Paris Fair—H. Gilloux. (*Radio Franç.*, pp. 4–8; June, 1948.) An illustrated account of selected exhibits, including receivers, radiograms, measuring equipment, etc.

53 Planck 3574

Max Planck—(*Phys. Blätter*, vol. 4, no. 4, pp. 133–180; 1948.) Comprehensive accounts, by several authors, of Planck's life and work, with many tributes from friends and former pupils. See also 1841 of July and 2675 of October.

654.19(494) 3575

Radio and the PTT—W. Felix. (*Tech. Mitt. Schweiz. Telegr.-TelephVerw.*, vol. 26, pp. 120–126; June 1, 1948. In German and French.) Discussion of the various activities of the Swiss PTT in the radio field.

621.396 3576

Essentials of Radio [Book Review]—M. Slurzberg and W. Osterheld. McGraw-Hill, New York, N. Y., 1948. 806 pp., $5.00. (*Electronics*, vol. 21, pp. 226–227; August, 1948.) "The authors . . . have presented here at an intermediate level, the principles of operation of the basic circuits and circuit elements used in conventional radio receivers, as essential background knowledge for understanding electronic circuits."

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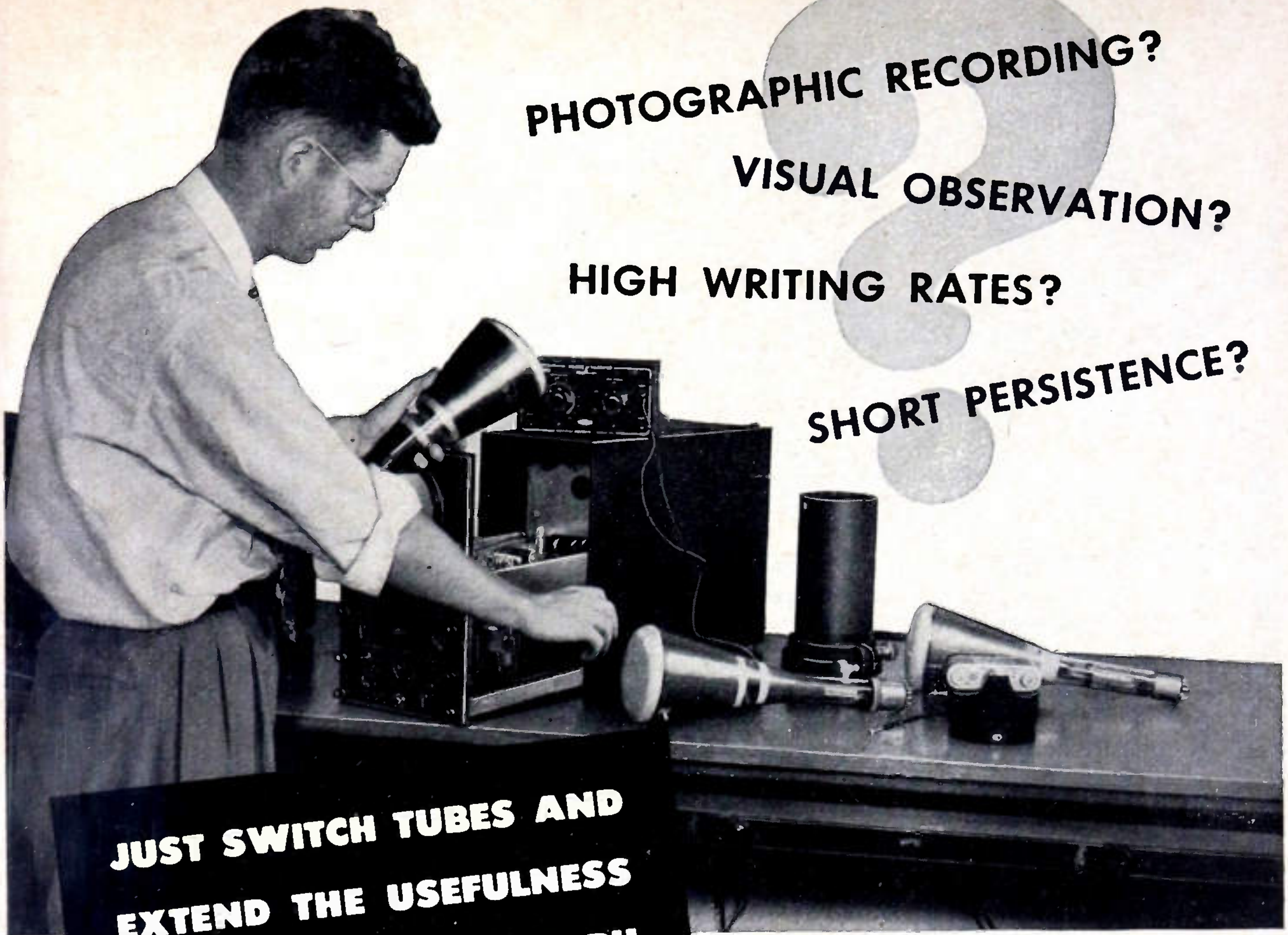

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SECTION MEETINGS

ATLANTA

"Beam Deflection Mixer Tubes," by E. W. Herold, RCA Laboratories; October 8, 1948.

BALTIMORE

"International Telecommunications," by E. K. Jett, A. S. Abell Company; October 28, 1948.

BUFFALO-NIAGARA

"Engineering Considerations in Locating Broadcast Transmitters," by S. L. Bailey, Jansky and Bailey, Consulting Engineers; October 14, 1948.

CEDAR-RAPIDS

"Television—Its Mechanism and Promise," by W. L. Lawrence, Radio Corporation of America; October 13, 1948.

CINCINNATI

"Bandwidth Reduction in Communications Systems," by W. G. Tuller, Melpar, Inc.; September 21, 1948.

"Mobile Telephone," by R. E. Kolo, Cincinnati and Suburban Bell Telephone Company; October 19, 1948.

CLEVELAND

"Citizens Radio Equipment," by A. Gross, Citizens Radio Corporation; October 28, 1948.

COLUMBUS

"Radio in Postwar Germany," by K. L. Tyler, Ohio State University; October 27, 1948.

CONNECTICUT VALLEY

"Properties of Aluminum Oxide Film as Used in Electrolytic Capacitors; Sprague Electric Company; October 16, 1948.

DAYTON

"Aircraft Computers," by O. H. Schuck, Minneapolis-Honeywell Company; October 14, 1948.

"Stratovision," by C. E. Nobles, Westinghouse Electric Corporation; November 11, 1948.

DES MOINES-AMES

"Television—Its Mechanism and Promise," by W. L. Lawrence, Radio Corporation of America; October 15, 1948.

DETROIT

"Printed Electronic Circuits," by C. Brunetti, National Bureau of Standards; September 17, 1948.

"The Pictorial Situation Display in Air Navigation and Traffic Control," by H. H. Spencer, Radio Corporation of America; October 15, 1948.

EMPORIUM

"Engineering Considerations in Locating Broadcast Transmitters," by S. L. Bailey, Jansky and Bailey, Consulting Engineers; October 14, 1948.

INDIANAPOLIS

"Problems in Television," by B. E. Shackelford, 1948 President, The Institute of Radio Engineers; September 16, 1948.

"Television Power Supplies, Synchronizing Circuits and Deflection Systems," by C. Honeywell, The Hallicrafters Company; October 22, 1948.

LONDON

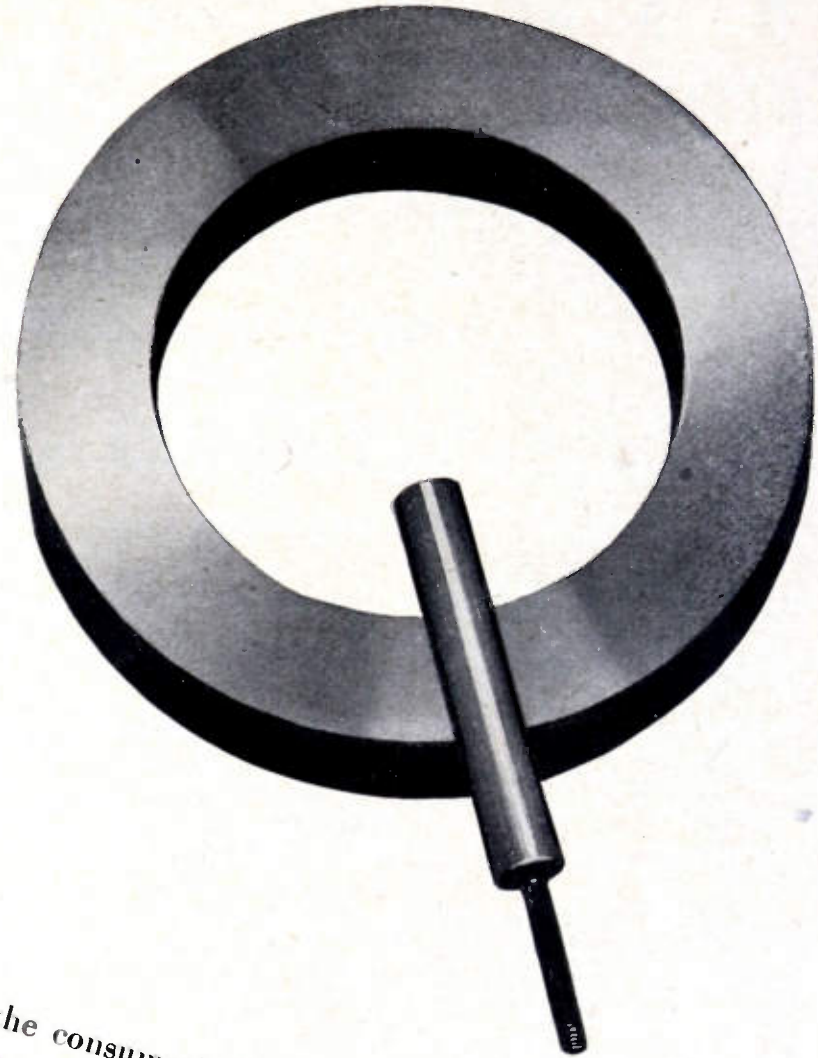
"Dry Batteries Today," by H. T. Tipple, Canadian National Carbon Company; September 24, 1948.

LOS ANGELES

"Voltage Stabilization by Two Terminal Stabilizing Elements," by D. Rutland, North American Aviation; October 19, 1948.

"Distributed Amplification," by N. B. Schrock, Hewlett-Packard Company; October 19, 1948.

(Continued on page 36A)



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SECTION MEETINGS

(Continued from page 34A)

LOUISVILLE

"Inspection of New Courier-Journal Building," by K. White, American Elevator Company, and J. W. May, American Air Filter Company; October 15, 1948.

NEW YORK

"Application of Microwave Radio Techniques to Telegraphy and to Sound and Video Program Transmission," by J. Z. Millar, Western Union Company; October 6, 1948.

OTTAWA

"Demonstration of the National Research Council's Radar Equipped Motor Vessel 'Radel.'" by H. R. Smythe, National Research Council; October 9, 1948.

PHILADELPHIA

"Intercarrier-Sound Circuits for Television Receivers," by A. A. Barco, RCA Laboratories; November 4, 1948.

PITTSBURGH

"The Airborne Magnetometer in Geophysical Exploration," by R. D. Wyckoff, Gulf Research and Development Company; October 11, 1948.

PORTLAND

"Design and Operation of a 5-Kw Television Transmitter," by H. W. Granberry, General Electric Company; October 29, 1948.

PRINCETON

"Electronic Properties of Semi-Conductors and the Transistor," by W. H. Brattain and J. Bardeen, Bell Telephone Laboratories; October 14, 1948.

"Electronic Reading Aids for the Blind," by V. K. Zworykin, L. E. Flory, and W. S. Pike, RCA Laboratories; November 11, 1948.

ROCHESTER

"Engineering Considerations in Locating Broadcast Transmitters," by S. L. Bailey, Jansky and Bailey, Consulting Engineers; October 14, 1948.

"A Television Station Selector Using Die-Stamped Inductances," by A. D. Sobel, A. W. Franklin Manufacturing Corporation; November 8, 1948.

"A Discussion of Image Sharpness in Photography and Television," by O. H. Schade, Radio Corporation of America; November 8, 1948.

"Application of Subminiature Tubes," by R. K. McClintock, Sylvania Electric Products, Inc.; November 8, 1948.

"The Transitrol, An Experimental AFC Tube," by J. Kurshan, RCA Laboratories; November 8, 1948.

"A New Low-Noise, Low-Microphonic Miniature Tube," by C. R. Knight and A. P. Haase, General Electric Company; November 8, 1948.

"What's When in America," by K. W. Jarvis, Consulting Engineer November 8, 1948.

"Radio Receiver Engineering Department Organization," by A. G. Rogers, Emerson Radio and Phonograph Corporation; November 9, 1948.

"Developments in Germanium Crystals," by S. T. Martin and H. Heins, Sylvania Electric Products, Inc.; November 9, 1948.

"A Television Distribution System for Laboratory Use," by J. Fisher, Philco Corporation; November 9, 1948.

"A Direct Coupled Video and ACC System for Television Receivers," by H. R. Shaw, Colonial Radio Corporation; November 9, 1948.

(Continued on page 38A)

FREQUENCY SHIFT EXCITER —Provides RF drive and frequency shift keying to transmitter

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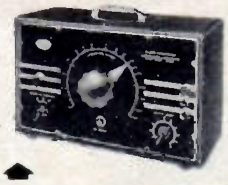

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SECTION MEETINGS

(Continued from page 36A)

"A Pulse-Cross Generator for Television Receiver Production," by R. P. Burr, Hazeltine Electronics Corporation; November 9, 1948.

"Lightweight Pickup Design for Microgroove Record Playing," by R. P. Haines, Philco Corporation; November 10, 1948.

"What Constitutes High Fidelity," by H. P. Fletcher, Bell Telephone Laboratories, J. K. Hilliard Altec Lansing Corporation; and C. J. LeBel, Consultant; November 10, 1948.

"High-Quality Audio System for Radio Receivers," by R. S. Anderson and B. E. Atwood, Stromberg-Carlson Company; November 10, 1948.

"Front Ends of Television Receivers," by J. O. Silvey, General Electric Company; November 10, 1948.

"A Picture-And-Sound-Modulated Generator for Television Receiver Production," by W. R. Stone, Hazeltine Electronics Corporation; November 10, 1948.

"Industrial Applications of Photography," by W. Clark, Eastman Kodak Company; November 10, 1948.

Sacramento

"TV Equipment and Components of Systems," by R. B. Newman, Radio Corporation of America; October 19, 1948.

St. Louis

"The Baldwin Electronic Organ," by J. F. Jordan, Baldwin Piano Company; October 20, 1948.

San Francisco

"The Application of Radar Techniques to Modern Aviation," by R. J. Shank, Hughes Aircraft Company; October 13, 1948.

Seattle

"Television Station Equipment and its Functions," by H. W. Granberry, University of Washington; October 25, 1948.

Toledo

"Electronic Applications to Industrial Weighing," by D. D. Mallory, Toledo Scale Company; October 18, 1948.

Toronto

"Objectives of a University Training in Electrical Engineering," by G. F. Tracy, University of Toronto; October 25, 1948.

Twin Cities

"Computers for Aeronautical Navigation," by H. Schuck, Minneapolis-Honeywell Company; October 19, 1948.

Washington

"A Field Test of Ultra-High-Frequency Television in the Washington Area," by R. D. Kell and G. H. Brown, RCA Laboratories; October 11, 1948.

"Application of Microwave Radio Techniques to Telegraphy and to Sound and Video Program Transmission," by J. Z. Millar, Western Union Telegraph Company; November 8, 1948.

SUB-SECTIONS

Hamilton

"Current Trends in Radio Engineering," by R. A. Hackbusch, Stromberg-Carlson Company; October 18, 1948.

"Television Antennas," by H. S. Dawson, Canadian General Electric Company; November 1, 1948.

Lancaster

"Aberrations of Lenses," by R. Kingslake, Eastman Kodak Company; October 12, 1948.

Long Island

"Tape Speed as it Affects Magnetic Recording," by R. H. Ranger, Rangertone, Inc.; October 13, 1948.

MICROWAVE PLUMBING

10 CENTIMETER

"S" BAND CRYSTAL MOUNT, gold plated, with 2 type "N" connectors $12.50
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MAGNETRON COUPLING to ⅞" rigid coax with TR pickup loop, gold plated $7.50

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RANDOM LENGTHS of waveguide, 6 in. to 18 in. long .... $1.10/ft.
WAVEGUIDE RUN, 1¼" x ½" guide, consisting of 4 ft. section with rt. angle bend on one end and 2", 45 deg. bend other end $8.00
WAVEGUIDE SECTION, 1¼" x ½" choke to choke, 4 ft. long $6.75
DUMMY LOAD, TS 332/UP $22.50
"X" Band pressurizing gauge section, with 15-lbs. gauge and pressurizing nipple $18.50
45 DEG. TWIST, 6" Long $10.00
12" SECTION, 45 deg. twist, 90 deg. bend $6.00
11" STRAIGHT WAVEGUIDE section choke to cover. Special heavy construction, silver plated $4.50
15 DEG. BEND, 10" choke to cover $4.50
5 FT. SECTIONS, choke to cover $14.50
18" FLEXIBLE SECTION $17.50
"E" and "H" PLANE BEND $12.50
BULKHEAD FEED THRU $15.00
"X" BAND WAVEGUIDE, 1¼" x ⅝" OD, 1/16" wall, aluminum .... per ft. $ .75
WAVEGUIDE, 1" x ½" I.D. per ft. $1.50
TR CAVITY for 724-A TR tube $3.50
3" FLEX SECTION, square flange to circular flange adapter $7.50
724 TR tube (41-TR-1) $2.50
WAVEGUIDE SECTION, CG 251/APS-15A, 26" long choke to cover, with 180 deg. bend of 2½" rad. at one end $6.00
SWR MEAS. SECTION, 4" L. with 2 type "N" output probes mtd full wave apart. Bell size guide. Silver plated $10.00
ROTARY JOINT with slotted section and type "N" output pickup $8.50
WAVEGUIDE SECTION, 12" long choke to cover, 45 deg. twist & 2½" radius, 90 deg. bend $4.50
SLUG TUNER/ATTENUATOR, W.E. guide, gold plated $6.50
TR/ATR DUPLEXER section with iris flange .. $8.00
TWIST 90 deg., 5" choke to cover, w/press nipple $6.50
WAVEGUIDE SECTIONS 2½ ft. long, silver plated, with choke flange $5.75
WAVEGUIDE 90 deg. bend E plane, 18" long .. $4.00
ROTARY JOINT, choke to choke $6.00
ROTARY JOINT, choke to choke, with deck mounting $6.00
S-CURVE WAVEGUIDE, 8" long cover to choke $3.50
DUPLEXER SECTION for 1B24 .... $10.00
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 db $15.00
CU 108/APS 33 Directional coupler, 25 db $15.00
CG 176/AP Directional coupler, 20 db $18.00
FLEX. WAVEGUIDE .... $4.00/Ft.
"X" BAND calibrated attenuator $85.00
SHIELDED KLYSTRON tube mounts with rough attenuator outputs $90.00
2½" FLEXIBLE SECTION, cover to cover $5.00
SHORT ARM "T" section, with additional choke output on vertical section $4.00

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
KBAND Rotary joint $45.00
ADAPTER, rd. cover to sq. cover $5.00
MITRED ELBOW and 8 sections choke to cover $4.50
DEHYDRATING UNIT, 60 lb. capacity, 115 v, 60 cy operation, 2' x 22" x 15", New and complete $425

VARISTORS W.E.

D-171631 ... $.95	D-168549 ... $.95
D-167176 ... $.95	D-162482 .. $3.00
D-168687 ... $.95	D-99136 ... $1.65
D-171812 ... $.95	D-166271 .. $2.50
D-171528 ... $.95	D-162356 .. $1.50
D-163298 ... $.95	D-161871A $2.85

THERMISTORS—W.E.

D-167332 (tube) .... $.95
D-170396 (bead) .... $.95
D-167613 (button) .... $.95
D-166228 (button) .... $.95
D-164699 FOR MTG. in "X" Band Guide $2.50
D-167018 (tube) .... $.95

COAX CONNECTORS

831SP ..... $.35	UG 254/U .. $.75
831AP ..... $.35	UG 255/U . $1.25
831HP ..... $.15	UG 146/U . $1.00
UG 21/U ... $.85	UG 85/U .. $1.25
UG 86 U ... $.95	

Homedell male to type "N" male adapter $1.25
D-166366 Baby "N" .... $ .85
Adapatre Cable Ass'y, Type "N" Male to Type "N" Female $2.25
Adapter Cable Ass'y, Sperry Male to Type "N" Male $2.25
Connector, #49579 for RG 18/U $1.80

INVERTER PE 218

Input: 27.5 V DC. 90 AMPS
Output: 115 V AC. 400 CY. 13 AMPS
1500 Volt—Amperes .9 PF
New, Original Packing $49.95
Slightly used, ex. cond. 25.00
Also available: PE 206. 800CY. 80V. New $15.75

MAGNETRONS

TUBE	FRQ. 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
2J38 Pkg.	3249-3263 mc.	5 KW.	25.00
2J39 Pkg.	3267-3333 mc.	87 KW.	25.00
2J55 Pkg.	9345-9405 mc.	50 KW.	25.00
2J61	3000-3100 mc.	35 KW.	65.00
2J62	2914-3010 mc.	35 KW.	65.00
3J31	24,000 mc.	50 KW.	55.00
5J30			39.50
714AY			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			20.09

MAGNETRONS 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

TUNABLE PKG'D "CW" MAGNETRONS

QK61 ..2975-3200 Mcs QK62 ..3150-3375 Mcs
QK60 ..2800-3026 Mcs QK59 ..2675-2900 Mcs
New .... $65 each

COAX CABLE

RG 17/U. 52 ohm imp. .... $.50/ft.
RG 18/U. 52 ohm imp. armored .... .51/ft.
RG 23/U. twin coax, 125 ohm imp. armored .50/ft.
RG 28/U. 50 ohm imp. pulse cable. Corona min. starting voltage 17 KV. .... .50/ft.
RG 35/U. 70 ohm imp. armored .... .50/ft.
RG 57/U. 95 ohm imp. twin coax. .... .35/ft.

MICROWAVE TEST EQUIPMENT

THERMISTOR BRIDGE: Power meter I-203-A. 10 cm. mfg W.E. Complete with meter, interpolation chart, portable carrying case $72.50
Bell Labs. Dual Mount mixer-beacon assemblies, 2 complete mixer-beacon mounts on gold-plated waveguide section $50.00
Slotted Line, Bell Labs. 1⅛" x 5/16" guide, gold plated $150.00
TS-238 GP. 10 cm. Echo box with resonance indicator and micrometer adjust cavity. 2700 to 2900 Mcs calibrated as shown $85.00
TS 108-AP dummy load $65.00
3 cm. Wavemeter, Micrometer head mounted on X-Band guide. Freq. range approx. 7900 to 10.000 Mc. $75.00

DYNAMOTORS

Type	Input Volts	Input Amps	Output Volts	Output Amps	Radio Set	Price
BD 77KM	14	40	1000	.350	BC 191	$20.00N
						14.00LN
PE 73	28	19	1000	.350	BC 375	24.50N
DM 21CX	28	1.6	235	.090	BC 312	3.45N
DM 25	12	2.3	250	.050	BC 367	2.49LN
DM 28R	28	1.25	275	.070	BC 348	8.75N
DM 33	28	7	540	.250	BC 456	5.50
DM 42	14	46	1030	.050		
			2/8			
CWD 21AAX	13	12.6	400	.135		17.50N
	26	6.3	800	.020		
			9AC	.112		
D 104	12		225	.100		14.95N
			440	.200		
DA 3A*	28	10	300	.260	SCR 522	8.95N
			150	.010		
			14.5	5		
#5053	28	1.4	250	.060	APN 1	3.95N
DA 7A	26.5		1100	.400	TA 2J	25.00N
PE 86	28	1.25	250	.060	RC 36	3.95
23350	27	1.75	285	.075	APN-1	3.50N
35X045B	28	1.2	250	.060		3.50N
ZA .0515	12/24	1/2	500	.050		3.95N
B-19 PACK	12	9.4	275	.110	Mark II	
			500	.050		9.95N

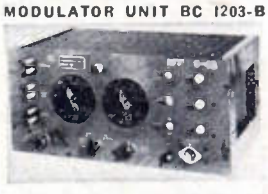
N—New. LN—Like new.
*—Less, Filter Box and Relays

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, revr-ampl, duplexer, HV supply, blower, pulse xfmr. 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

PULSE EQUIPMENT

MODULATOR UNIT BC 1203-B



Provides 200-4,000 PPS. Sweep time: 100 to 2,500 microsec. in 4 steps. fixed mod. pulse, suppression pulse, sliding modulating pulse, blanking voltage, marker pulse, sweep voltages, calibration voltages, fil. voltages. Operates 115 vac. 50-60 cy. Provides various types of voltage pulse outputs for the modulation of a signal generator such as General Radio #804B or #804C used in depot bench testing of SCR 695, SCR 595, and SCR 535. New as shown $125.00
MIT. MOD. 3 HARD TUBE PULSER: Output Pulse Power: 114 KW (12 KV at 12 amp). Duty Ratio: .001 max. Pulse duration: .5, 1.0, 2.0 microsec. Input voltage: 115 v, 400 to 2400 cps. Uses 1-715-B, 1-829-B, 3-72's, 1-73. New $110.00
APQ-13 PULSE MODULATOR, Pulse Width .5 to 1.1 Micro Sec. Rep. rate 624 to 1348 Pps. Pk. Pwr. out 35 KW. Energy 0.018 Joules ... $49.00
TPS-3 PULSE MODULATOR. Pk. power 50 amp. 24 KV (1200 KW pk); pulse rate 200 PPS, 1.5 microsec; pulse line impedance 50 ohms. Circuit—series charging version of DC Resonance type. Uses two 705-A's as rectifiers. 115 v. 400 cycle input. New with all tubes $49.50
APS-10 MODULATOR DECK, Complete, less tubes $75.00
APS-10 Low voltage power supply, less tubes $18.50

PULSE NETWORKS

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

PULSE TRANSFORMERS

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 $12.00
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 $2.00
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. $19.50
W. E. #D169271 Hi Volt input pulse Transformer $9.95
G.E. K2450A Will receive 13KV. 4 micro-second pulse on pri., secondary delivers 14KV. Peak power out 100KW GE $15.00

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. Prices subject to change without notice.

COMMUNICATIONS EQUIPMENT CO.

131-R Liberty St., New York, N.Y. **Digby 9-4124**

STUDENT BRANCH
MEETINGS

Alabama Polytechnic Institute, IRE Branch

Election of Officers; October 11, 1948.

University of Arizona, IRE-AIEE Branch

"Large Business versus Small Business for the Graduate Engineer," by S. Becker, University of Arizona; October 27, 1948.

California State Polytechnic College, IRE Branch

"The Relationship of the Engineering Society to the Student," by B. E. Shackelford, 1948 President, The Institute of Radio Engineers; September 29, 1948.

Iowa State College, IRE-AIEE Branch

"Television," by T. A. Hunter, Regional Director IRE; October 19, 1948.

State University of Iowa, IRE Branch

General Electric Company Demonstration; October 19, 1948.

Manhattan College, IRE Branch

"Insulation Testing," by W. G. Long, J. G. Biddle Company; October 20, 1948.

University of Michigan, IRE-AIEE Branch

"Electrical Engineering Applications in Geophysical Prospecting," by J. T. Wilson, University of Michigan; October 6, 1948.

"The Transistor," by R. R. Barnes, Michigan Bell Telephone Company; October 27, 1948.

Michigan State College, IRE-AIEE Branch

"What the Professional Engineering Organizations Can Do For You," by A. Rauth, Consumers Power Testing Laboratories; October 13, 1948.

College City of New York, IRE Branch

"Design of a High-Fidelity Audio System," by L. M. Glantz, Student; October 19, 1948.

"High-Fidelity Systems," by L. Feldman, Student; October 19, 1948.

University of North Dakota, IRE-AIEE Branch

"The Freeman Headbolt Heater," by E. Zobel Five-Star Manufacturing Company; October 13, 1948.

University of Notre Dame, AIEE-IRE Branch

"The South Bend AIEE and its Activities," by R. Koontz, Consulting Engineer; October 13, 1948.

Oregon State College, IRE-AIEE Branch

"IRE-AIEE Societies History and Background," by F. O. McMillan, Oregon State College; September 29, 1948.

"Student Summer Jobs," by E. P. Single, J. F. Richardson, P. Foley, M. M. Matthews, J. Lynch, D. G. Wohlgemuth, D. Poley, H. Lowry, and N. Austin, Students; October 14, 1948.

Rutgers University, IRE-AIEE Branch

"Placement of Engineers," by J. L. Price, Rutgers University; October 26, 1948.

(Continued on page 41A)

STUDENT BRANCH MEETINGS

(Continued from page 40A)

St. Louis University, IRE Branch

"Commercial Radio Licenses," by W. Fisher, Student; October 7, 1948.

"Mathematical Short Cuts," by O. Haly, Student; October 7, 1948.

"Operation of Station WEW-FM," J. Goforth, Student; October 14, 1948.

"Microwaves in your Telephone," by M. Waggoner, Southwestern Bell Telephone Company; October 21, 1948.

University of Tennessee, IRE-AIEE Branch

"Electrical Engineering and Atomic Energy," by D. W. Cardwell, University of Tennessee; October 26, 1948.

University of Wyoming, IRE Branch

"Radio-Frequency Induction Heating," by C. Kane, Student; October 7, 1948.

MEMBERSHIP

The following transfers and admissions were approved to be effective as of January 1, 1949:

Transfer to Member Grade

Berry, F. M., 1200 E. 49 St. Terr., Kansas City, Mo.
Birkenhead, W. D., 910 Second St., Manhattan Beach, Calif.
Campbell, C. A., 44 Central St., W. Concord, Mass.
Chernosky, E. J., 6608 Hillandale Rd., Chevy Chase 15, Md.
Crandon, L. H., Electronics Department, Westinghouse Research Laboratories, East Pittsburgh, Pa.
Davis, R., 2460 Silverlake Blvd., Los Angeles 26, Calif.
Davis, S., 351-99 St., Brooklyn 9, N. Y.
Dobbs, C. I., 5701 S. Hoyne Ave., Chicago 36, Ill.
Frenette, C., 54 Bougainville Ave., Quebec City, Que., Canada
Morse, M. S., Box 322, Red Bank, N. J.
Murphy, J. L., 1456 E. 54 St., Chicago 15, Ill.
Nelson, D. E., 411 W. Illinois St., Urbana, Ill.
Qureshi, M. A. A., P.O. Husain Agahi, Multan City, W. Punjab, Pakistan
Saigeon, N. D., 26783 Outer Dr., Escorse 18, Mich.
Tomiyasu, K., 2 Gray Gardens East, Cambridge 38, Mass.

Admission to Member Grade

Betts, A. L., 220 S. Husband, Stillwater, Okla.
Corwin, N. J., 2040 Baker Ave., Schenectady 8, N. Y.
Dinnick, G. M., c/o Andrew Corp., 363 E. 75 St., Chicago 19, Ill.
Eddy, R. C., 510 N. Hanover St., Elizabethtown, Pa.
Gilmer, P., 27 Beechwood Rd., Florham Park, N. J.
Greaves, H. A., Mackenzie, Rio Demerara, British Guiana, S. A.
McClain, E. F., Jr., Naval Research Laboratory, Building 1, PH, Washington 20, D. C.

(Continued on page 42A)

MEMBERSHIP

(Continued from page 41A)

Smoll, A. E., 2511 S. State, Syracuse 5, N. Y.
Spicer, W. T., 1929 E. 31 St., Baltimore 18, M.d
Stewart, R. M., Jr., Physics Department, Iowa State College, Ames, Iowa
Tapy, R. W., University of New Mexico, Albuquerque, N. Mex.
Urquhart, K. J., Sandia Base Branch, Albuquerque, N. Mex.
Wase, A. E. N., 62 Holdenhurst Ave., London N.12, England
Zielesch, B. E., 461 Oak, Wisconsin Rapids, Wis.

The following admissions to Associate grade have been approved and were effective as of December 1, 1948:

Adler, R. L., 602 W. 165 St., New York 32, N. Y.
Aggarwol, N. H., c/o L. Banarsi Das, Advocate Hoshiar Par, East Punjab, India
Allred, C. Mc., National Bureau of Standards, Washington, D. C.
Appuhamy, F. A., "San Jose," Dankotvwa, Ceylon
Ardelian, T. F., Hudson's Bay Co., Fur Trade Department (Radio), Hudson's Bay House, Winnipeg, Manitoba, Canada
Argall, J. R. C., 104-53 Ave., Lachine, Montreal 32, Ont., Canada
Barnett, H. A., Jr., Radio Station WIST, Charlotte 2, N. C.
Berry, R. F., 17945 Harwood Ave., Homewood, Ill.
Beyer, A. R., 2019 Dunstan, Houston 5, Tex.
Blais, R. L., 282 Pleasant St., Berlin, N. H.
Buchsbaum, W. H., 80-21-74 Ave., Glendale, L. I., N. Y.
Caffrey, G. H., 103 N. Chester Ave., Hatboro, Pa.
Cancro, C. A., Naval Ordnance Laboratory, Bldg., 201, White Oak, Silver Spring, Md.
Chaney, A. L., Jr., 612 Grandin Rd., Charlotte, N. C.
Christopher, O., Box 68, Air Base Branch, Belleville, Ill.
Cooper, L. V., 1930 AACS APO 731, c/o PM Seattle, Wash.
Couch, P. W., 22 Hopeland St., Dayton, Ohio
Croumlich, R. M., Box 171, New Baden, Ill.
DeDominicis, C. M., 34 Via Filippino Lippi, Milano, Italy
Dickinson, I. E., McClatchy Broadcasting Co., 7 & Eye Sts., Sacramento, Calif.
Earley, J. M., 2120 Douglass Ave., Maplewood 17, Mo.
Edmanson, J. W. E., 2436 Shakespeare Rd., Houston 5, Tex.
Elam, F. E., 1345 Madison St., N. W., Washington 11, D. C.
Endres, J. M., Jr., 29 Fulmer Ave., Llanerch, Havertown, Pa.
Erickson, E. A., 4625 Second Blvd., Detroit 1, Mich.
Farley, B. G., Bell Telephone Laboratories, Murray Hill, N. J.
Foakes, P. F. J., 16 Bouverie Rd., Chelmsford, Essex, England
Foley, T. U., 109 E. Evesham Ave., Magnolia, N. J.
Freeman, G. A., Jr., 2117 Ave. L., Galveston, Tex.
Galbraith, A. H., 43 Crestmont Rd., Bangor, Me.
Gerstein, B., 928 George St., Chicago 14, Ill.
Gibbs, L. L., 349 Orchard Dr., Dayton 9, Ohio
Graham, J. M., Rm. 309 CAP, General Electric Company, Schenectady, N. Y.
Grdseloff, L., c/o Green's Commercial Agencies, P.O.B. 600, Cairo, Egypt
Gregersen, K. J., 4408 Greenwich Village Ave., Dayton 6, Ohio
Greulich, H. J., Box 228, APO 925, c/o Postmaster, San Francisco, Calif.
Gulczynski, R. J., 2004 W. Cullerton St., Chicago 8, Ill.
Guttwein, G. K., 24 Liberty St., Long Branch, N. J.
Heppel, H. H., Jr., 2202 S. Boulevard, Houston 6, Tex.

MEMBERSHIP

Hullegard, E. W., 53 A St. Eriksgatan, Stockholm, Sweden
Hyzinski, B., 64 North Empire St., Wilkes Barre, Pa.
Innes, F. T., 1919 Arch St., Philadelphia 3, Pa.
Isaacs, S. J., 2043 Holland Ave., New York 60, N. Y.
Jones, G. E., Jr., University of Pittsburgh, Pittsburgh 13, Pa.
Jones, O. L., 3006 Nichols Ave., S. E., Washington 20, D. C.
Jose, R. S., Hdqs. 1804, AACS Group, APO 942, c/o Postmaster Seattle, Wash.
Keep, H. A., 1110 Winbern St., Houston 4, Tex.
Kelley, N. J., 123 Ontario St., Toledo 2, Ohio
Kiley, W. A., 1235 Avant Ave., San Antonio, 3 Tex.
Koenig, J. G., 129 Hornaday Rd., Pittsburgh, 10 Pa.
Kothe, E. W., 68 Prince St., Hastings-on-Hudson 6, N. Y.
Kovner, H., 400 E. 57 St., New York 22, N. Y.
Lojkasek, T., 1961 Bartolome Mitre, Buenos Aires, Argentina
Lorenz, L. H., 1250 Wakefield Dr., Houston 18, Tex.
Lundahl, S. E., 190 Nagle Ave., New York 34, N. Y.
Mack, S. J., 81 Brook St., Wollaston 70, Mass.
Mahaffey, R. E., 615 S. Darling, Angola, Ind.
Manson, W. R., 4016 44 Ave. S., Minneapolis, Minn.
Margulius, M. P., Acoyte 143-VII H, Buenos Aires, Argentina
McCarrell, S. G., 8925 Bishop St., Chicago, Ill.
McCormick, C. E., Box 1138, Grand Junction, Col.
McDonald, H. D., 10 E. 138 St., New York 35, N. Y.
McNamara, T. R., 254-31-84 Rd., Floral Park, L.I., N. Y.
Mehta, P. E., 13 Jal Chamber, Charni Road Junction, Bombay 4, India
Myers, V. V., Jr., 413 South Broadway, Albuquerque, N. Mex.
Narula, V. P., c/o G. R. Narula Esqr., Near Water Tank A, Patiala, India
New, H. J., Jr., 3504 Harrison, Amarillo, Tex.
Pacher, F. A., 1227 39 Ave., San Francisco 22, Calif.
Padilla, D. J., R-217 D, 13 St., Sandia Base, Albuquerque, N. Mex.
Parks, R. J., 10000 La Tuna Canyon Rd., Roscoe 7, Calif.
Perelman, I. B., 28 Windermere Dr., Bronxville Hgts., Yonkers, N. Y.
Perlman, A., 160 Beach 82 St., Rockaway Beach, L.I., N. Y.
Pickens, R. T., 58 W. 75 St., New York 23, N. Y.
Platt, S., 8 Stuyvesant Oval, New York 9, N. Y.
Rabinowitz, M. H., 324 E. 81 St., New York 28, N. Y.
Rahner, H. B., Georgia Institute of Technology, Box 5388, Atlanta, Ga.
Redfield, A. H., Jr., 621 Quintana Pl., N.W., Washington 11, D. C.
Reedy, J. F., 4652 Kenmore Ave., Chicago, Ill.
Roberts, J. T., Barnard Hall, North Brother Island, New York 54, N. Y.
Root, J. S., 621 Craig St. W., Montreal 3, Ont., Canada
Rosner, M., 306 E. 75 St., New York 21, N. Y.
Rothman, L., 1820-81 St., Brooklyn 14, N. Y.
Rynder, L. S., 2372 Franklin Ave., Toledo 10, Ohio
Scovill, J. M., 20 W. Garfield St., Seattle 99, Wash.
Simon, J. M., 1511 20 St., N.W., Washington, D. C.
Smith, E. A., 2235 Kensington Rd., Dayton 6, Ohio
Stauffer, C. R., 1553 N. Wells St., Chicago 10, Ill.
Sterne, H. J., 7071 N. Wilbur St., Portland 3, Ore.
Swain, J. E., Jr., 5058½ N. Wolcott, Chicago 40, Ill.
Tompson, R. N., Mathematics Department, University of Nevada, Reno, Nev.

(Continued on page 44A)

MEMBERSHIP

(Continued from page 43A)

Tufani, H. D., 1819 Humboldt Blvd., Chicago 47, Ill.
Ulander, C. G., 1209 Shepherd St., N.W., Washington, D. C.
Wagenseller, J. E., 1017 North Ave., Secane, Pa.
Wallace, R. M., 2921 Cherrywood Ave., Dallas 9, Tex.
Whitehurst, C. H., Jr., 201 N. Highland Ave., Winter Garden, Fla.
Wilson, J. O., Box 193, Espanola, N. Mex.
Wofford, W. D., 1502 Elmen, Houston 19, Tex.
Wolfe, R. E., 9306 Harford Rd., Baltimore 14, Md.
Wynn, W. W., R.F.D. 2, Maple, Ont., Canada

The following transfers to Associate grade were approved to be effective as of December 1, 1948:

Baird, G. A., Conestoga Rd., Ithan, Pa.
Bardi, G. B., 114 West 13 St., New York 11, N. Y.
Bein, W., 610 Riverside Dr., New York 31, N. Y.
Bend, C. M., Jr., 107 Farrington St., St. Paul 2, Minn.
Billingsley, F. C., 1926 Fifth Ave., Troy, N. Y.
Boehm, W. R., 1044-26 St., S., Arlington, Va.
Borgese, A. J., 115 Harding Ave., White Plains, N. Y.
Britton, C. C., 722 Grand Ave., Ames, Iowa.
Burnett, J. R., School of Electrical Engineering, Purdue University, Lafayette, Ind.
Chambers, G. R., 1945 Holliston Ave., Pasadena, Calif.
Chatterjee, A. K., 503 E. Greer St., Champaign, Ill.
Chin, F. K., 251 Victoria Ave., Hampton, Va.
Christmas, W. I., 63 Columbia Ave., Westmount, Montreal 6, Que., Canada
Clark, C. M., 3206 Nevin Ave., Richmond, Calif.
Cohen, E., 71-31 Manse St., Forest Hills, L.I., N. Y.
Culling, E. J., 107 Tenth St., N.E., Washington 2, D. C.
D'Amato, R. J., 989 New York Ave., Brooklyn 3, N. Y.
Davis, J. H., 560 E. St. John, San Jose 12, Calif.
Eisenberg, I., 1412 Vyse Ave., New York 59, N. Y.
Ellis, T. E., Jr., R.F.D. 5, Kenton, Ohio
Evenson, R. K., 463 West St., New York 14, N. Y.
Fenaughty, A. L., 2039 Stanford Ave., St. Paul X5, Minn.
Flackbert, W. D., 90 Broad St., New York 4, N. Y.
Freinmuth, W. L., 4107 Warner St., Kensington, Md.
Fusfeld, M., 710 West 173 St., New York 32, N. Y.
Gaines, W. M., 462 Third St., Schenectady, N. Y.
Godbey, J. K., 3120 Dutton Dr., Dallas 11, Tex.
Green, P. E., Jr., 226 Broad St., Oxford, N. C.
Gutsheon, W. Y., 309 W. 57 St., New York, N. Y.
Hodgin, D. M., Jr., 3206 Vine Ave., S.E., Cedar Rapids, Iowa
Huebscher, H., 1745 Eastburn Ave., New York 57, N. Y.
Hufford G., 3814 Jocelyn St., N.W., Washington 15, D. C.
Hull, D. E., 1792 37 St., Los Alamos, N. Mex.
Jacobi, W. J., Department of Electrical Engineering, Princeton University, Princeton, N. J.
Jakubowski, M. C., c/o Fairchild Aerial Surveys, Inc., 224 East 11 St., Los Angeles 15, Calif.
Jennings, S. J., 967 Main Ave., Schenectady 3, N. Y.
Joerger, J. C., 729 W. Roscoe St., Chicago 13, Ill.
Jones, P. D., Station 0-40, Sperry Gyroscope Company, Great Neck, L.I., N. Y.
Jones, W. N., Box 323, Swarthmore, Pa.
Kilby, J. S., 4924 West Lloyd St., Milwaukee 8, Wis.
Laurent, J. A., Jr., 2000 Grand Blvd., Schenectady, N. Y.
Lotz, W. E., Jr., Signal Corps, Fort Monmouth, Red Bank, N. J.

(Continued on page 45A)

MEMBERSHIP

(Continued from page 44A)

Myerhoff, A. A., 236 First St., Perth Amboy, N. J.
Minor, C. M., 5728 Blackstone Ave., Chicago 37, Ill.
Moon, R. F., U. S. Navy Electronics Laboratory, San Diego 52, Calif.
Nakahara, T., 1031 King St., Seattle 4, Wash.
Pickens, D. H., 8215 S. Vernon, Chicago, Ill.
Press, M., 137-77 - 70 Ave., Kew Garden Hills, L.I., N. Y.
Price, T. P., PUD 2, Raymond, Wash.
Quintin, W. P., Jr., 120 Lincoln Ave., Edgewood, Pa.
Richardson, L. D., 38 Hillcrest Rd., Reading, Mass.
Richter, F. A., 1031 E. 48 St., Chicago 15, Ill.
Rosar, M. T., 50 Moore Rd., Bronxville 8, N. Y.
Sample, G. A., 1123 Bellevue Ave., Syracuse, N. Y.
Santa, M. M., 805 W. 19 St., Wilmington, Del.
Saron, R. B., 345 E. 81 St., New York, N. Y.
Schrieber, W. F., 48 Logan St., Brooklyn 8, N. Y.
Schwartz, E. I., 1878 Harrison Ave., New York 63, N. Y.
Sinclair, R. O., Jr., Bell Telephone Laboratories, Whippany, N. J.
Spear, E. D., 316 B. South Columbia, Albuquerque, N. Mex.
Steele, H., 227 Hopkins St., Brooklyn 6, N. Y.
Stouffer, R. L., 630 N. Cedar, Galesburg, Ill.
Tobler, C. R., 20 Sutton Pl., Verona, N. J.
Viales, L. O., 252 16 Ave., San Francisco 18, Calif.
Vincent, R. C., O.R.L., Box 30, State College, Pa.
Weaver, P. F., Bell Telephone Laboratories, Inc., 463 West St., New York 14, N. Y.
Weinstein, D., 2726 Valentine Ave., New York 58, N. Y.
Weissman, S., 307 Throop Ave., Brooklyn 6, N. Y.
Williams, A. M., 817 Belleview St., Amarillo, Tex.
Wilmot, R. D., 303 First Ave., Ottawa, Ont., Canada
Willson, F. E., American Telephone & Telegraph Company, 7 East Clinton, Joliet, Ill.
Yeslin, A., 2739 Augusta Blvd., Chicago 22, Ill.
Yang, T., University of Toledo, Toledo 6, Ohio

News—New Products

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

Wavemeter and Modulation Indicator

The new Model 180 is a pocket-size control unit complete in itself which is being produced by the **Simpson Electric Co.**, 5216 W. Kinzie St., Chicago, Ill.

Designed for use by hams, ships, police radio, etc., it is an all-purpose instrument which makes it possible to accurately monitor quality of transmission, spot a transmitter at any point on the band desired, and keep a constant check on percentage modulation.

Encased in cast aluminum, the Model 180 wavemeter serves also to plot antenna field patterns, indicate changes in actual radiated power output, and search the region between bands for harmonics and parasitics. Separate coils for the 10-, 20-, 40-, and 80-meter bands and hand-drawn calibration curves are supplied. For coverage of all possible field-strength conditions, the 2-foot antenna can be plugged into the panel jack provided. Range of the instrument covers bands up to and including 420 Mc.

(Continued on page 46A)

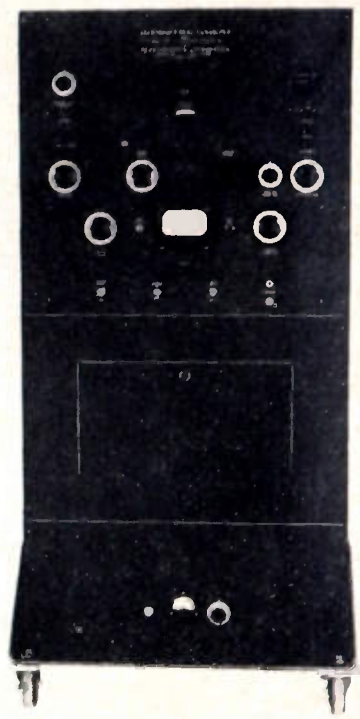
News—New Products

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

(Continued from page 45A)

New Test Equipment

Measurements Corporation, Boonton, N. J., has announced a new television standard signal generator, their Model #90. With a carrier range extending from 20 to 250 Mc, it is said to be the first such really commercial unit to reach the market. Engineering design features include a buffer stage and harmonic operation to provide separation of the output stages from the oscillator, so that the stability will not be disturbed through the coupling system. Modulation of the instrument extends to 5 Mc, assured through the use of band-pass over-coupled circuits. A crystal calibrator within the instrument permits accurate setting of the output signal frequency over the eight bands used to cover the wide range of generated signal. The video modulator provides a standard RMA composite signal, while the oscilloscope provided permits continuous monitoring of the output levels of the keyed dc potentiometer for observing the level of the signal delivered to the test under way. Dual coaxial output connectors are provided, with an attenuator of the balanced mutual-inductance type to afford smooth control of the delivered signal. The unit is mounted on casters for easy portability about the laboratory.



Kay Electric Co., Pine Brook, N. J., has announced that their Mega-Match is now available in a modified design to indicate reflected energy in a band width of 30 Mc anywhere in the spectrum between 10 and 500 Mc. In many applications it

(Continued on page 47A)

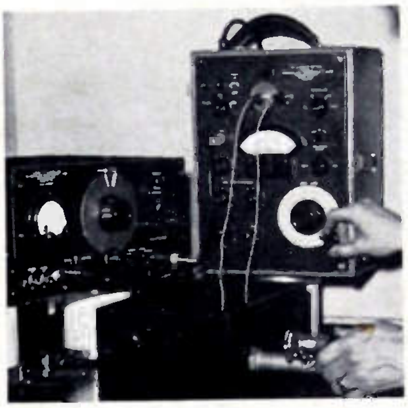
News—New Products

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

(Continued from page 46A)

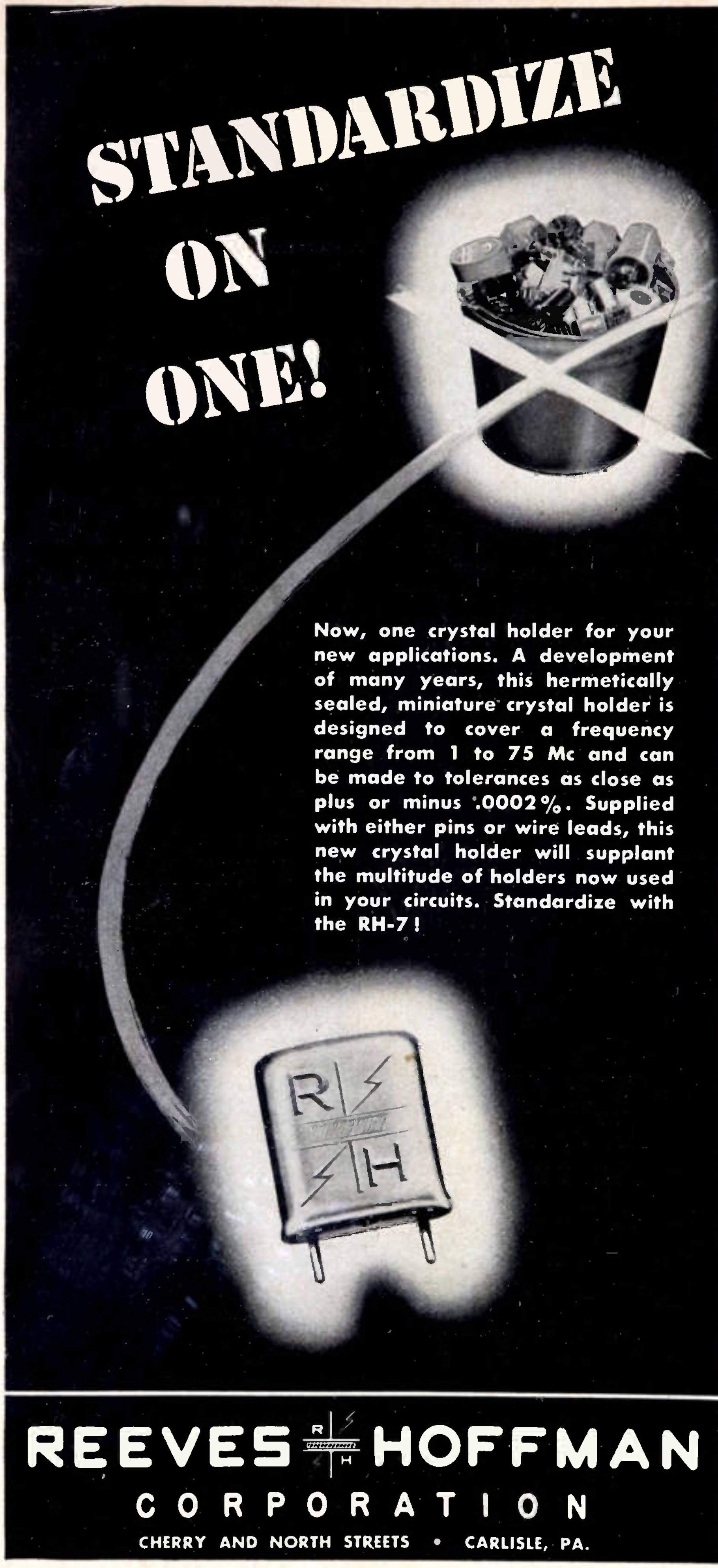
can be used to frequencies as high as 1,000 Mc. This is accomplished by a special coaxial detector. A high-quality relay line is included with the instrument.

Technology Instrument Corp., 1058 Main St., Waltham, Mass., brings forth a radio-frequency model of their renowned Z-angle meter, to make such tests as impedance, phase angle, etc., on antennas and transmission lines, and in general testing of radio frequency equipment. It makes such test possible without complex mathematics and the need for elaborate groupings of many test instruments. Those familiar with the audio-frequency model Z-angle meter will welcome an instrument specifically designed to make the same testing rapidly, at the radio end of the communication spectrum. Detailed technical data is available by addressing a written request to the makers.



DeMornay Budd, Inc., 475 Grand Concourse, New York, N. Y. have added to their line of waveguide apparatus with a new calorimeter to measure high orders of power in the microwave region. Although measuring average power as high as 500 watts at frequencies reaching 26,500 megacycles the calorimeter will make measurements starting at as low as 5 watts average power, and the accuracy of measurement increases with an increase in the amount of power measured. Accuracies as high as 2 watts in 500 watts are claimed possible for this unit at these extreme ranges and the oigh degree of engineering acumen built into the unit will be appreciated when it is realized that the instantaneous amounts of power can run to millions of watts. Measurement is accomplished by causing a water bath to absorb the entire rf power under test, this

(Continued on page 48A)

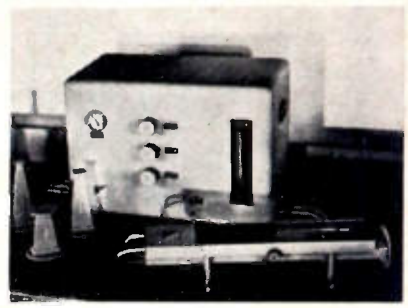


News—New Products

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

(Continued from page 47A)

water bath in turn then being cooled through the use of a circulating pump and connection to the external water supply. Accurate flow meters and needle valves enable measurement of the amount of heat sent into the water bath, recirculated in the system by a $\frac{1}{4}$ hp motor actuating a vane-type impeller pump. A higher-powered model is now in process of development.



Tracerlab, Inc., 55 Oliver St., Boston, Mass., has brought forth a new radiation testing meter of the ionization chamber variety. Battery-operated with a life rating of over 1,000 hours to a set of cells with three ranges calibrated on a 3″ meter and a pistol grip, the scope of the instrument is such that the range of beta or gamma rays measured and detected is well below the safe dosage value and well above the amount found in most radioactive laboratories today. The Model SU-1A as it is known is suitable for measurements of both radioisotopes as well as X-ray radiation as low as 50 kilovolts in origin.

Beta Electronics Co., 1762 Third Ave., New York, N. Y., have a new kilovoltmeter with f.s. ranges of 15/30 and 25/50 kilovolts, and 50 kilohms/volt and 25 kilohms/volt, respectively. Safety of operating personnel is a design feature incorporated in these instruments, the high-voltage test lead being equipped with a lucite handle of good length, making it possible to test circuits while the high voltage is applied thereto. The leads are

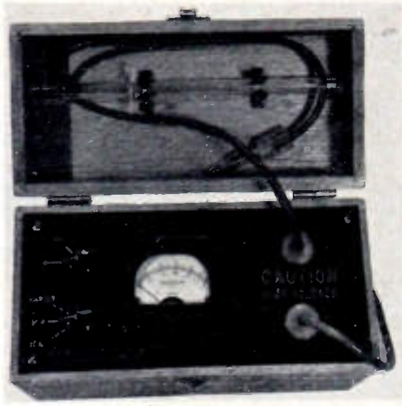
(Continued on page 49A)

News—New Products

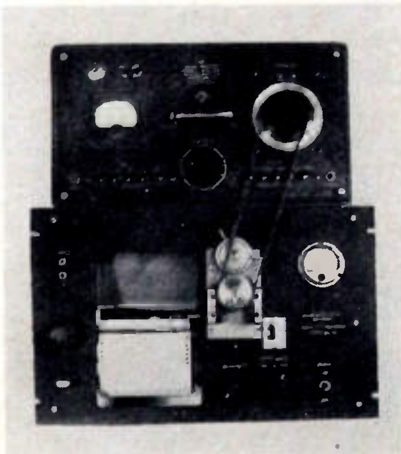
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 48A)

permanently attached to the instrument, so that there is no danger of contact with live wires at this potential. All switching takes place at low voltage, and the panels is protected from any lethal potentials. The input resistance is constant on both ranges, and the meter is a 20-micro-ampere full-scale instrument, for high sensitivity properly shunted to indicate the two ranges enumerated.



Sound Apparatus Co., 233 Broadway, New York, N. Y., has combined their recording sound analyzer with a General Radio Sound Analyzer Model 760-A to make a record on a 4" wide tape record. Complex waveforms are thus interpreted in a visible record with full scale ranges of 20, 40, 60 or 80 db range. The two instruments are mechanically linked to enable rapid analysis of the spectrum or complex waves as the unit eliminates the tedious point-by-point measurement technique, necessary without the use of a continuously recording ink recorder, for this type of laboratory study. As the instrument is portable, it lends itself ideally to field and laboratory use alike. Full data can be had by writing the makers.



Recent Catalogs

• • • A pictorial description in Catalog No. 900 of the complete line of clips and electrical specialties offered by **Mueller Electric Co.**, 1583 E. 31 St., Cleveland 14, Ohio.

(Continued on page 60A)

Positions Open for

PHYSICISTS
SR. ELECTRONIC ENGINEERS

Familiar with ultra high frequency and micro wave technique.

Experience with electronic digital and/or analog computer research and development program.

Salaries commensurate with experience and ability. Excellent opportunities for qualified personnel.

CONTACT:
C. G. Jones, Personnel Department
Goodyear Aircraft Corporation
Akron 15, Ohio

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . .

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.
1 East 79th St., New York 21, N.Y.

MATHEMATICIANS, ENGINEERS, PHYSICISTS

Men to train in oil exploration for operation of seismograph instruments, computing seismic data, and seismic surveying. Begining salary $250.00 to $300.00 per month depending upon background. Excellent opportunity for advancement determined by ingenuity and ability. The work requires changes of address each year; work indoors and out; general location in oil producing locations. Send resume and include snapshot to National Geophysical Co., Inc., 8800 Lemmon Ave., Dallas 9, Texas.

ELECTRONICS ENGINEER

Well known, 40 year old manufacturer of electrical and electronic instruments wants research engineer experienced in design of radio electronic apparatus at high frequencies, for the development of military and civilian test equipment. Box 542.

(Continued on page 52A)

ELECTRONICS ENGINEER

Senior engineer wanted for Engineering Department of large established, well-known firm, which designs and uses electronic instruments, counters and computing aids extensively for its own operations. Applicants must have degree in Electrical Engineering with Communications major. Graduate study highly desirable. Experience must include two or more years of research or development work in private industry. Location North side of Chicago. Salary open. In reply, please state age, give summary of education, experience and past employers.

Box #557
The Institute of Radio Engineers
1 East 79th St. New York 21, N.Y.

ENGINEERS – ELECTRONIC

Senior and Junior, outstanding opportunity, progressive company. Forward complete résumés giving education, experience and salary requirements to

Personnel Department
MELPAR, INC.
452 Swann Avenue
Alexandria, Virginia

WESTINGHOUSE RESEARCH

Openings in Pittsburgh

ENGINEER having three or more years experience in radar or television for work on television systems and television receiver research.

ENGINEER or PHYSICIST with experience in underwater acoustics.

PH.D. PHYSICIST with background for working on solid state problems.

ENGINEER or PHYSICIST for work on magnetic circuits and materials.

ELECTRONIC ENGINEER to work on advanced circuit design problems.

ENGINEER or PHYSICIST with three centimeter wave guide experience.

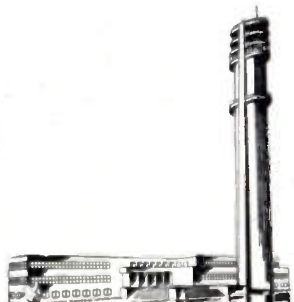
For application address Manager, Technical Employment, Westinghouse Electric Corporation, 306–4th Avenue, Pittsburgh, Pennsylvania

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.



RCA

"HIS MASTER'S VOICE"

RCA VICTOR

has excellent opportunities for Electronic Engineers and Physicists to do Design and Development work on:

- ✓ Radar Transmitters; Receivers and Indicators; Pulse Circuits; Microwave Antennas.
- ✓ Servo—Mechanisms and Controls.
- ✓ Audio Communications and Supersonics.
- ✓ Television Transmitters, Antennas, Studio Equipment.
- ✓ Mobile Communications Equipment.
- ✓ Aviation Communications and Electronic Equipment.

Write to: —

Employment Manager
Personnel Division
RCA Victor Division
Camden, New Jersey

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

POSITIONS OPEN

(Continued from page 52A)

ENGINEERS

A young rapidly growing mid-western research laboratory has openings for the following types of personnel:
(1) ELECTRONIC DESIGN ENGINEERS—Experienced circuit design engineers wanted for indicator timing circuit development.
(2) SERVO DESIGN ENGINEERS OR PHYSICISTS—Experience required for the development of airborne computor equipment.
Top pay offered to those capable of project responsibility. Local university offers graduate courses in servo and computer designed and applied mathematics. Send complete resume. Our Engineering Dept. knows of this announcement. Box 544.

ELECTRONIC ENGINEERS—PHYSICISTS

A leading electronics company in Los Angeles, California offers permanent employment to persons experienced in advanced research and development. State qualifications fully. Box 545.

ELECTRONIC ENGINEER

An opportunity for a man with considerable experience to head a small development and engineering group in a growing company located in Chicago. Pulse experience a necessity and a background of work with nuclear radiation instruments and nuclear detectors very desirable. Please give full details. Box 546.

(Continued on page 53A)

ZENITH RADIO CORPORATION
needs
Research and Development Engineers and Physicists

Project, senior and junior engineers and physicists are required for prosecution of several very interesting developments. Openings exist for men experienced in various aspects of radio and television receiver and transmitter development, in radar, and in applications of electronics to ordnance problems. *Salaries commensurate with experience and ability.*

WRITE: Director of Research
Zenith Radio Corp.
6001 W. Dickens Ave.
Chicago 39, Illinois

WANTED PHYSICISTS ENGINEERS

Engineering laboratory of precision instrument manufacturer has interesting opportunities for graduate engineers with research, design and/or development experience on radio communication systems, Servomechanisms (closed loop), electronic & mechanical aeronautical navigation instruments and ultra-high frequency & microwave technique.

WRITE FULL DETAILS
TO
EMPLOYMENT SECTION

SPERRY GYROSCOPE
COMPANY
DIVISION OF SPERRY CORP.
Marcus Ave. & Lakeville Rd.
Lake Success, L.I.

PHILCO

To maintain the Philco tradition of progressive research and development in the electronic field an ever increasing staff of engineers and physicists has been employed over the last two decades. Continuing expansion of Philco's engineering and research activities is producing excellent opportunities for engineers and physicists.

The scope of the work in the Philco laboratories includes basic research on the theory of semiconductors; vacuum tube research and design, including cathode ray tubes; and the design of special circuits, radio, television, television relay and radar systems.

IF YOU ARE INTERESTED IN YOUR OPPORTUNITY AT PHILCO,

WRITE . . . Engineering Personnel Director
Philco Corporation
Philadelphia 34, Pa.

POSITIONS OPEN

(Continued from page 52A)

RADIO AND TELEVISION ENGINEERS

The Industry Service Laboratory (Formerly License Laboratory), New York, has several positions open for Senior and Junior engineers having qualifications for development and consultation work in television and radio. Good technical education and some experience required. Interesting work, broadening experience, and wide contacts. Write fully to Director, Industry Service Lab., RCA Laboratory Division, 711 Fifth Ave., New York 22, N.Y.

CRYSTAL ENGINEER

Manufacturer of Piezo electric crystals desires experienced engineer familiar with quartz oscillating crystals and their applications to radio frequency control. Write full details. Box 548.

RESEARCH AND DEVELOPMENT ENGINEERS

Wanted for advanced research and development. Should have extensive experience on analysis of electronic systems in the fields of microwaves, missiles, radar, servomechanisms communications, navigational devices. Outstanding ability in E.E. or physics required. Please furnish complete resume, salary requirements and availability to: Personnel Manager, W. L. Maxson Corporation, 460 West 34th Street, New York, N.Y.

ELECTRICAL ENGINEER

Nationally known Chicago company is in need of high grade, experienced electrical engineer. Must be a college graduate with either a B.S. or E.E. degree, with high scholastic record. Should have from 2 to 5 years experience in electronics, preferably with a minimum of two years in the design of audio amplifiers. In reply give all particulars and state expected salary. Address Box 549.

DESIGN AND DEVELOPMENT ENGINEER

Design and development engineer to take care of engineering and development of receiving antennae and associated equipment. U.H.F. experience desirable. Upstate New York manufacturer. Reply giving age and qualifications to Box 450.

SALES ENGINEER

Several territories open east of Rocky Mountains, experienced sales engineer representative capable of selling and installing FM two-way radiotelephone systems for mobile operations. Nationally advertised product. Exceptional opportunity for right man. Send detailed qualifications, education, past experience and territory desired. Radiotelephone operators license, 1st or 2nd class preferred. Must have had previous experience in FM radiotelephone. Reply Box 451.

SCIENTISTS AND ENGINEERS

Wanted for interesting and professionally challenging research and advanced development in the fields of microwaves, radar, gyroscopes, servomechanisms, instrumentation, computers, and general electronics. Scientific or engineering degrees required. Salary commensu-

(Continued on page 54A)

Positions Open

(Continued from page 53A)

rate with experience and ability. Direct inquiry to Manager, Engineering Personnel, Bell Aircraft Corporation, P.O. Box 1, Buffalo 5, N.Y.

AERO DYNAMICIST ENGINEERS

Aero dynamicist engineers wanted to work on the design of Analog computers, to simulate the flight characteristics of specific aeroplanes. 3 years experience in stability and control essential. Knowledge of servomechanisms dynamics of free flight and applied mathematics desirable. Apply in person, or submit resume to Personnel Dept. Curtis Wright Corp., Propellor Division, Route 6, Caldwell Township, New Jersey.

ELECTRICAL ENGINEER

Opening for a man who has the ability to teach advanced electronic circuits and theory, frequency modulation and television. Should eventually teach course in ultra-high frequency techniques and electric wave phenomena and pulse systems. Reply to Director of Academic Administration, 1020 North Broadway, Milwaukee 2, Wisconsin.

SOUND ENGINEER

For large radio and television manufacturer in Chicago. Experience in loud speaker design, audio circuits, and acoustics necessary. Fine opportunity for advancement. Please write giving full particulars. Box 552.

PROFESSOR

Professor of communications engineering needed for fall 1949 by southeastern university. Will be in charge of graduate work and research activities. $6,000.00 for nine months with extra income for summer teaching. Must have Ph.D. or D.Sc. degree. Write Box 553.

ENGINEERS—PHYSICISTS

Southwestern state college engaged in research, development and operation in guided missile field has openings for graduate electrical engineers or physicists with experience in electro-mechanical devices, telemetering, audio, RF, VHF and antenna design. Salary depends on education and experience which should be fully described in first letter. Box 554.

COMMUNICATIONS—ELECTRONICS ENGINEER

Work will involve aeronautical communications and aircraft electronic projects. Some development work but primarily application engineering. Box 555.

ENGINEERS AND PHYSICISTS

An expanded program of research has created opportunities in the fields of acoustics, hydrodynamics, electronics, mechanics and mathematics. Positions flexible enough to permit part-time college teaching if desired. Opportunity for graduate study. Salary commensurate with experience. Appointments carry academic rank ranging from Research Assistant to Professor. Inquiry invited. P.O. Box 30, State College, Pa.

ELECTRICAL AND ELECTRONICS ENGINEERS

The Engineer Research and Development Laboratories of the Corps of Engineers, located at Fort Belvoir, Virginia, has a continuing need for electrical and electronics engineers to fill positions involving design and development and paying entrance salaries, depending on grade, of from $3727.20 to $6235.00 per annum.

Positions Open

(Continued from page 54A)

Applications should be mailed to the Civilian Personnel Office, Bldg. 211, Fort Belvoir, Virginia.

ELECTRICAL ENGINEERS

1 experienced engineer wanted for each of the following: radar pulse circuit; circuit development; radar modulators; radar devices; radar receivers; radar indicator circuit design and development. Send resume in confidence to Box 556.

ELECTRONICS ENGINEER MANAGER

Openings in equipment Development Department of Electronics Division, Boston, Mass. for graduate engineer with ability to administer and direct an engineering activity. An electrical engineer specialized in communications or electronics preferred, with industrial or equivalent in radar or communications systems design. Responsible for development and manufacture of microwave X & K band radar systems, FM test equipment, industrial and laboratory test equipment, and electronic computers. Boston or New York interviews will be arranged for qualified applicants submitting complete resume including salary requirements to Supervisor of Employment, Sylvania Electric Products, Inc. Industrial Relations Dept. 500 Fifth Ave., New York 18, N.Y.

MAGNETRONS DEVELOPMENT ENGINEER

An opening at our Electronics Division, Boston, Mass., for a Senior Engineer, to work in the tube development engineering group as a design and process engineer on magnetrons. Will be in charge of a project and have one or more engineers and technicians reporting to him. Will be responsible for manufacturing the tube after it has been designed in small quantities. B.S. in E.E. or physics required, and two to three years experience in the use of vacuum systems, and considerable experience on microwave type metal tube assembly. Furnish full particulars in writing regarding age, education, experience, present salary and salary requirements to: Supervisor of Employment, Sylvania Products, Inc. Industrial Relations Dept. 500 Fifth Ave., New York 18, N.Y.

ELECTRONICS ENGINEER SECTION HEAD

Our Electronics Division, Boston, Mass., is seeking a graduate electrical engineer (S.B. or S.M. electronics option preferred) to supervise a group of 17 engineers and several technicians working on X & K band radar and other electronic equipment development, including computers. Early interviews will be granted qualified applicants with actual experience in design of large scale electronics equipment and administration of engineers. Mail complete resume to Supervisor of Employment, Sylvania Electric Products, Inc. Industrial Relations Dept. 500 Fifth Ave., New York 18, N.Y.

ELECTRONICS ENGINEERS

Top flight engineers. Must have 10 years design and development experience on servomechanisms and amplifiers, circuits and equipment layout. Apply in person or submit complete resume to Personnel Dept., Curtis Wright Corp., Propeller Division, Route 6, Caldwell Township, New Jersey.

INFORMATION SERVICE

I.R.E. Yearbook and I.R.E. Industry Research Division

A two-fold service is available without cost to firms supplying products for or rendering technical services to the radio-and-electronic industry. (1) The I.R.E. Yearbook, which is noted for its reference value, lists more than 2,000 firms and furnishes product or industry classification. (2) A classified index of products and services is maintained by the I.R.E. Industry Research Division. Radio engineers continuously draw upon this bank of statistical data.

These two sources of information are not only invaluable to I.R.E. members and other engineers, but the proper listing of your company may bring you new business.

To help keep up to date, please check-off your products and services. These may be rewritten on the coupon on next page and sent to: Industry Research Div., Proceedings of the I.R.E. Room 707, 303 West 42nd St., New York 18, N.Y.

Products to Be Checked By Radio-Electronic Manufacturers

1. AMPLIFIERS, Audio Frequency.
( ) a. Broadcast Speech Input Equipment.
( ) b. Dynamic Noise Suppressors.
( ) c. High Fidelity.
( ) d. Inter-communication Systems.
( ) e. Medical Equipment.
( ) f. Peak Limiting.
( ) g. Phonograph Preamplifiers-equalized.
( ) h. Power Amplifiers.
( ) i. Pre-amplifiers.
( ) j. Public Address Systems.
( ) k. Recording Amplifiers.

2. ANTENNAS.
( ) a. AM Broadcast.
( ) b. Dummy.
( ) c. FM Broadcast.
( ) d. Miscellaneous.
( ) e. Receiving Types, all services.
( ) f. Television Broadcast.

3. ANTENNA ACCESSORIES.
( ) a. Feeder Systems.
( ) b. Insulators.
( ) c. Phasing & Tuning Equipment.
( ) d. Support Towers.
( ) e. Tower Lighting Equipment.

4. ATTENUATORS.
( ) a. Audio Frequency.
( ) b. Radio Frequency.

5. BATTERIES.
( ) a. Flashlight & Miscellaneous Dry.
( ) b. Hearing Aid.
( ) c. Portable Radio Types.
( ) d. Storage.

6. BLOWERS & COOLING FANS.
Bridges, see 60a.

7. CABLE & WIRE.
( ) a. Co-axial Cable.
( ) b. Pre-formed Harnesses
( ) c. Rubber Insulated Wires.
( ) d. Shielded Types.
( ) e. Synthetic Insulated Wires.
( ) f. U.H.F. Types.

8. CAPACITORS: Fixed.
( ) a. Ceramic.
( ) b. Electrolytic.
( ) c. Mica.
( ) d. Oil Filled.
( ) e. Paper.
( ) f. Pressurized Gas.
( ) g. Vacuum.

9. CAPACITORS: Variable.
( ) a. Neutralizing.
( ) b. Precision.
( ) c. Temperature Frequency Compensating.
( ) d. Trimmers.
( ) e. Tuning.
( ) f. Vacuum.

10. CERAMICS.
( ) a. Coil Forms.
( ) b. Custom Fabrication.
( ) c. Rods.
( ) d. Sheets.

11. CHASSIS & RELAY RACK CABINETS: Metal.
( ) a. Open Stock.
( ) b. Special Order & Custom Fabrication.
Coil Forms, see 10a.

12. COILS.
( ) a. A. F. Chokes.
( ) b. Miscellaneous.
( ) c. R. F. Chokes.
( ) d. Toroids.
( ) e. Transformer Coils.
( ) f. Tuning.
Condensers, see 8 & 9.

13. CONNECTORS.
( ) a. AN Standard Types.
( ) b. Microphon.
( ) c. Power.
Consoles, see 1a.

14. CONVERTERS.
( ) a. Frequency.
( ) b. Vibrator.
Rotary, see 38.

15. CORES & CORE MATERIALS.
( ) a. Complete Cores.
( ) b. Laminations.
( ) c. Powdered Metal.

16. CRYSTALS.
( ) a. Germanium & Silicon, etc.
( ) b. Oscillating Quartz.
( ) c. Piezo-Electric.

17. CRYSTAL HOLDERS.
Discs, Recording, see 51a.

18. DRAFTING EQUIPMENT & SUPPLIES.
Dynamotors, see 38a.

19. ELECTRONIC CONTROL EQUIPMENT.
( ) a. Air Conditioning Controls.
( ) b. Burglar Alarm & Protection Devices.
( ) c. Combustion & Smoke Elimination.
( ) d. Fire Prevention & Detection.
( ) e. Production Controls, Counting & Sorting.
( ) f. Variable Speed Regulators.
( ) g. Voltage Control & Stabilization.

20. EQUALIZERS.
( ) a. Dialogue.
( ) b. Line.
( ) c. Magnetic Reproducer Types.
( ) d. Sound Effects.

21. FACSIMILE EQUIPMENT.

22. FILTERS.
( ) a. Band Pass & Band Rejection.
( ) b. Dividing Networks.
( ) c. Noise Elimination.
Sound effects, see 20d.
Frequency Measuring Equipment, see 37a, 59a, 61b, c, d, and e

23. FUSES & FUSE HOLDERS.
Generators, see 38c.

24. GRAPHIC RECORDERS.

25. HARDWARE & MANUFACTURING AIDS.

26. INDUCTION HEATING EQUIPMENT.
( ) a. Manufacturing Processes.
( ) b. Medical Applications.
Inductors, see 12.

27. INSULATION.
( ) a. Cloth.
( ) b. Glass.
( ) c. Mica.
( ) d. Paper.
( ) e. Varnished Cambric.
See also 10 & 45.

28. JACKS, JACK FIELDS, PLUGS, & PATCH CORDS.

29. KEYS.
( ) a. Switching.
( ) b. Telegraph.
Knobs, see 40c.

30. LACQUERS.
( ) a. Finishing.
( ) b. Fungus Proofing.
( ) c. Moisture Proofing.

31. LOUDSPEAKERS & HEADPHONES.
( ) a. Commercial Grade Loudspeakers.
( ) b. Headphones.
( ) c. High Fidelity Loudspeaking Systems.
( ) d. High Frequency Types.
( ) e. Low Frequency Types.

32. MACHINERY, FIXTURES, & TOOLS FOR RADIO-ELECTRONIC MANUFACTURING.

33. MAGNETS.
( ) a. Electro.
( ) b. Permanent.
Measuring Equipment, see 35, 59, 60, 61.

34. METALS: Base.
( ) a. Copper.
( ) b. Ferrous.
( ) c. Non-ferrous, excluding Copper.
( ) d. Powdered.
( ) e. Precious & Rare.

35. METERS.
( ) a. Ammeters.
( ) b. Elapsed Time.
( ) c. Frequency Indicating.
( ) d. Voltmeters.
( ) e. Volume Level Meters (dB & VU).
( ) f. Wattmeters & Watt Hour Meters.
Vacuum Tube Voltmeters, see 60i.

36. MICROPHONES.
( ) a. Carbon.
( ) b. Condenser.
( ) c. Crystal.
( ) d. Magnetic.

37. MONITORING EQUIPMENT.
( ) a. Frequency.
( ) b. Modulation.

38. MOTOR GENERATORS.
( ) a. Dynamotors.
( ) b. Frequency Changers.
( ) c. Motor-Generators.
( ) d. Rotary Converters.

39. MOTORS: Very Small.
( ) e. Special Fabrication.
( ) a. Blower Motors.
( ) b. Selsyn Controls.
( ) c. Timing Devices.

40. MOULDED PRODUCTS & SERVICES.
( ) a. Cabinets.
( ) b. Insulators.
( ) c. Knobs & Parts.
( ) d. Proprietary Mouldings.

41. OPTICAL SYSTEMS, MIRRORS, SCREENS, & ACCESSORIES.
Oscillators, see 59a, d & 61b, c, d, e.

42. OSCILLOGRAPHS & ACCESSORIES.
( ) a. General Purpose, Cathode Ray.
( ) b. Recording.
( ) c. Synchroscopes, Cathode Ray.
( ) d. U.H.F. Cathode Ray Equipment.
Panels, see 11.

43. PHONOGRAPH & TRANSCRIPTION REPRODUCING EQUIPMENT.
( ) a. Crystal Pick-ups.
( ) b. Magnetic Pick-ups.
( ) c. Phonograph Motors.
( ) d. Playback arms.
( ) e. Record Changers.
( ) f. Turntables, complete.
Pre-amplifiers, see 1i.

44. PILOT LIGHTS & ASSEMBLIES.
( ) a. Incandescent.
( ) b. Neon.

45. PLASTICS.
( ) a. Raw Powders for Moulding.
( ) b. Rods.
( ) c. Sheets.
Plugs, see 13 & 28.

46. POINT TO POINT COMMUNICATION EQUIPMENT.
( ) a. Aircraft & Airport Equipment.
( ) b. Citizen Radio.
( ) c. Emergency Communications.
( ) d. Fleet Dispatching.
( ) e. Police & Fire Department Equipment.
( ) f. Ship to Shore Equipment.

47. POWER SUPPLIES.
( ) a. Electrically Powered.
( ) b. Gasoline Driven.
( ) c. Voltage Regulated Output types.
Printed circuits, see 25.

48. PUMPS: Vacuum.
Racks, see 11.
Radar Equipment, see 64b.

49. RADIO RECEIVERS
( ) a. Broadcast.
( ) b. Communications.
( ) c. Fixed Frequency.
( ) d. Frequency Modulation.
( ) e. Special Purpose.
( ) f. Television.

50. RECORDING EQUIPMENT.
( ) a. Disc Recording Machines.
( ) b. Magnetic Tape Recorders.
( ) c. Magnetic Wire Recorders.

51. RECORDING ACCESSORIES & SUPPLIES.
( ) a. Blanks.
( ) b. Cutting Needles.
( ) c. Disc Recording Heads.
( ) d. Magnetic Recording Playback, & Biasing Heads.
( ) e. Magnetic Recording, Tape.
( ) f. Magnetic Recording Wire.

52. RECTIFIERS.
( ) a. Metallic.
( ) b. Vacuum Tube.
Regulators, Voltage, see 67.

53. RELAYS.
( ) a. Hermetically sealed.
( ) b. Instrument.
( ) c. Keying.
( ) d. Mercury.
( ) e. Power Control & Overload.
( ) f. Stepping.
( ) g. Telephone Types.
( ) h. Time Delay.

54. REMOTE CONTROLLING EQUIPMENT.
( ) a. Automatic Tuning Mechanisms.
( ) b. Remote Controls.
( ) c. Switching Functions.
( ) d. Servo-Mechanisms.

55. RESISTORS.
( ) a. Carbon fixed.
( ) b. Carbon Variable.
( ) c. Precision.
( ) d. Vacuum Sealed.
( ) e. Wire Wound, Fixed.
( ) f. Wire Wound, Variable.

56. SOCKETS, VACUUM TUBE.
( ) a. Receiving Tube Types.
( ) b. Transmitting Tube Types
( ) c. Underwriters Laboratories Approved Types.

57. SOLDER.
( ) a. Acid Cored.
( ) b. Plain.
( ) c. Precious Metal.
( ) d. Pre-forms.
( ) e. Rosin Cored.
Speakers, see 31.

58. SWITCHES.
( ) a. Band Switches.
( ) b. Circuit Breaking.
( ) c. Key.
( ) d. Mercury Switches.
( ) e. Momentary Contact.
( ) f. Power.
( ) g. Precision Snap-Acting.
( ) h. Rotary
( ) i. Time Delay.
( ) j. Toggle & Push Button.

59. TESTING & MEASURING EQUIPMENT: Audio Frequency.
( ) a. Beat Frequency Oscillators.
( ) b. Distortion & Noise Analyzers.
( ) c. Intermodulation Distortion Analyzers.
( ) d. Resistance Capacity Oscillators.
( ) e. Square Wave Generators.
( ) f. Wave Form Analysis Equipment.

60. TESTING & MEASURING EQUIPMENT: General.
( ) a. Bridges, all types.
( ) b. Capacitance Decades
( ) c. Capacitor Testers.
( ) d. Multi-meters.
( ) e. Resistance Decades.
( ) f. Resistor Testers-Ohmmeters.
( ) g. Stroboscopes.
( ) h. Tube Testers.
( ) i. Vacuum Tube Voltmeters.
( ) j. Vibration Testing Equipment.

61. TESTING & MEASURING EQUIPMENT: Radio Frequency.
( ) a. "Q" Meters.
( ) b. Signal Generators, AM.
( ) c. Signal Generators, FM.
( ) d. Signal Generators, TV.
( ) e. Standard Frequency Generators & Multi-Vibrators.
( ) f. Sweep Generators & Calibrators.
( ) g. Wavemeters.

62. TRANSFORMERS.
( ) a. Audio Frequency.
( ) b. Hermetically Sealed.
( ) c. High Fidelity Audio Types.
( ) d. Power Components.
( ) e. Pulse Generating Types.
( ) f. Radio Frequency.
( ) g. Voltage Regulating.

63. TRANSMITTERS.
( ) a. AM Broadcast.
( ) b. Communications.
( ) c. FM Broadcast.
( ) d. Special Types.
( ) e. TV Broadcast.

64. ULTRA HIGH FREQUENCY EQUIPMENT & ACCESSORIES.
( ) a. Antennas & Reflectors.
( ) b. Radar Apparatus
( ) c. Receivers.
( ) d. Transmitters.
( ) e. Waveguides.

65. VACUUM TUBES.
( ) a. Cathode Ray.
( ) b. Geiger-Mueller.
( ) c. Industrial types.
( ) d. Klystrons & Magnatrons.
( ) e. Phototubes.
( ) f. Pirani Tubes.
( ) g. Receiving Types.
( ) h. Rectifiers.
( ) i. Special Purpose.
( ) j. Thyratrons.
( ) k. Television Types.
( ) l. Transmitting Types.
( ) m. Voltage Regulators.

66. VACUUM TUBE COMPONENT PARTS.
Varnishes, see 30.

67. VOLTAGE REGULATORS.
( ) a. Automatic.
( ) b. Manually Controlled

68. WAXES, POTTING & SEALING COMPOUNDS

69. WOODEN CABINETS.
( ) a. Radio Sets.
( ) b. Record Storage.

NON-MANUFACTURING CLASSIFICATIONS.

91. BOOKS AND BOOK PUBLISHERS.

92. BROADCASTING STATIONS & COMMUNICATION COMPANIES.

93. CONSULTING ENGINEERS.
( ) a. Acoustical.
( ) b. Electrical.
( ) c. Mechanical.
( ) d. Radio.

94. DISTRIBUTORS & JOBBERS OF RADIO & ELECTRONIC EQUIPMENT & SUPPLIES.

95. FREQUENCY MEASURING & MONITORING SERVICES.

96. LABORATORIES & CUSTOM BUILDERS OF EQUIPMENT.

97. RECORDING STUDIOS & SERVICES.

98. TECHNICAL SCHOOLS & INSTITUTIONS.

99. TRANSCRIPTION LIBRARIES.

COUPON

Send to: Industry Research Div., Proceedings of the I.R.E.
Rm. 707, 303 West 42nd St., New York 18, N.Y.

Date ....

Firm Name: ....

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The following is a list of code numbers and letters for our products or services:

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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 announcemen twithout assignment of reason.

ENGINEER

S.B.E.E., M.I.T. June 1948, major in electronics, minor in servomechanisms. Mechanical engineering Ohio State University in Army S.T.P., major in internal combustion engines. Experience: 6 months machine shop, 1½ years inspector and designer in precision bearing manufacturing. 1½ years industrial engineering in Manhattan project. Age 27. Single. Prefer development work but would like to hear from any firm for which I can be an asset. Box 189 W.

ELECTRONIC ENGINEER

10 years experience in research, design, development and supervision with automotive and aircraft fields and guided missile projects. Seeking administrative position. Aggressive; personable. Box 188 W.

RADIO ENGINEER

English radio engineer seeks position in communications. Fully qualified by examination. 12 years experience. Leslie F. Bennett, c/o Mercantile Trust Bank, Baltimore, Md.

JUNIOR ENGINEER

Electrical engineer graduate. Age 24. Single. B.S.E.E. February 1948. Some experience in television field. Seeks interesting position with good company in New York City area. Box 190 W.

TELEVISION ENGINEER

Graduated American Television Institute of Technology November 1948 with B.S.T.E. Age 26. Married. 1st class F.C.C. License. 4 years maintenance Navy radio equipment. Trained in operation and maintenance of R.C.A. Image Orthicon and DuMont equipment. Desires position in television broadcasting field. Box 192 W.

ENGINEER

Graduated University of Michigan August 1948 with B.S.E.E. in communications. Age 24. Married. Two years Army radar (G.C.A.). Two years shop experience. Interested in sales or development engineering. Prefer midwest area. Box 194 W.

JUNIOR ENGINEER

Graduate R.C.A. Institutes. Age 27. Married, 1 child. Desires work west coast

(Continued on page 59A)

Positions Wanted

(Continued from page 59A)

in television, radio, electronics research or development. Limited Air Force Experience. Ambitious, persevering. Box 195 W.

TECHNICAL EXECUTIVE

Engineering physicist, M.A., Gold medallist, 30, experienced in radio, radar and X-ray, Division manager Montreal branch of European concern wants change to progressive North American concern that can use his talents not necessarily in these fields. Box 203 W.

ELECTRONIC ENGINEER

Currently engaged in production, design and development work for capacitor manufacturer, desires position in electronic industry within Chicago area. B.S.E.E. January 1948 Illinois Institute of Technology. Communications major. Age 25. Married. Two years Navy electronics experience including supervision of Radio Teletype station. Box 204 W.

ELECTRONIC ENGINEER

B.S.E.E. January 1949. College of the City of New York. Age 23. Single. 2 years experience as Navy electronic technician. Location immaterial if near graduate school. Box 214 W.

RADIO ENGINEER

B.S.E.E. June 1949, University of Cincinnati. Age 31. Married. Experience: 2½ Co-op in Quality control and television production; 9 months advanced development high frequency guided missile electronic equipment; 2½ years Army as Communications Chief in Air Forces; 8 years in managerial and executive field. Desires position as research or development engineer with a progressive firm in the eastern or central states. Member of Tau Beta Pi and Eta Kappa Nu. Box 215 W.

TELEVISION ENGINEER

Graduating American Television Institute of Technology, January 1949 with B.S.T.E. Age 28. Married, 1 child. Past supervisory experience. Desires sales or development engineering. Box 216 W.

MICROWAVE ENGINEER

Experienced in wave guide and antenna work; seeks responsible position with a future. New York City or Long Island. Box 217 W.

ELECTRONIC ENGINEER

M.S.E. February 1949; B.S.E. June 1948 University of Michigan. Interested in development, non-basic research in electronics and communications. Prefer New York or New Jersey area. Ex-Navy technician, summer development job experience. Age 23. Single. Box 218 W.

ELECTRICAL ENGINEER

Advanced experience in development and production of magnetic amplifiers for all applications. Vast experience in electronic, control and magnetic devices fields. Will accept position with good opportunity. Box 219 W.

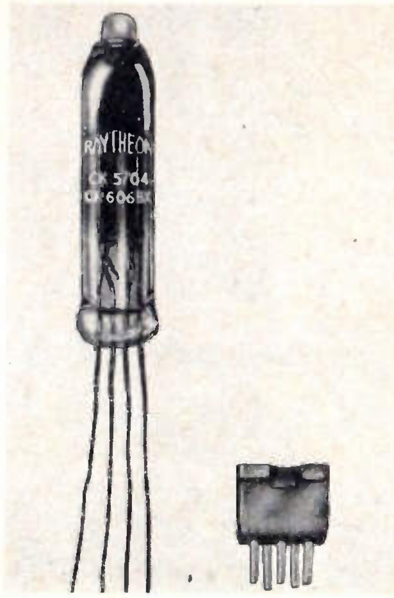
News—New Products

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

(Continued from page 49A)

New Tubes

Raytheon Manufacturing Co., 60 E. 42 St., New York, N. Y. widens the scope of their line of sub-miniature tubes with the CK606BX/CK5704, a diode of such small dimensions that its natural resonance is



well above 1,200 Mc. It is similar to one-half of a 6AL5 dual diode of the more conventional-appearing tubes. With a 6.3-volt 150 ma heater, the tube lends itself to rf probes for vtvm applications, or in ufh applications where small size as well as ability to operate at microwave ranges is called for.

Eclipse-Pioneer Division, **Bendix Aviation Corp.**, Teterboro, N. J., has announced a TT-1 ultra-high-frequency diode for television receiver testing, where it is desired to observe sensitivity and to determine the operating range at which it will perform, satisfactorily. The new diode is also applicable to radar receiver sensitivity measurements.

This noise diode was originally developed by RCA, and Bendix is currently making it under license from that organization. It is a noise-generating unit of very wide-band characteristics, so that the sensitivity of the receiver can be ascertained by determining what level above inherent receiving set noise will be needed for producing satisfactory aural and video signals. Laboratory models are now in process of production and the maker will answer all inquiries about their procurement.

(Continued on page 61A)

News—New Products

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

(Continued from page 60A)

Model TAA-16 Amplifier

The TAA-16, an amplifier useful in the determination of standing-wave voltage ratios when used in connection with square-law detector probes and slotted wave-guides, is available from **Browning Laboratories, Inc.**, Winchester, Mass.

This amplifier has two inputs, selectable by switch for rapid comparison. Operating frequencies are from 500 to 5000 cps, and operation can be wide band or highly selective as needed. The selective network is panel-tuned through the range noted above. The output meter is calibrated directly in standing-wave voltage ratio. Full-scale meter readings are obtained with 15 microvolts at the input under wide band operation, while a 10-microvolt signal will produce this result under selective operation.

The shielding employed permits full gain to be used without disturbance or the 60-cps harmonics generated in the unit.

Description: the TAA-16 is supplied in a cabinet and is suitable for rack mounting; height 9″, width 20″, depth 12″, weight 30½ lbs.

Recent Catalogs

Clarkstan Corp., 11927 W. Pico Blvd., Los Angeles 34, Calif., has issued a new composite catalog describing recent additions to their line of audio-frequency products and associated items. Included are data sheets on their pickup, a sweep-frequency generator, in addition to their sweep frequency record, a pickup scale, and similar items that the engineer particularly interested in audio items will want to have. A copy will be sent upon written request to the company, addressing them with your corporate connection.

Kay Electric Co., 14 Maple Ave., Pine Brook, N. J., has issued bulletins on their products for video testing, each describing in detail the uses to which the unit can be put. These items find application in video set testing and field servicing, and the catalogs and reprints are offered to all in the field who are working in this branch of the art.

National Carbon Co., 60 E. 42 St., New York 17, N. Y., has a booklet listing the carbon brushes of its manufacture. This publication details the methods needed to standardize on types of brushes used in rotating machinery, and the need for such a step, to save large inventories, increase field servicing of units in operation, etc. Ask for Catalog Section B-2106 when writing to the company.

(Continued on page 64A)

PROFESSIONAL CARDS

INDEX AND DISPLAY ADVERTISERS

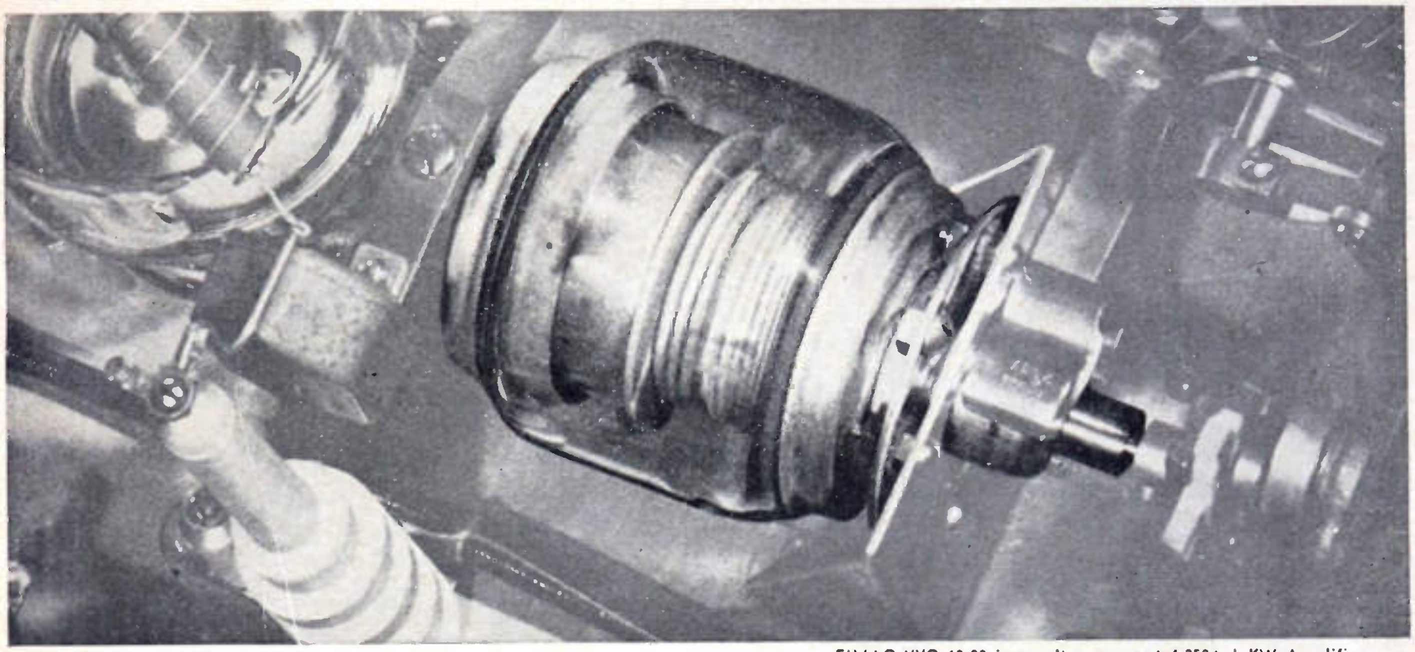
DISPLAY ADVERTISERS

Follow the Leaders to

Eimac Tubes

The Power for R-F

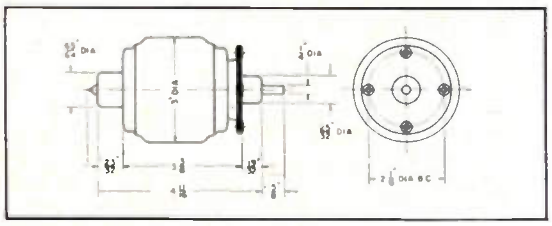
ESSENTIAL IN MODERN CIRCUITRY



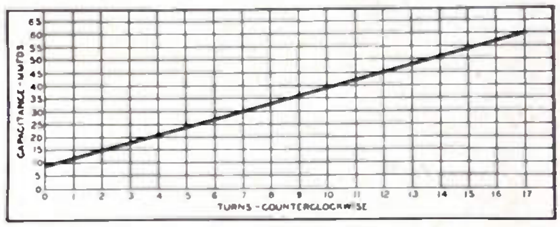
EIMAC VVC 60-20 in an ultra-compact 4-250A I KW Amplifier.

Consider the advantages . . . and Eimac Variable Vacuum Capacitors become the essential component in modern circuitry.

- Extremely compact size reduces equipment bulk. Type VVC 60-20 is less than one-sixth the size of air-dielectric capacitors with similar ratings.



- Structural rigidity eliminates electro-mechanical vibration.
- Low-torque tuning mechanism.
- Unaffected by dusty or humid atmospheres. Ideal for industrial application.
- Capacitance variation is linear with shaft rotation.



- Low temperature coefficient. Negligible change in capacitance due to temperature variance. (.004 mmfd. per degree cent.)

Eimac variable vacuum capacitors are immediately available. In addition to the type VVC 60-20 illustrated here, there are types VVC2 60-20 and VVC4 60-20.

GENERAL CHARACTERISTICS

	Capacity	R-F Peak Voltage	Maximum RMS Current
VVC 60-20	10-60 mmf.	20-KV	40 amp.
VVC2-60-20			
Parallel	20-120 mmf.	20-KV	80 amp.
Split-stator	5-30 mmf.	40-KV	40 amp.
VVC4-60-20			
Parallel	40-240 mmf.	20-KV	160 amp.
Split-stator	10-60 mmf.	40-KV	80 amp.

EITEL-McCULLOUGH, INC.

206 San Mateo Ave., San Bruno, California

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

for MORE POWER

The Type V-20 VARIAC* . . . one of the newer additions to our VARIAC line . . . is proving exceedingly popular. With its load rating of almost 3½ kw it is filling a definite need which could not be met with a single Type 100 which it replaces. The V-20 is rated at 20 amperes, with a 30 ampere maximum; the 230-volt V-20H at 8 and 10 amperes respectively.

These latest VARIACS are provided with heavier barrier terminals in a box equipped with knockouts for standard ¾-inch BX or conduit. The new combination knob and hand wheel makes it much easier to vary voltage. The V-20 dials have extra large calibration figures, easy to read at a distance.

As with all VARIACS, output voltages are continuously adjustable from ZERO to 17% above line voltage.

SPECIFICATIONS

	TYPE V-20	TYPE V-20H
LOAD RATING (KVA)	3.45	2.3
Input Voltage	115	230 or 115
Output Voltage Zero to	115 or 135	270 or 230
Rated Current (Amperes)	20	8
Maximum Current (Amperes)	30	10
PRICE	$55.00	$55.00

WRITE FOR THE "VARIAC BULLETIN"

* The trade name VARIAC is registered at the U. S. Patent Office. VARIACS are manufactured and sold under Patent No. 2,009,013.

GENERAL RADIO COMPANY Cambridge 39, Massachusetts

90 West St., New York 6 920 S. Michigan Ave., Chicago 5 1000 N. Seward St., Los Angeles 38