Proceedings of the I·R·E

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

December, 1947

Volume 35 Number 12

General Electric Company

GARGANTUAN TUBE-TEST RACK

Thirty-two modern cathode-ray tubes, for television reception, simultaneously endure thousands of hours of steady overload at voltages from 1200 to 60,000, interrupted only by their brief removal during special periods devoted to measurements.

1948 I.R.E. NATIONAL CONVENTION—MARCH 22-25

PROCEEDINGS OF THE I.R.E.

Waves and Electrons Section

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The Institute of Radio Engineers

LS SERIES

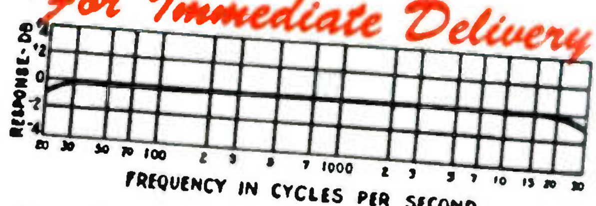
The Ultimate in Quality

UTC Linear Standard Audio Transformers represent the closest approach to the ideal component from the standpoint of uniform frequency response, low wave form distortion, high efficiency, thorough shielding and utmost dependability. Wartime restrictions having been lifted, and UTC production running at full capacity, we now offer these transformers for immediate delivery.

UTC Linear Standard Transformers feature...

- **True Hum Balancing Coil Structure** . . . maximum neutralization of stray fields.
- **Balanced Variable Impedance Line** . . . permits highest fidelity on every tap of a universal unit . . . no line reflections or transverse couplings.
- **Reversible Mounting** . . . permits above chassis or sub-chassis wiring.
- **Alloy Shields** . . . maximum shielding from induction pickup.
- **Multiple Coil, Semi-Toroidal Coil Structure** . . . minimum distributed capacity and leakage reactance.
- **Precision Winding** . . . accuracy of winding .1%, perfect balance of inductance and capacity; exact impedance reflection.
- **Hiperm-Alloy** . . . a stable, high permeability nickel-iron core material.
- **High Fidelity** . . . UTC Linear Standard Transformers are the only audio units with a guaranteed uniform response of ± 1.5DB from 20-20,000 cycles.

For Immediate Delivery



Typical Curve for LS Series

Type No.	Application	Primary Impedance	Secondary Impedance	Max. Level	Relative hum-pickup reduction	Max. unbalanced DC in primary	List Price
LS-10	Low impedance mike, pick-up or multiple line to grid.	50, 125, 200, 250, 333, 500/600 ohms	60,000 ohms in two sections	+15 DB	−74 DB	5 MA	$25.00
LS-10X	As above	As above	50,000 ohms	+14 DB	−92 DB-Q	5 MA	$32.00
LS-21	Single plate to push pull grids Split primary and secondary	8,000 to 15,000 ohms	135,000 ohms; turn ratio 3:1 overall	+14 DB	−74 DB	0 MA	$24.00
LS-30	Mixing, low impedance mike, pickup, or multiple line to multiple line	50, 125, 200, 250, 333, 500/600 ohms	50, 125, 200, 250 333, 500/600 ohms	+17 DB	−74 DB	5 MA	$25.00
LS-30X	As above	As above	As above	+15 DB	−92 DB-Q	3 MA	$32.00
LS-50	Single plate to multiple line	8,000 to 15,000 ohms	50, 125, 200, 250, 333, 500/600 ohms	+17 DB	−74 DB	0 MA	$24.00
LS-55	Push pull 2A3's, 6A5G's, 300A's, 275A's, 6A3's, 6L6's	5,000 ohms plate to plate and 3,000 ohms plate to plate	500, 333, 250, 200, 125, 50, 30, 20, 15, 10, 7.5, 5, 2.5, 1.2	20 watts			$28.00
LS-57	Same as above	5,000 ohms plate to plate and 3,000 ohms plate to plate	30, 20, 15, 10, 7.5, 5, 2.5, 1.2	20 watts			$20.00

The above listing includes only a few of the many units of the LS Series. For complete listing — write for catalogue.

United Transformer Corp.

150 VARICK STREET NEW YORK 13, N. Y.

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

Proceedings of the I.R.E., December, 1947, Vol. 35, No. 12. Published monthly in two sections by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price $1.50 per copy. Subscriptions: United States and Canada, $12.00 a year; foreign countries $13.00 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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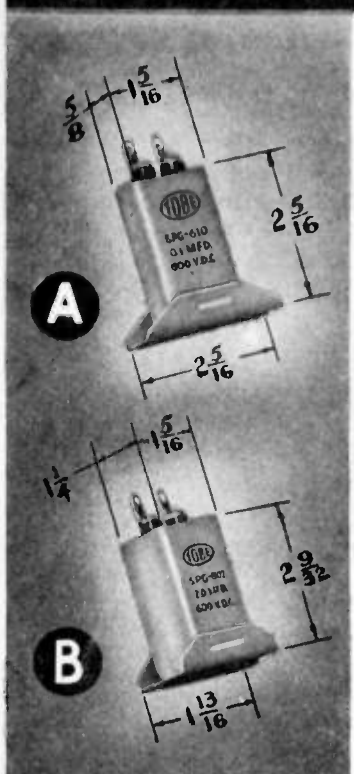
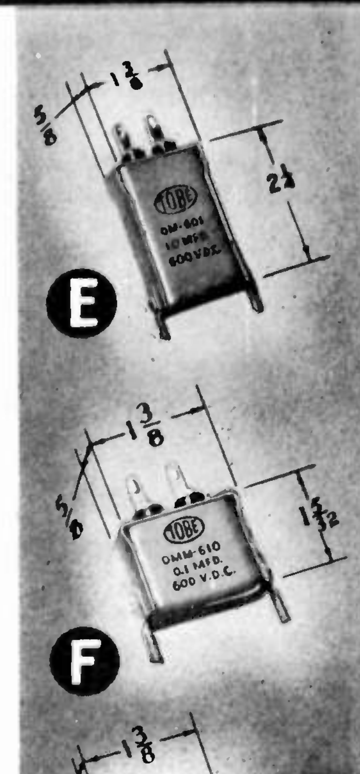
CHECK *these* SPECIFICATIONS





THESE data explain the outstanding performance of Tobe "Oil-Mites" . . . demonstrate their qualifications for use under extreme humidity and temperature environment . . . show the diversity of mounting provisions, sizes, housings, and electrical ratings for convenient incorporation in electronic and electrical apparatus.

Winding: non-inductive.
Impregnation: mineral oil.
Case: seamless drawn steel, hermetically sealed; non-magnetic case (copper or brass) can be furnished.
Terminals: non-removable tinned copper solder lugs riveted to phenolic bushings.
Terminal Seal: oilproof gaskets between all adjacent surfaces in terminal assembly; terminal solder-sealed to assembly rivets; metal-to-glass-sealed terminals can be furnished if specified.
Case Finish: tinned all over.
Markings: type number, voltage and capacitance rating, and terminal identification ink-stamped on case.
Insulation Resistance: never less than 2,000 megohms.
Dissipation Factor: less than 0.008 at 1,000 cycles.
Operating Temperature: minus 55C to plus 85C.

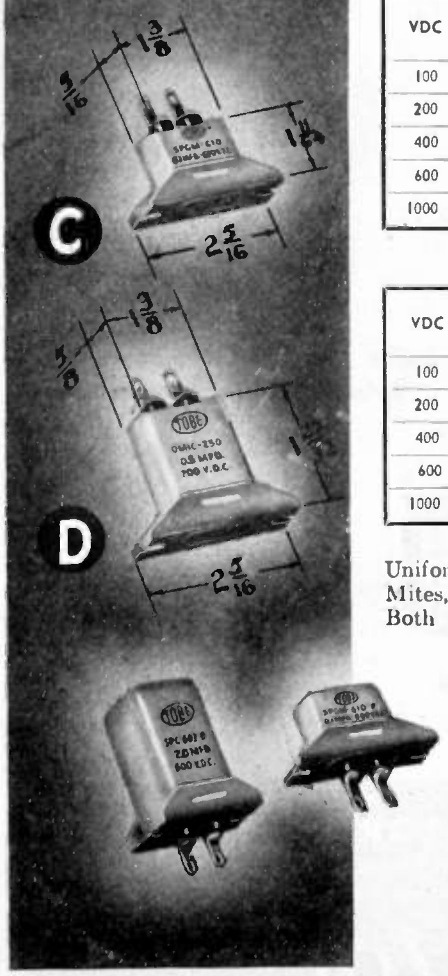
With Attached Channel Bracket

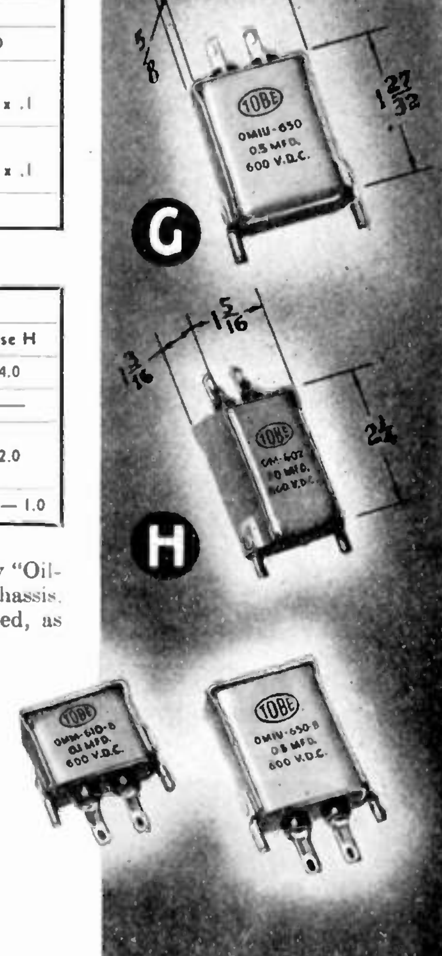
VDC	MFD			
	Case A	Case B	Case C	Case D
100	—	4.0	.01 — .25 2 x .05, 2 x .1	.01 — 1.0 2 x .05, 2 x .1
200	—	—		
400	—	2.0		.01 — .50 2 x .05, 2 x .1
600	.01 — 1.0			
1000	.05 — .50	1.0	.01 — .1	.01 — .25

With Reversible Hold-Down Bracket

VDC	MFD			
	Case E	Case F	Case G	Case H
100	—	.01 — .25 2 x .05, 2 x .1	.01 — 1.0 2 x .05, 2 x .1	4.0
200	—			—
400	—		.01 — .5 2 x .05, 2 x .1	2.0
600	.01 — 1.0			
1000	—	.01 — .1	.01 — .25	.05 — 1.0

Uniformity of size adds to the convenience afforded by "Oil-Mites," allowing gang installation above or below the chassis. Both upright and inverted mounting can be furnished, as illustrated. Where necessary, variation can be made in style and position of terminal lugs.





Reprints of this specification page are available and will be sent on request. For detailed data on "Oil-Mites" and other Tobe Capacitors ask for Catalog 4712RE.

TOBE

TOBE DEUTSCHMANN *Corporation* **CANTON, MASSACHUSETTS**

SMALLER, LIGHTER

"EVEREADY"

"A-B" BATTERY

for more compact portables!

Will outlast any other battery pack of its size!

SPECIFICATIONS:

Voltage:
"A"-9, tapped at 7½.
"B"-90.

Size:
9 7/32" x 2 23/32" x 4 5/16".

Net weight:
4 lbs. 15 ozs.

NOW the "Eveready" combination "A-B" 90-volt battery pack is available in even a smaller size... for the more compact, lighter portables. This pack provides plenty of power—it will last longer than any other "A-B" pack of comparable size.

This longer life is the result of the exclusive flat-cell principle found only in "Eveready" "Mini-Max" batteries.

It will pay you, in designing your new portables, to take advantage of this powerful, lighter-weight, smaller "Eveready" battery pack. For more details, consult National Carbon Company, Inc.

• Ordinary battery (*left*) is made of *round* cells and wasted space! "Eveready" battery (*right*) is made of *flat* cells—*no space between them* wasted by air, pitch, or cardboard!

EVEREADY
TRADE-MARKS
MINI-MAX
RADIO BATTERIES

The registered trade-marks "Eveready" and "Mini-Max" distinguish products of

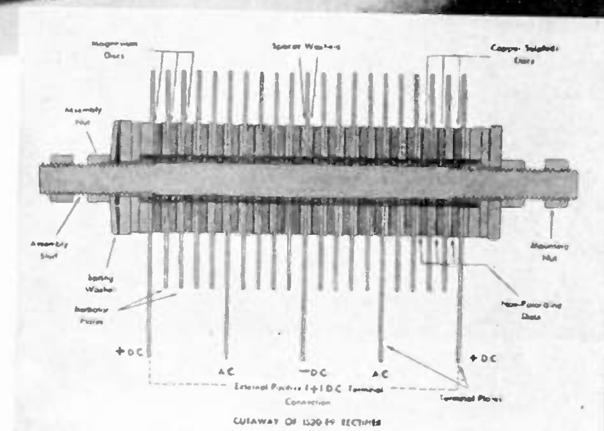
NATIONAL CARBON COMPANY, INC.

30 EAST 42nd STREET, NEW YORK 17, N. Y.

Unit of Union Carbide and Carbon Corporation

UCC

Hardly Bigger Than a Milk Cap . . .



yet this magnesium copper sulphide rectifier cell will carry 100 amperes!

This small-sized disc is the vital heart of the Mallory magnesium copper sulphide rectifier. Fitted between radiating fins, as illustrated at the right, it is protected from exterior damage.

More important, this disc will carry much greater current loads — operate under much higher temperatures — than similar junctions in other type rectifiers. Its overall area is smaller — a direct contribution to the compactness of the rectifier itself.

The fact is that space requirements of the Mallory magnesium copper sulphide rectifier are usually many times less than those of other dry disc rectifiers. Just one of many reasons why Mallory rectifiers outsell all other types of dry disc rectifiers for low-voltage, high current applications. See your Mallory distributor for Rectifier Catalogs. Or write direct for engineering assistance.

MCSR's Are the World's Toughest Rectifiers

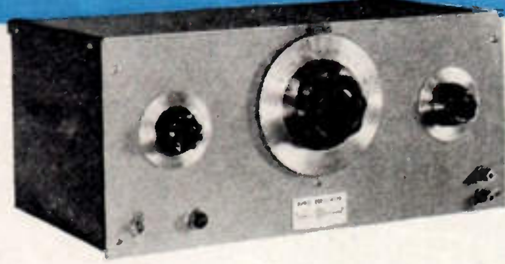
Check These Features

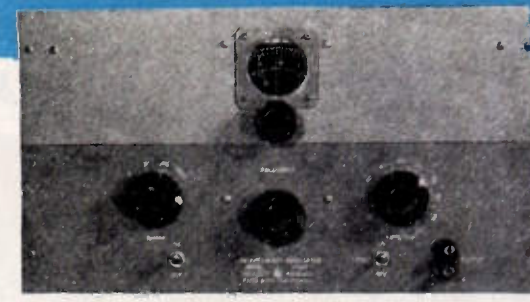
- Self-healing rectifying film
- Durable all-metal construction
- Small size, light weight
- No moving parts to wear out
- Resists harmful atmospheric conditions
- Output unaffected by temperatures
- Maximum overload range
- Constant output during rectifier life
- Low cost of operation

P. R. MALLORY & CO Inc
MALLORY RECTIFIERS
MAGNESIUM COPPER SULPHIDE RECTIFIERS — STATIONARY AND PORTABLE D. C. POWER SUPPLIES — BATTERY CHARGERS AND AVIATION RECTOSTARTERS*
*Reg. U. S. Pat. Off.

P. R. MALLORY & CO., Inc., INDIANAPOLIS 6, INDIANA

**Rectostarter is the registered trademark P. R. Mallory & Co., Inc., for rectifiers for use in starting internal combustion engines.*





200 SERIES AUDIO OSCILLATORS

Available in six standard models. *-hp-* 200A and *-hp-* 200B have transformer-coupled output delivering 1 watt into matched load. Primarily designed for audio testing. *-hp-* 200C and *-hp-* 200D have resistance-coupled output and supply constant voltage over wide frequency range. The *-hp-* 202D is a modification of the 200D, extending frequency downward to 2 cps. *-hp-* 200I is a spread-scale oscillator designed for interpolation work and for applications where oscillation frequency must be known with utmost accuracy.

202B LOW FREQUENCY OSCILLATOR

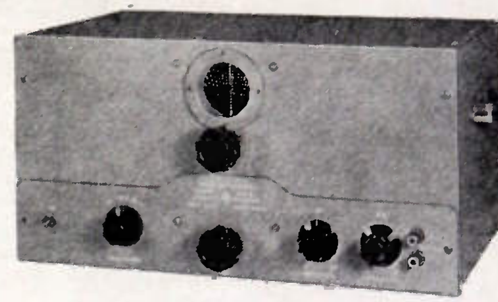
Specially designed for work between ½ cps and 1000 cps. Provides excellent wave form, good stability, split-hair measuring accuracy in the very low frequencies. Ideal for vibration or stability checks on mechanical systems, for testing geophysical, electro-cardiograph or electro-encephalograph equipment, checking response of seismographs, or electrical simulation of mechanical phenomena.

Resistance-Tuned Oscillators ...For Every Measuring Job
½ cps to 10 mc!

From A to Z in measuring, there's an *-hp-* resistance-tuned oscillator engineered to fit your exact need. Nine precision oscillators in all...and each bears the famed *-hp-* family characteristics of no zero set, constant output, low distortion, great stability, and decade tuning.

Brief data on these *-hp-* oscillators are given here. For complete details, write or wire today!

HEWLETT-PACKARD COMPANY
1508D PAGE MILL ROAD • PALO ALTO, CALIFORNIA

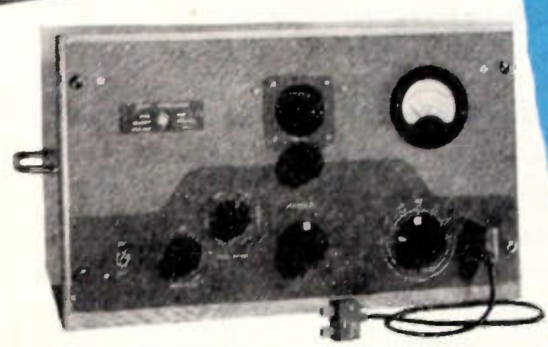


201B AUDIO OSCILLATOR

Meets every requirement for speed, accuracy, wave form purity and ease of operation in FM and other fields where high fidelity is most important. Provides 3 watts output into a 600 ohm resistive load. Distortion held to 1% or less, at 3 watts, ½% at 1 watt output. Excels in testing high fidelity amplifiers, speakers, and in comparing frequencies.

BRIEF SPECIFICATIONS

Instrument	Freq. Range	Output	Distortion	Freq. Response
-hp- 200A	35 cps to 35 kc	1 watt/22.5v	Less than 1%	±1 db to 15 kc
-hp- 200B	20 cps to 20 kc	1 watt/22.5v	Less than 1%	±1 db to 15 kc
-hp- 200C	20 cps to 200 kc	100 mw/10v	Less than 1% to 20 kc	±1 db to 150 kc
-hp- 200D	7 cps to 70 kc	100 mw/10v	Less than 1% 10 cps to 20 kc	±1 db throughout range
-hp- 202D	2 cps to 70 kc	100 mw/10v	Less than 2% 10 cps to 70 kc	±1 db, 7 cps to 70 kc
-hp- 200I	6 cps to 6 kc	100 mw/10v	Less than 1% above 10 cps	±1 db, 6 to 6000 cps
-hp- 202B	½ cps to 1000 cps	100 mw/10v	Less than 1% 1 to 1000 cps	±1 db, 10 to 1000 cps
-hp- 201B	20 to 20,000 cps	3 w/42.5v	Less than 1%	±1 db throughout range
-hp- 650A	10 cps to 10 mc	15 mw/3v	Less than 1% 100 cps to 100 kc	±1 db throughout range



650A WIDE-BAND OSCILLATOR

Continuous frequency coverage, 10 cps to 10 mc. Highly stable, versatile. Output flat within 1 db throughout frequency range. Available voltages range from .00003 to 3 v. Other advantages include 94" scale length, 6 to 1 micro-controlled tuning drive, 50 db output attenuator variable in 10 db steps, output voltage divider providing 6 ohm internal impedance (reducing output voltage 100 to 1).

hp laboratory instruments
FOR SPEED AND ACCURACY

AFTER YEARS IN COLUMBIA RECORDS' FILES

—they STILL speak for themselves

"Master safety disc No. 15B — an AUDIODISC — recorded December 12, 1939, was taken from our files and played back on September 12, 1947. This test showed that after almost eight years the recorded quality was still excellent and there was no measurable increase in surface noise. Surface noise of a new cut, made on this disc at the same date in 1947, was no different from the original cut."

This is the brief, factual report by Columbia recording engineers on a test made to measure the lasting qualities of AUDIODISCS. In the photograph the two large bands show the orchestral recording made in 1939. Close to these are the unmodulated grooves cut this year.

One more convincing proof of a most important claim — *"AUDIODISCS do not deteriorate with age either before or after recording, and there is no increase in surface noise from the time of recording to playback or processing—whether it be a few days or many years."*

AUDIO DEVICES, INC., 444 Madison Avenue, New York 22, N.Y.

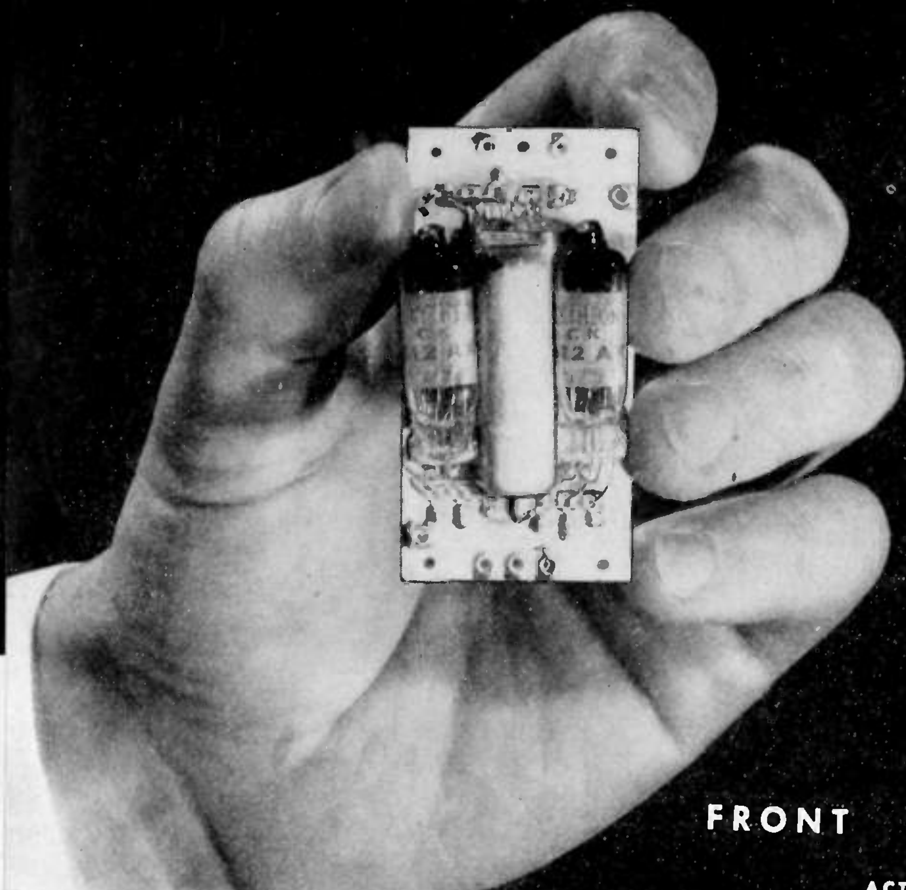
Export Department: Rocke International Corp., 13 E. 40th Street, New York 16, N. Y. * REG. U. S. PAT. OFF.

Audiodiscs are manufactured in the U.S.A. under exclusive license from PYRAL, S.A.R.L., Paris

they speak for themselves **audiodiscs**

Announcing the "Ampe

A typical new application of Centralab's revolutionary "printed electronic circuit"!

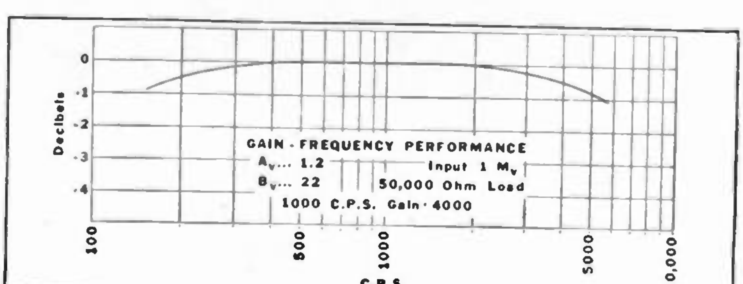


Important News!

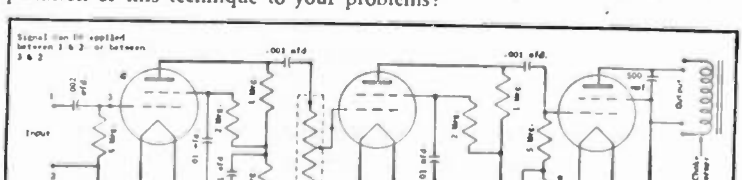
"Ampec" — a typical application of Centralab's printed electronic circuit — is a compact, highly efficient, dependable 3-stage audio amplifier. Can be designed for hearing aids, mike preamps or any electronic voltage or frequency applications where small compact size, high efficiency and reliability are required.

FRONT BACK

ACTUAL SIZE ILLUSTRATED



GAIN-FREQUENCY CHART above shows flat response of typical "Ampec" within 1 decibel between 200 and 5000 cycles. Why not investigate the application of this technique to your problems?



SCHEMATIC DIAGRAM of typical "Ampec" illustrated shows what components can be used. "Ampecs" can be designed to meet a wide range of gain or frequency response requirements.

Miniature, one-piece amplifier unit can offer complete electrical circuit from input to output! There's never been an electronic device like Centralab's new "Ampec"! Lightweight, durable, with reliability and efficiency heretofore unobtainable in small units, "Ampec" illustrates how you can get all components of an audio-amplifier — tube sockets, capacitors, resistors, wiring — "printed" on one, compact ceramic chassis according to your special requirements.

Look at these advantages: no jumble of wires to shift or come loose. Since "Ampecs" can be made from one "master plate", characteristics of all units are uniform . . . a complete unit can be replaced by an exact duplicate. Only 2.250" long, 1.156" wide, .187" thick over tube clips. Weight with tubes, 0.63 oz.

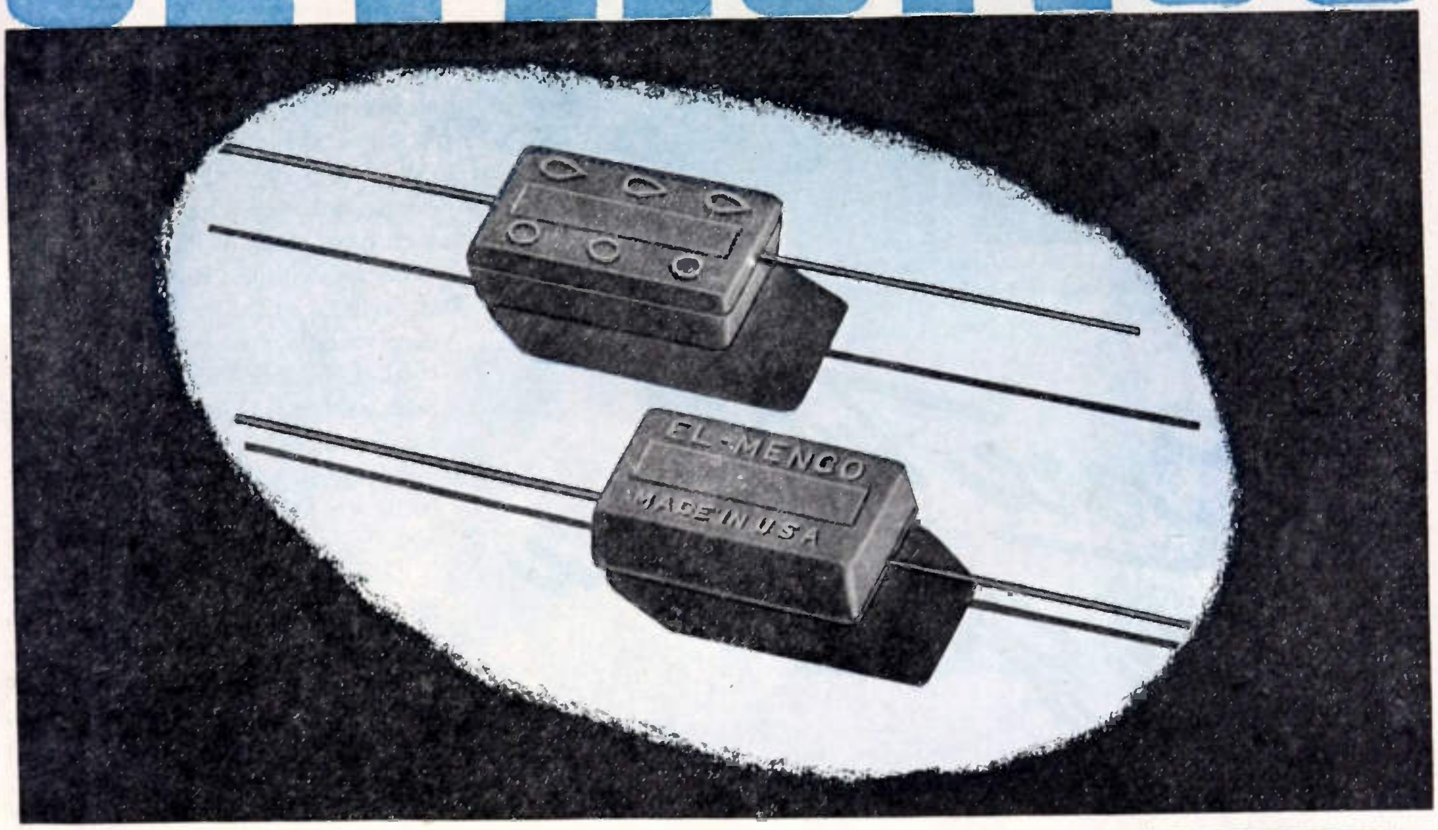
Write for Bulletin R973.

Centralab CRL

Division of GLOBE-UNION INC., Milwaukee

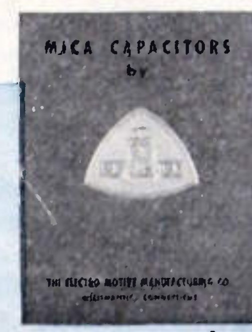
El-Menco

GOOD PRODUCTS BUILD GOOD NAMES



Because the makers of EL-MENCO Capacitors have always insisted on quality at any cost, the name EL-MENCO is now recognized as the identification mark of leadership.

EL-MENCO Capacitors—Molded Mica—Mica Trimmer. We are bending every effort to satisfy all manufacturers who want these quality products. Send us your specifications and we will do our level best to supply you.



Our new catalogue for manufacturers and engineers is now available . . . send for your copy.

THE ELECTRO MOTIVE MFG. CO., Inc. Willimantic, Conn.

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Is Sole Agent for El-Menco
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and Canada

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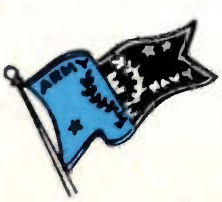
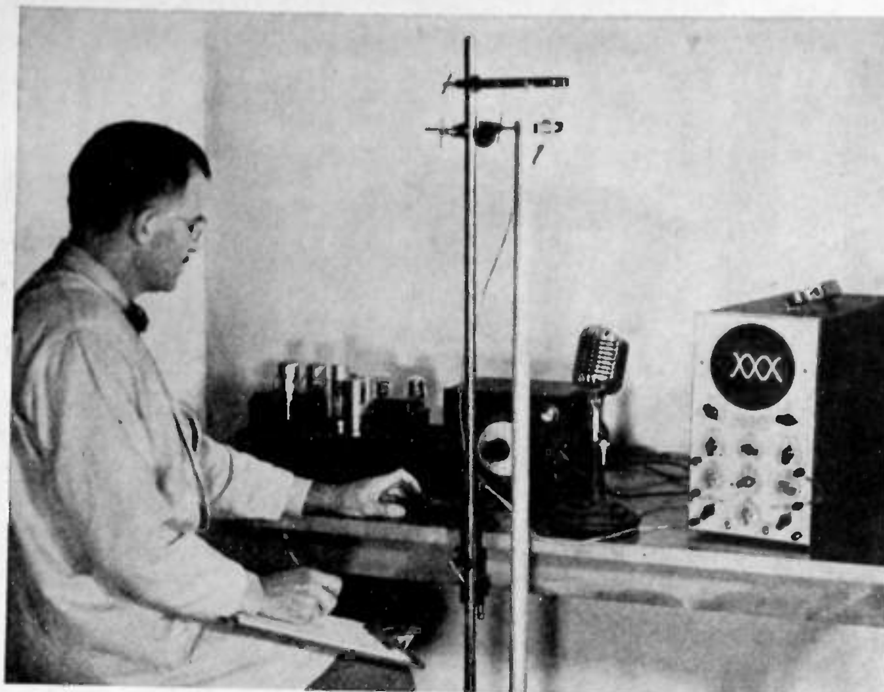


MOLDED MICA El-Menco **MICA TRIMMER**

CAPACITORS



Laboratory set-up for measuring tone of chime tubes. Lissajous figure on screen of cathode ray oscilloscope is being used to determine the frequency (cycles per second) of the chime's fundamental note.

Revere Tubes make Good Music

BECAUSE of the importance of the market for brass tube used in door chimes, Revere some time ago embarked upon a complete scientific study of the musical qualities of such tube, to determine the factors responsible for pleasing tone. Here is a brief report of the work, which offers an example of the thoroughness with which Revere attacks problems concerning the application of its mill products.

The first step was purely experimental. We proceeded by ear. Over 100 samples of tubes in various alloys, tempers and gauges were hung up, struck, listened to, and preferences obtained from many people. These tests indicated not only what was the best alloy, but also what were the proper temper and wall thickness requirements to produce the most acceptable and desirable tone. But Revere did not stop there. It was desirable to know what made that tone preferable, what were the factors that influenced it, and how they could be controlled. It was felt that only with such complete information in hand could Revere be in position to control chime tube quality accurately, and fill customers' orders reliably with a standard product.

The project then was turned over to a laboratory physicist who is also a talented musician. Here began the most ambitious and lengthy and scientific part of the work, employing the most modern electronic apparatus, including a beat-frequency oscillator and a cathode ray oscilloscope. These made it possible to dissect the tone produced, measuring the frequency and intensity of the fundamental note and its partials with an accuracy of one cycle per second. Much new information was uncovered. For example, the strike tone so clearly heard when the chime is struck does not actually exist in the tube, but is a difference tone between the 1st and 3rd partials. Hence, for good tone, those partials must be equal in intensity and duration.

It requires seven closely-typed pages just to sum up the work in general terms; the laboratory records fill a large volume. The net of it is that Revere really knows about all there is to know about chime tube, scientifically, musically, physically, and, of course, how to produce it. If you need such tube, come to Revere.

Perhaps you use brass tube not for its sound, but for its corrosion resistance, strength, machinability, the polish it takes, the ease with which it can be bent, soldered, brazed, plated. Revere also knows how to control the factors influencing such applications, so come to Revere for brass tube for any purpose.

Revere also makes other types of tube, including copper water tube, condenser tube in such alloys as Admiralty, Muntz, cupro-nickel, tube in aluminum and magnesium alloys, lockseam tube in copper alloys and steel, and electric welded steel tube. Many of these can be had not only round, but also square, rectangular, oval, and in various flutings and special shapes. The Revere tube line therefore is complete, and awaits your orders.

The Technical Advisory Service will gladly collaborate with you in such matters as selection of alloys, tempers and gauges, and in fabrication processes.

REVERE

COPPER AND BRASS INCORPORATED

Founded by Paul Revere in 1801

230 Park Avenue, New York 17, New York

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

VALUABLE EASY-TO-BUY ELECTRONICS

large stocks...

wide selection...

quick delivery, too!

Authorized WAA Electronics Distributors

EASTERN

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Carr Industries, Inc.
1269 Atlantic Ave.
Brooklyn, New York, N. Y.

Tobe Deutschmann Corp.
863 Washington Street
Canton, Mass.

Electronic Corp. of America
353 West 48th Street
New York, N. Y.

Emerson Radio & Phonograph Corp.
76 Ninth Ave.
New York, N. Y.

General Electric Co.
Bldg. 267; 1 River Road
Schenectady, N. Y.

General Electronics, Inc.
101 Hazel Street
Paterson, N. J.

Hammarlund Mfg. Co., Inc.
460 West 34th Street
New York, N. Y.

Hytron Radio & Electronics Corp.
76 LaFayette Street
Salem, Mass.

Johanns & Keegan Co., Inc.
62 Pearl St.
New York, N. Y.

Newark Electric Co., Inc.
242 West 55th St.
New York, N. Y.

Radio Parts Distributing Co.
128 West Olney Road
Norfolk, Va.

Smith-Meeker Engineering Co.
125 Barclay Street
New York, N. Y.

Standard Arcturus Corp.
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Newark, New Jersey

Sylvania Electric Products Inc.
Emporium, Pa.

Technical Apparatus Co.
165 Washington St.
Boston, Mass.

Tung-Sol Lamp Works, Inc.
95 Eighth Ave.
Newark, New Jersey

W. & H. Aviation Corp.
Municipal Airport
Rochester, New York

MIDWESTERN

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4410 N. Ravenswood Ave.
Chicago, Ill.

Belmont Radio Corp.
3633 S. Racine Ave.
Chicago, Ill.

Electro-Voice, Inc.
Carroll & Cecil Streets
Buchanan, Michigan

E. F. Johnson Co.
206 Second Ave., S. W.
Waseca, Minnesota

SOUTHERN

Navigation Instrument Co., Inc.
P. O. Box 7001, Heights Station
Houston, Texas

Southern Electronic Co.
611 Baronne St.
New Orleans, La.

PACIFIC

Cole Instrument Co.
1320 S. Grand Avenue
Los Angeles, Calif.

Hoffman Radio Corp.
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Los Angeles, Calif.

MANUFACTURERS WHOLESALERS—JOBBERS

See these WAA Authorized Distributors for *your* share! Their inventories include many types of unused electronic devices, tubes and equipment.

Purchasing of surplus electronic tubes, devices and equipment has been simplified to a high degree. These WAA Approved Distributors were selected on a basis of their ability to serve you intelligently and efficiently.

Write, phone or visit your nearest Approved Distributor for information concerning inventories, prices and delivery arrangements. Fill your needs—NOW—while inventories still permit large purchases and wide selection. You'll find you can "Save with Surplus."

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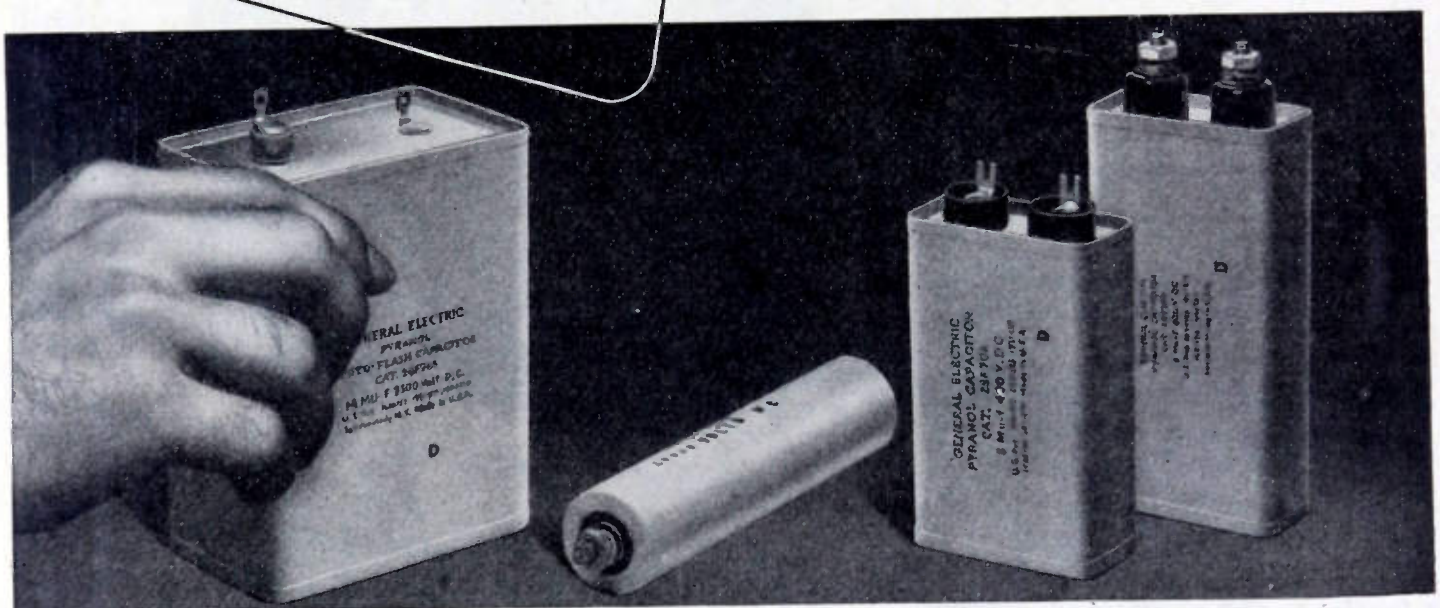

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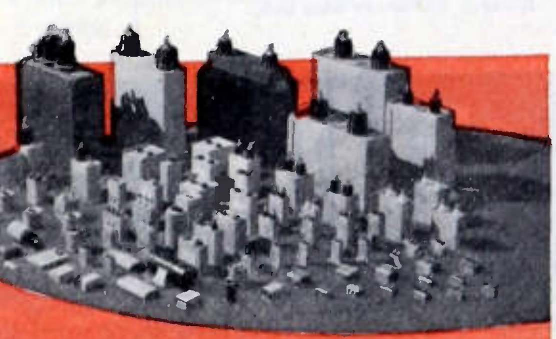

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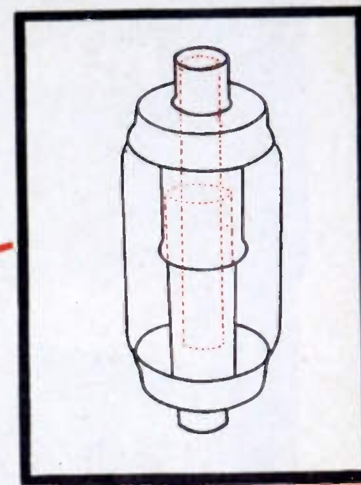

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PROCEEDINGS OF THE I.R.E.

(Including the WAVES AND ELECTRONS Section)

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PROCEEDINGS OF THE I.R.E.

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INSTITUTE NEWS AND RADIO NOTES SECTION

WAVES AND ELECTRONS SECTION

Dorman D. Israel

Mr. Israel was born in Newport, Ky., on July 21, 1900. He started in amateur "wireless" activities in 1914, and received his commercial operator's "ticket" in 1918. He was active in "wireless" clubs both at school and elsewhere in and around Cincinnati, until World War I came with its mandatory closing down of amateur activity.

He entered the University of Cincinnati as a co-operative student in electrical engineering in 1918, and for the first three years literally forgot "wireless" mainly because his co-operative job was with an electrical machinery manufacturer. Then in early 1921, he met Powel Crosley, Jr., who was determined to get into the radio business. Arrangements were made for Mr. Israel to go to work for what was then the Crosley Manufacturing Company as its first employee. This too was a co-operative job, since Mr. Israel was then a pre-junior at the University of Cincinnati (which is the third year of a five-year engineering course). In this job he soon found himself designing parts and equipment for home radio; and, to round things out, he also designed and installed the first two WLW transmitters of 100 and 1000 watts, respectively. It should be added that somewhere along the line he found the time to spend the required half of his daylight hours in school so that he received the E.E. degree in 1923. After graduation he stayed on with the Crosley organization as development engineer, and also made sales trips for the corporation through the southeast and southern parts of the United States.

In 1924 Mr. Israel and some associates became connected with Cleartone Radio Corporation in Cincinnati, and it was during his connection with this company that he pioneered considerably in tuned-radio-frequency receiver circuit developments and a.c.-operated vacuum-tube receivers.

He returned as chief development engineer to the Crosley Corporation in 1929, and became active there in the development of mass production techniques for superheterodyne and screen-grid-tube receivers. He was chief engineer of Grigsby-Grunow Corporation (Majestic Radio) for about a year in 1932, returning again to Crosley in 1933 as chief radio engineer.

Early in 1936, he became chief engineer of Emerson Radio and Phonograph Corporation in New York, and is still actively identified with the operations of this company. He is now vice-president in charge of engineering and production and a member of the board of directors of Emerson. In this connection, Mr. Israel has made many effective and valuable contributions to the art of engineering "small radio." He is also president and director of two Emerson subsidiary companies, Radio Speakers, Inc., of Chicago and Plastimold Corporation of Attleboro, Mass., as well as a director of another subsidiary, Jefferson-Travis, Inc.

Mr. Israel was identified with the start of the Television Broadcasters Association, having organized and conducted the panel sessions at their first convention in December, 1944. He has contributed much tc the engineering work of the Radio Manufacturers Association, serving on many receiver and systems committees. He is now chairman of the receiver section and of the Receiver Section Executive Committee in RMA. He recently was active in the formation and work of the Talking Book Systems Committee of the RMA. He is one of the organizers of the Cincinnati Section of I.R.E., of which he was Chairman in 1931.

He has been identified at various times with the work of the I.R.E. Sections, and has served on the following I.R.E. committees: Receivers, Television, Public Relations, Awards, RMA-I.R.E. Coordination, Convention Requirements, and Annual Review, and has given long service in the critical and demanding post of General Chairman of the Papers Procurement Committee.

He taught elementary radio engineering at the University of Cincinnati night college in 1928, 1929, and 1930. The range of his published papers varies in scope from a study of automatic volume control to a discussion of engineering education. Mr. Israel became an Associate of the I.R.E. in 1923, was made a Member in 1930, and a Fellow in 1941. He was awarded the Certificate of Appreciation from the War Department in 1946 for his outstanding work in connection with vacuum-tube fuzes.

One of the industrially important factors in technology is *systems engineering*. Illustratively, unless all aspects of a communication system are closely considered, both individually *and* collectively, and unless their corresponding specifications are correlated in a fashion consistent with system performance, the over-all effectiveness of the communication system and its parts will be lowered.

The author of the following guest editorial, who is the Editor of the journal *Audio Engineering*, has appropriately directed attention to one major aspect of present-day communication system engineering. This aspect may merit even closer technical study than it has at times received.—*The Editor.*

Audio Aspects of Postwar Radio Engineering

JOHN H. POTTS

Now that the transition to peacetime operation has been largely effected, it is interesting to survey some of the technological effects of the war upon radio engineering. During the prewar depression years, engineering emphasis was mainly on mass production of low-cost apparatus, with quality of construction and performance of secondary importance. The exigencies of war called for entirely different standards, with precision construction and excellence of operation imperative. Engineers learned how to make fine instruments and achieve quantity production without sacrifice of quality—knowledge badly needed by the radio industry.

Although, in the scramble to resume peacetime production of radio equipment in the least possible time, many manufacturers elected to revert to prewar designs and production standards, the resulting inferior apparatus found little public acceptance. The public expected something better. Those manufacturers who took a little longer to get into production, but did a better job with their war-gained know-how, suffered less.

But we must remember that, for a good many years, engineering emphasis has been placed largely on the development of carrier techniques, improved methods of transporting sound or other forms of intelligence from one point to another. During the war there was no need to improve the character of sound reproduction for esthetic purposes; this would have no military value. Research was confined to new concepts, such as radar, direction-finding, and the like. Thus the radio industry found itself at the end of the war with the production facilities and the skill needed to turn out fine apparatus in quantity, but without any additional experience in the design of equipment for improved reproduction of sound. Yet, insofar as broadcasting is concerned, we must remember that the sound quality is of paramount importance. In the past two years many improvements have been made in loudspeaker design and in phonograph pickups, but much remains to be done.

We need to provide better audio channels than those now available in reasonably priced receivers. We need a better demodulator; for many years there has been little research done on detectors, despite the well-known limitations of the diode. We need better i.f. amplifiers which will pass sidebands without attenuating the higher audio frequencies. Cabinets for the larger sets should be improved, acoustically and artistically. Our war-gained knowledge of mass production with close tolerances has provided us with smaller, improved tubes; it can also help in improving other components.

Radio broadcasting originally sprang into popularity because it provided better musical reproduction than the old mechanical phonograph then available. With greatly improved recordings and pickups, many receivers provide better reproduction from some records than is obtained from radio broadcast signals. Unless the radio manufacturers take heed of the situation before it is too late, public preference may revert to the phonograph.

Frequency Stabilization of Microwave Oscillators*

R. V. POUND†

Summary—**Two electronic circuits for frequency stabilization of electronically tunable microwave oscillators are described and discussed. One of these uses a microwave circuit equivalent to the low-frequency discriminator, in conjunction with a d.c. amplifier, to control the oscillator frequency at the frequency of a cavity resonator. The other circuit obtains frequency control of the oscillator by the cavity through a circuit operating almost entirely at an intermediate frequency. With both systems, frequency modulation of a highly degenerative type is provided. The resulting stability over long periods is essentially that of the cavity. The stability over short periods is such that the signal obtained occupies a band of less than 1 part in 10^8 in width. The practical limit to the stabilization obtainable with given components is estimated, and several applications are suggested.**

Introduction

AT LOW FREQUENCIES, the piezoelectric quartz crystal is used to control the frequency of an oscillator. Because it is very difficult to make quartz crystals that resonate at frequencies higher than the ordinary short-wave region, stable signal generators for higher frequencies are often obtained through the use of a crystal-controlled oscillator followed by frequency multipliers. This technique has been used to obtain stable frequencies in the microwave region, but the resulting signal generator is complex. Multiplier tubes for the higher microwave frequencies, for use in the last stage of such a device, are not readily available.

Cavity resonators having many of the desirable properties of the quartz crystal can be made for the microwave region. Although the resonant circuit of a microwave oscillator usually consists of a cavity resonator, it is not of a type having the highest Q or the greatest frequency stability obtainable because of the requirements imposed on its use as the tank circuit of the oscillator. On the other hand, cavities of the types used for wave meters and frequency standards, having unloaded Q's as high as 50,000 or more, can be made. Temperature compensation, through the use of tuning structures made from materials having different thermal coefficients of expansion, allows the resonant frequency of such cavities to be made temperature-independent. To obtain independence from the changes of atmospheric dielectric constant, the cavity may be hermetically sealed. Since the cavities for these frequencies need not be large, further independence of the resonant frequency from the ambient temperature can be obtained through the use of a temperature-regulated oven, as is common with quartz crystals. A property of the high-Q resonant cavity, not possessed by the quartz-crystal resonator, is the ability to be tuned continuously through a wide band of frequencies by a simple mechanism. If an oscillator could be made to possess the frequency stability of such a cavity, such an oscillator would compare favorably, as a source of signal power having a stable frequency, with the crystal-controlled oscillator and multiplier. In addition, a stable, tunable source of signal power could be obtained at any frequency for which oscillators are available.

An external cavity can be made to control the frequency of an oscillator in a direct manner by coupling the cavity to the oscillator in such a way that the external cavity appears to be the tank circuit of the oscillator. If, for instance, the high-Q external cavity is coupled to the cavity of the oscillator through a transmission line having an effective length of an integral number of half-wavelengths, the rate of change of susceptance of the combined circuit is determined mainly by the external cavity. Considerable improvement in frequency stability can be obtained in this way, but the circuit is not easy to set up. With most oscillator tubes a part of the coupling circuit must be a built-in output line, and the effective electrical length of this line varies among different tubes of the same type. As a result, to obtain frequency control over the same range of frequencies with different tubes, an adjustable circuit must be used.

The circuits for frequency control to be described here utilize the external cavity in a special microwave circuit. This circuit develops a voltage which is a measure of the difference between the frequency of an oscillator fed into it and the resonant frequency of the cavity. When this voltage is amplified and superimposed in the correct sense on the supply voltage of an element of the oscillator, the potential of which affects the frequency of the oscillator, the difference frequency is reduced. The time of response of the circuits has been kept small in order to reduce the frequency-modulation components having audio- and higher-frequency periods. In

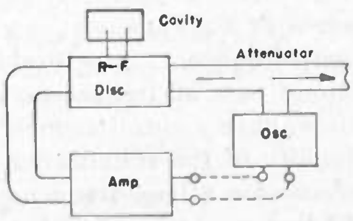

Fig. 1—Block diagram of electronic frequency-stabilization system.

this way a very narrow frequency band is occupied by the resulting signal, making it useful for measurements on extremely high-Q circuits and for narrow-band voice communication. If only the low-rate drifts are removed

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† Formerly, Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.; now, Society of Fellows, Harvard University, Cambridge, Mass.

from an oscillator in the microwave region, as is the more common practice, the resulting signal is still too broad for these purposes in most cases.

To obtain such a reduction in the signal bandwidth, the control circuit must be made to act as rapidly as possible. A block diagram of a control circuit of the type under discussion is shown in Fig. 1. Suppose the output terminals of the amplifying device are disconnected from the control terminals of the oscillator. If an alternating voltage were to be applied to the control terminals of the oscillator, frequency modulation would result, and this, in turn, would produce an alternating voltage of the same frequency at the output terminals of the amplifier. The entire device may, then, be considered as a voltage amplifier and, to obtain stabilization of the oscillator frequency, the input terminals of the amplifier must be connected to the output terminals. The device resulting is, therefore, analogous to an amplifier with a very large amount of negative feedback.

Suppose the oscillator signal to contain, before application of the stabilizing circuit, frequency components corresponding to frequency modulation at a given audio frequency and to a deviation from an average frequency of $\pm d\nu_0$. There will be an output voltage from the stabilization amplifier equal to $Gd\nu_0$ where G expresses the output voltage per unit of frequency deviation and is a complex function of the modulation frequency. When the output terminals of the amplifier are connected to the control terminals of the oscillator, the frequency deviation is reduced to $d\nu$.

$$d\nu = \frac{d\nu_0}{1 + AG}$$

where A is the frequency change produced by unit change in the voltage supplied at the control terminals of the oscillator. The stabilization factor S is

$$S = \frac{d\nu_0}{d\nu} = 1 + AG.$$

The analogy to a negative-feedback amplifier is apparent in this expression. For the operation to be stable, the quantity AG must not be equal to -1 at any frequency in the feedback loop. This puts restrictions on the amplifying system, for, as shown by Bode,[1] there is a minimum phase shift accompanying a given rate of change of gain with frequency, in realizable networks. The amplifier cannot pass all frequencies equally well and must, therefore, have a cutoff at high frequencies. For complete stability of the stabilization circuit, the gain cannot decrease for a large frequency range at a rate as great as 12 db per octave at frequencies less than the frequency of unity gain. An amplifier of many stages ordinarily has a cutoff rate far exceeding 12 db per octave unless special precautions are taken. The only precaution taken in the systems to be described is the use of amplifier circuits having a wide pass band compared with the pass band of a single resistance and capacitance circuit that produces the cutoff at high frequencies. Without doubt this aspect of the systems could be improved if it were found desirable. One notable difference between the present circuits and the ordinary negative-feedback amplifier is the fact that in the present circuits it is not necessary that the gain be constant through the frequency band for which stabilization is desired. The frequency-modulation components of the greatest magnitude are usually at the power-supply frequency and the first few harmonics of it, and the gain at these frequencies is of the most importance.

In most practical circuits the quantity A is not a function of the oscillator tube alone, but is partially determined by the character of the load circuit of the oscillator. This is particularly true with low-power oscillators, where the attenuation between the oscillator and the circuit containing the high-Q cavity cannot be very great. Under this condition the presence of the high-Q cavity as a part of the load circuit affects the dependence on frequency of the susceptance within the tank circuit of the oscillator. Therefore, the amount of frequency change produced by unit change in reflector voltage depends on the nature of the load circuit. The rate of change of susceptance of the tank circuit with frequency may be increased or decreased by the presence of a resonant cavity in the load circuit, depending on the effective electrical length of the coupling circuit. An increase in the rate increases the stability of the tube before the application of the electronic feedback circuit, and a decrease decreases the stability. The magnitude of A is correspondingly decreased or increased, respectively. Therefore, the gain through the feedback loop can be altered by a change in the effective phase length of the line coupling the external cavity to the oscillator, and instability of the feedback circuit can result even if the magnitude of the coupling and the amplifier gain are the same as at a line length for which stable operation results.

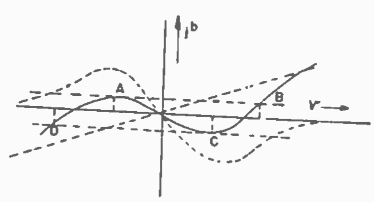

Fig. 2—Susceptance versus frequency in an oscillator with a high-Q load circuit.

In some instances the presence of a resonant cavity in the load circuit of the oscillator can cause the oscillator to tune discontinuously through the frequency of resonance of the cavity, completely skipping the resonant frequency. If this happens the stabilization circuit cannot function properly, and coupling circuits resulting in this kind of operation must be avoided. Curves of the

[1] H. W. Bode, "Relations between attenuation and phase in amplifier design," *Bell Sys. Tech. Jour.*, vol. 19, pp. 421–454; July, 1940.

susceptance as a function of frequency, showing how this discontinuous operation comes about, are shown in Fig. 2. The straight dashed line represents the susceptance of the tank circuit of the oscillator, and the curve formed by the short dashes represents the susceptance produced by a resonant cavity coupled to the oscillator through a line an odd number of quarter-wavelengths in effective length. If the operation of the oscillator is continuous for a line of this length it must be continuous for all other line lengths because, as reference to an admittance chart will show, the maximum possible negative rate of change of susceptance occurs for this line length. In the figure the solid line represents the total susceptance as a function of frequency, and the coupling is sufficient to produce a frequency discontinuity. If the tube were tuned through the resonance of the external cavity from the low-frequency side, the oscillator would skip discontinuously from the frequency corresponding to A to that corresponding to B. Approaching resonance from the other direction results in a skip from C to D.

To avoid such a discontinuity, the coupling must be such that the magnitude of the rate of change of susceptance of the load circuit with frequency is less than that of the oscillator circuit, when measured at the same point in the coupling line. If they are equal the operation is continuous, but the tube is very unstable at the resonant frequency of the cavity and the quantity A is infinite. This is illustrated in Fig. 3.

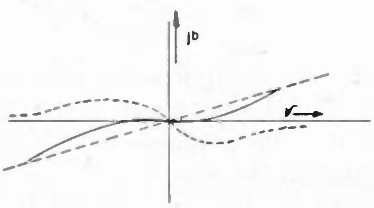

Fig. 3—Susceptance versus frequency at critical coupling.

To measure the rate of change of susceptance of the tank circuit of an oscillator with frequency at a point in the output coupling line where that of the load circuit can be expressed, a number related to the "pulling figure" of the oscillator may be used. The change in oscillator frequency per unit change in the load susceptance, in units of the characteristic admittance of the wave guide, may be used and termed C. The condition for continuous operation is that the rate of change of the load susceptance in these units with frequency must not exceed $1/C$ in magnitude.

For ordinary low-power oscillator tubes and high-Q cavities, this condition is not easily met. To meet it, the coupling aperture to the cavity may be made very small; or a matched dissipative attenuator, such as a tapered strip of carbon-coated bakelite, if the coupling line is a wave guide, may be used between the cavity and the oscillator. In the latter instance the amount of attenuation required may be calculated as follows:

In units of the characteristic admittance of the coupling line, the cavity admittance, as can easily be shown from an equivalent simple shunt-resonant circuit, is, to a very good approximation,

$$Y_c = \frac{\delta_0}{\delta_1} + j\,\frac{2\Delta\nu}{\delta_1}, \qquad (1)$$

in a plane of reference to which the equivalent shunt circuit applies. The quantitity δ_0 is the reciprocal of the unloaded Q of the cavity, δ_1 is $(\delta_L - \delta_0)$ where δ_L is the reciprocal of the Q resulting when the cavity is loaded with a semi-infinite input line, and $\Delta\nu$ is $(\nu-\nu_0)/\nu_0$ where ν is the frequency of operation and ν_0 is the resonant frequency of the cavity. A susceptance varying with frequency like that of the load circuit of Fig. 2 is obtained at a point an odd number of quarter-wavelengths toward the generator from the position in the line feeding the cavity to which (1) applies. At such a point the admittance is Y_1, the reciprocal of Y_c in the same units. The effect of an attenuator may be taken into account by writing the reflection coefficient Γ_1 associated with Y_1.

$$\Gamma_1 = \frac{1 - Y_1}{1 + Y_1}.$$

An attenuator reducing the power incident on the cavity to r times that incident on the attenuator reduces the reflection coefficient to

$$r\Gamma_1 = \frac{r(1 - Y_1)}{1 + Y_1}.$$

The admittance at the input to the attenuator is

$$Y_{cr} = \frac{1 - r\Gamma_1}{1 + r\Gamma_1}.$$

The susceptance is the imaginary part of this, and the rate of change of the susceptance with frequency is, at the resonant frequency of the cavity where it is a maximum,

$$\frac{dB}{d\nu} = \frac{8r\alpha}{\nu_0\delta_0\left[(1 - r)\alpha + (1 + r)\right]^2}$$

where α has been written for δ_1/δ_0. The condition for continuous tuning of the oscillator through the cavity resonance is, then,

$$\frac{8r\alpha}{\nu_0\delta_0\left[(1 - r)\alpha + (1 + r)\right]^2} < \frac{1}{C}. \qquad (2)$$

This is satisfied if

$$r < \frac{\left(\alpha^2 + \dfrac{4\alpha C}{\delta_0\nu_0} - 1\right)}{(\alpha-1)^2}\left\{1 - \left[1 - \frac{(\alpha-1)^2(\alpha+1)^2}{\left(\alpha_2 + \dfrac{4\alpha C}{\delta_0\nu_0} - 1\right)^2}\right]^{1/2}\right\}.$$

For most applications the second term in the square brackets is small compared with unity, and a series expansion gives

$$r < \frac{1}{2} \frac{(\alpha + 1)^2}{\left(\alpha^2 + \dfrac{4\alpha C}{\delta_0 \nu_0} - 1\right)} \tag{3}$$

neglecting terms in the expansion to higher than the first power. For stabilization circuits it will be shown that the best results are obtained for α equal to unity, and for this condition (2) leads directly to

$$r < \frac{\delta_0 \nu_0}{2C}. \tag{4}$$

In the 10,000-Mc. region and with a 2K25 oscillator tube, δ_0 might be 4×10^{-5}, and C about 10 Mc. per unit change in susceptance. This is a value for C found by measurement of the coupling conditions under which continuous operation of several 2K25's could be obtained for all parts of the reflector-voltage mode between the points at which the delivered power was one-fourth that at the center of the mode. With these values, r must be less than 0.02. Thus an attenuation greater than 17 db must be used between the cavity and the oscillator, in order that continuous tuning of the oscillator through the resonant frequency of the external cavity will be obtained, for this most restrictive effective length of the coupling line. In the stabilization circuits, special circuits called "magic tees" are used between the oscillator and the cavity, and the effect of each of these is equivalent to a 3-db attenuator. About 12 db of additional attenuation is required, therefore.

The Magic Tee

The magic tee is a circuit which can be formed from wave guides having rectangular cross sections, in the manner illustrated in Fig. 4. From the symmetry and

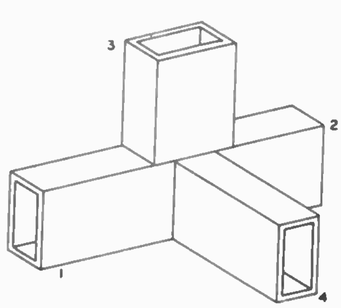

Fig. 4—Wave-guide "magic tee."

considerations of the fields in these wave guides, it is easily shown that a wave sent into arm *3* of the structure excites waves of equal amplitudes traveling outward from the junction in arms *1* and *2*, and that these excited waves have like phases at planes equidistant from the junction. On the other hand, a wave sent into arm *4* excites waves of equal amplitudes but having opposite phases at planes equidistant from the junction in arms *1* and *2*. Because of the opposite kinds of symmetry of the waves excited in arms *1* and *2* by waves sent into arms *3* and *4*, no direct coupling exists between the latter arms. If arms *1* and *2* are terminated in nonreflecting loads, no power is delivered to a load on arm *4* if a wave is sent into arm *3*, and similarly no power is delivered to a load on arm *3* if power is sent into arm *4*.

It is easily shown that the addition of matching irises to eliminate the reflections at the junctions, for waves sent into arms *3* and *4* with arms *1* and *2* terminated in reflectionless loads, results also in zero direct coupling between arms *1* and *2*. Such matching structures having a wide pass band have been developed at the Radiation Laboratory[2] and elsewhere, and it is to the resultant device that the term "magic tee" is applied. Circuits equivalent to the magic tee, such as that shown in Fig. 5, can be used with equal success at wavelengths where they are more convenient.

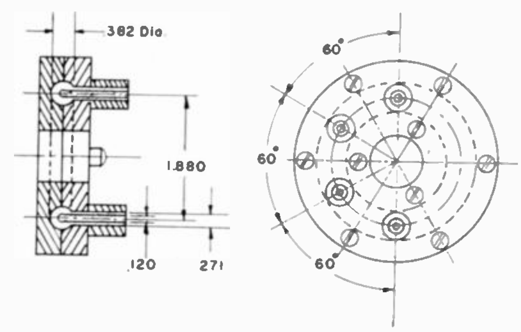

Fig. 5—10-centimeter coaxial line equivalent to a "magic tee."

The magic tee can be represented by an equivalent network having four pairs of terminals. The terminals may be supposed to lie in planes chosen to be equidistant from the junction in arms *1* and *2*. In arms *3* and *4* the terminals may be taken to lie in planes at which zero admittance would be found if arms *1* and *2* are short-circuited in the planes chosen for the terminals in those arms. There are such planes in arms *3* and *4* every half-wavelength from those closest to the junction. For one choice of these two planes, the relations between the voltages and currents in the terminals of the equivalent network can be shown to be given by the relations

$$\begin{aligned} i_1 &= j\frac{\sqrt{2}}{2}(e_3 + e_4)Y_0 \\ i_2 &= j\frac{\sqrt{2}}{2}(e_3 - e_4)Y_0 \\ i_3 &= j\frac{\sqrt{2}}{2}(e_1 + e_2)Y_0 \\ i_4 &= j\frac{\sqrt{2}}{2}(e_1 - e_2)Y_0 \end{aligned} \tag{4a}$$

[2] R. V. Pound, Radiation Laboratory Series, "Microwave Mixers," vol. 16, McGraw-Hill Book Co., New York, N. Y., to be published.

where Y_0 is the characteristic admittance of the wave guide. The signs in these relations are changed if the plane of the terminals in either arm *3* or arm *4* is changed by an odd number of half-wavelengths.

As an example of the use to which the equivalent circuit may be put, the power delivered to a load having an admittance Y_4 on arm *4* from a generator having an admittance Y_3 on arm *3* may be calculated, for arms *1* and *2* terminated with admittances Y_1 and Y_2, respectively. Such a calculation yields

$$P_4 = 4P_0 g_3 g_4 \left| \frac{Y_1 - Y_2}{(1+Y_1Y_4)(1+Y_2Y_3)+(1+Y_1Y_3)(1+Y_2Y_4)} \right|^2 \quad (5)$$

where P_0 is the power available from the generator; g_3 and g_4 are, respectively, the conductive parts of Y_3 and Y_4; and all admittances are expressed in terms of the characteristic admittance of the wave guide.

The Microwave Discriminator

Two different kinds of stabilization circuits have been constructed. One of them utilizes a microwave circuit that is equivalent to the frequency discriminator used at low frequencies. In Fig. 6 a symbolic diagram of one

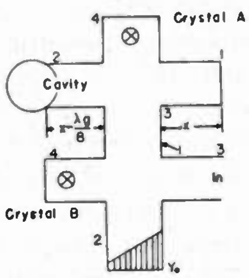

Fig. 6—Magic-tee frequency discriminator.

form of this microwave discriminator is shown. There are two symbols representing magic tees, and the numbers on the arms refer to the arms numbered in the same way in Fig. 4. Other orientations of the magic tees can be used because of the complete symmetry of the structures, but it is most convenient to use arms *1* and *2* for the cavity and the comparison short circuit.

The source of power is connected to arm *3* of the lower magic tee. One-half the power sent into the circuit is delivered to the matched termination on the lower arm, while the other half is sent upward into the other tee through arm *3*. This excites waves of equal amplitudes and like phases in arms *1* and *2*, and these travel outward toward the cavity and the short circuit. At frequencies far removed from the resonant frequency of the cavity, the cavity reflects completely, and it is so positioned that it appears like a short circuit one-eighth wavelength farther from the junction than the short circuit on the opposite arm. At these frequencies the waves reflected in the two arms reconverge on the junction with a relative phase of $\pi/2$ radians, at planes equidistant from the junction, because the wave on the side containing the cavity has traveled a total of a quarter-wavelength farther than the other. Because arm *4* is excited by a wave possessing odd symmetry about the junction plane and arm *3* by one possessing even symmetry, waves of equal amplitude are excited and travel outward in arms *3* and *4* of this magic tee at this frequency. A matched crystal in arm *4* detects this wave, and one-half the power returned out arm *3* is detected by a matched crystal on arm *4* of the lower magic tee. The function of the lower tee is to detect the power returned from the upper one without coupling directly to the input signal. Because only one-half the returned power is delivered to the lower crystal, if the crystals are "square law" the detected voltage at the upper crystal is twice that at the lower crystal. An attenuator on either the r.f. or the d.c. side of the upper crystal can be used to make the output voltage of that crystal the same as that at the lower crystal for all frequencies far removed from the resonant frequency of the cavity.

At the resonant frequency of the cavity, the reflection coefficient of the cavity has the same or the opposite phase to that at frequencies far removed from resonance, corresponding to a conductance either larger or smaller than the characteristic admittance of the wave guide. As a result, equal amounts of power are sent back out arm *3* and out arm *4* of the upper tee for an input signal at the resonant frequency. The same attenuator results in equal output voltages from the two crystals. At frequencies not at the cavity resonance but adjacent to it on either side, the reflection coefficient of the cavity is either advanced in phase or retarded in phase relative to the phase at resonance. Therefore, for frequencies near resonance on one side, the power delivered to arm *4* of the upper magic tee is greater than that returned to arm *3*. On the other side of resonance, arm *3* receives the greater power. The difference between the voltages rectified by the two crystals is, therefore, with the balancing attenuator in place, a function of frequency similar to the output voltage of the conventional discriminator circuit.

If Y_c from (1) is substituted for Y_1 (5) and Y_2 is set equal to $-j$, the admittance of a short-circuited one-wavelength line, an expression for the power delivered to the upper crystal is obtained. The power delivered to the lower crystal, from the above description of the operation, can be seen to be one-half this power with the sign of $\Delta\nu$ reversed. Taking one-half the power delivered to the upper crystal less the power delivered to the lower crystal yields an expression proportional to the output voltage of the discriminator, assuming matched magic tees and matched crystals producing voltages proportional to the incident power. Thus the discriminator output voltage is found to be

$$V = P_0 D \frac{\alpha a}{(1+\alpha)^2 + a^2} \text{ volts} \quad (6)$$

where P_0 is the power available from the matched generator connected to the discriminator, a is $2\Delta\nu/\delta_0$, and D

is the rectification efficiency of the crystals in volts per unit incident power. The rate of change of the discriminator voltage with frequency is greatest at resonance, or for a equal to zero, and is

$$\frac{dV}{d\nu} = DP_0 \frac{Q_0}{\nu_0} \frac{2\alpha}{(1+\alpha)^2}. \tag{7}$$

This is a maximum for α equal to 1, and for the frequency-stabilization circuit this is the optimum value of α. Curves of V/P_0D from (6) are plotted in Fig. 7 for values of α from 0.5 to 10.

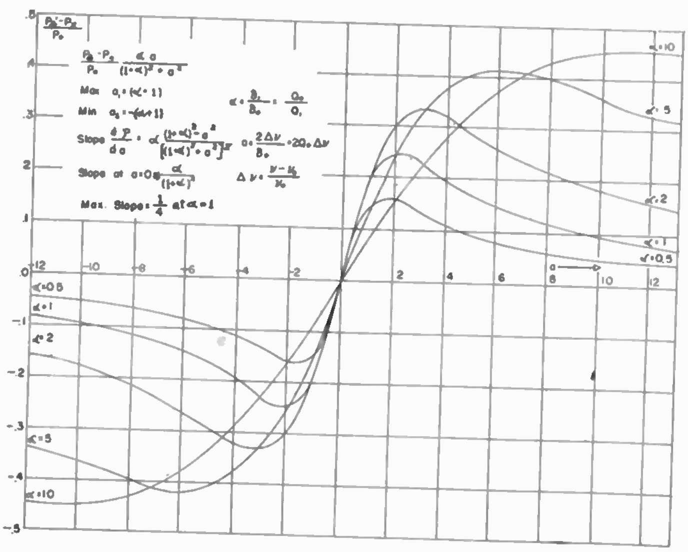

Fig. 7—Output voltage versus frequency for various cavity-coupling factors of the balanced magic-tee discriminator.

If the coupling between the oscillator and the discriminator is limited by the need to avoid frequency discontinuities, for the range of coupling in which the crystals remain square law, the maximum obtainable slope is independent of the cavity Q and α, and is determined by the quantity C and the power available from the oscillator. This results because the attenuation r, from (3), must be used, and P_0 is proportional to r. Nevertheless, it will be shown later that α equal to unity and the highest possible Q_0 give the best operation of the stabilizing circuit for a tube with a given C and available power.

The D.C. Stabilizing Circuit

A frequency-stabilizing circuit using the microwave discriminator can be made in conjunction with a d.c. amplifier, as illustrated in Fig. 8. The circuit diagram of a d.c. amplifier that has been used in this application is shown in Fig. 9. The amplifier has two push-pull stages using 6SH7G tubes. The balancing of the output of the two crystals is obtained by the adjustment of a potentiometer between the upper crystal and the amplifier tube. A potentiometer in the plate circuit of the first stage is used to balance the amplifier. A large negative voltage was required to lower the d.c. level of the output voltage of the amplifier to the −100-volt region for application to the reflector of a 2K25 oscil-

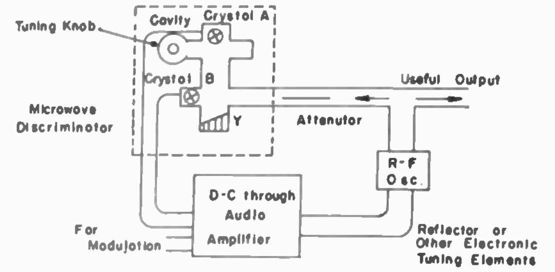

Fig. 8—Block diagram of a d.c. stabilizer

lator tube (cathode grounded) through a potentiometer, without a large sacrifice in gain. This same negative voltage is used to obtain stability through the large common cathode resistors in each stage. Further stability is obtained by the use of negative feedback from the plates of the second stage to the cathodes of the first. The voltage gain of the amplifier between the input terminals and the plates of the second stage is 2000, although the gain from the input to the reflector is only about 600 because the reflector can be supplied only from an unbalanced line.

A capacitor of about 0.01 μfd. capacitance connected from the reflector to ground potential provides the high-frequency cutoff and prevents singing of the stabilizing circuit. With this amplifier used in a stabilizing circuit for a 2K25 tube in the region of 3.2 centimeters with a TE_{01}-mode wavemeter cavity, a stabilization factor of several hundred is obtained. A discriminator slope of about 1 volt per Mc. is obtained, and the oscillator frequency is changed by approximately 1 Mc. per volt.

Once locked, the frequency of the oscillator follows changes in the cavity frequency over the range of electronic tuning available at the reflector. If a wider tuning range than this is required, a 2K45 or similar oscillator tube may be used. This tube can be tuned electronically by a bias voltage at the grid of a special triode contained

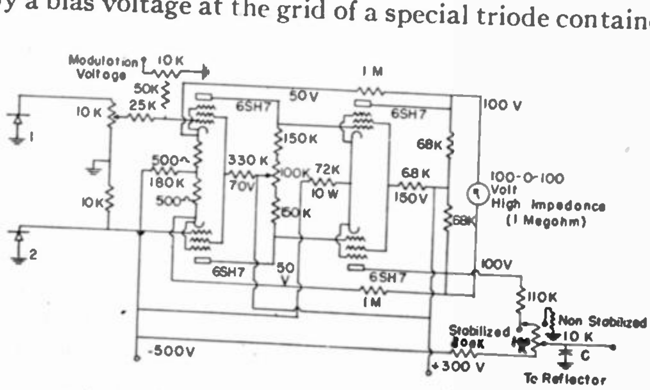

Fig. 9—D.c. amplifier for the electronic frequency stabilizer.

in the tube. The potential of this grid controls the current to the plate and, therefore, the temperature of the plate. The plate is so connected to the oscillator resonator that a distortion produced by a change in the temperature of the plate changes the tuning of the oscillator, and an electronic tuning range of 12 per cent is obtained in this way. The rate of tuning is relatively slow, however, and for frequency control sufficient to remove audio-frequency modulation components from the oscillator signal the control voltage must be connected to the reflector. Connection also to the tuning grid, however, through a circuit similar to that shown in Fig. 10, has been used to obtain single-knob tuning over more than a 10 per cent range of frequencies. Only a few volts are required to tune the tube over this range, and the difference of the oscillator frequency from the cross-over frequency of the discriminator required to develop the tuning voltage is not large if the amplifier gain is sufficient. In the circuit diagram, 1N34 crystals are

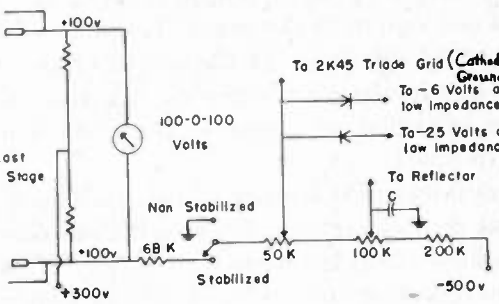

Fig. 10—Supplementary circuit for wide-band single-knob tuning.

shown used as clamping diodes to prevent the grid from being driven positive or too far negative. These also prevent motorboating of the oscillator into and out of oscillation. Fortunately, the change in reflector voltage required to keep the tube in oscillation when a change is made in the grid voltage is in the same direction as the change in grid voltage. Therefore, the tube can be kept in oscillation over the 10 per cent band. It does not, however, remain at the center of a mode. This would be remedied by reduction in the tuning rate of the thermal triode by use of a degenerative cathode resistor.

The frequency of the stabilized oscillator can be modulated very simply. To obtain frequency modulation the modulating voltage is superimposed on the output of the discriminator. A change in voltage at the input of the oscillator by this means results in a change in frequency with amplitude sufficient to produce a compensating change in the output voltage of the discriminator. The frequency modulation is thus very highly degenerative, and the response of the system is uniform from zero modulating frequency up to a frequency so high that the stabilization factor is not much larger than unity. The linearity of the modulation is determined by the constancy of the slope of the discriminator characteristic. Little harmonic distortion is produced for deviations as large as $\pm\nu_0/4Q_L$ where Q_L is the loaded Q of the cavity in the discriminator circuit. With the amplifier of Fig. 8, a uniform deviation in frequency for a given amplitude of modulating voltage was obtained for modulation frequencies from zero to 50 kc.

The I.F. Stabilizing Circuit

To overcome certain limitations in the stabilization associated with the d.c. amplifier and the use of crystals as detectors, another stabilization system eliminating these components has been developed. A block diagram of this system is shown in Fig. 11. From the output

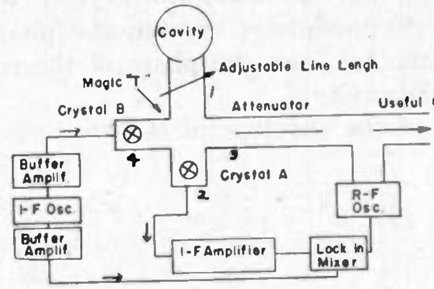

Fig. 11—Block diagram of the i.f. stabilizer.

terminals of a crystal mixer in the microwave circuit of this system is obtained an i.f. voltage proportional in magnitude and dependent in phase on the imaginary part of the reflection coefficient of the cavity.

This i.f. voltage is obtained through the use of a magic tee, terminated on arm *1* with the cavity fed through a line of variable length. A crystal connected to an i.f. oscillator is connected to arm *4*, and arm *2* is terminated by the mixer crystal. The oscillator to be stabilized is fed into arm *3*. The oscillator is fed into the magic tee through a matched attenuator to insure continuous operation of the oscillator, and one-half the available power from this attenuator is delivered, without reflection, to the mixer crystal A. In the opposite arm, the wave excited by the oscillator is reflected by the cavity in a phase and amplitude depending upon the frequency relative to the resonant frequency of the cavity. The reflected wave couples in part to the attenuator in the input arm and in part to the modulator crystal B. This crystal does not reflect when zero voltage exists across its i.f. terminals, but when driven by the large i.f. voltage it reflects two sideband frequencies, above and below the oscillator frequency by an amount equal to the intermediate frequency. The sideband waves travel back to the cavity and to the mixer crystal. The waves returned to the cavity are again reflected and are absorbed with some production of waves at the original frequency and second-order sideband frequencies. These have little effect on the operation of the system and may be neglected.

Arriving at the mixer crystal A are waves at three different frequencies, and the linear combination of these waves may be seen to be

$$E_B = \frac{\sqrt{2}}{2} E_0 \left\{ \cos \omega_1 t + \frac{|\Gamma_c| m}{4} \cos [(\omega_1 + \omega_2)t + \delta] + \frac{|\Gamma_c| m}{4} \cos [(\omega_1 - \omega_2)t + \delta] \right\} \tag{8}$$

where E_0 is the matched-load peak voltage at the input to the magic tee, ω_1 is 2π times the r.f.-oscillator frequency, ω_2 is 2π times the intermediate frequency, Γ_c is the reflection coefficient of the cavity, m is a factor describing the efficiency of the modulator crystal, and δ is a phase factor dependent on the length of the line between the magic tee and the cavity, the length of line between the tee and the modulator crystal, the characteristics of the modulator crystal, the phase characteristics of the tee, and the phase of the reflection coefficient of the cavity.

The square of the envelope of these waves may be shown to be

$$E_t^2 = \frac{E_0^2}{2}\left\{1 + \frac{|\Gamma_c|^2 m^2}{8} + \frac{|\Gamma_c|\, m}{2}\cos(\omega_2 t + \delta) + \frac{|\Gamma_c|\, m}{2}\cos(\omega_2 t - \delta) + \frac{|\Gamma_c|^2 m^2}{8}\cos(2\omega_2 t)\right\}. \quad (9)$$

It will be shown later that the best operation of the system is obtained when $|\Gamma_c|$ is very small in the region of the resonant frequency of the cavity. Therefore the terms in $|\Gamma_c|^2$ may be neglected, and the envelope is given by

$$E_t \cong \frac{E_0\sqrt{2}}{2}\left\{1 + |\Gamma_c|\, m\cos(\delta)\cos(\omega_2 t)\right\}^{1/2}. \quad (10)$$

This may be expanded by the binomial theorem and terms in $|\Gamma_c|$ to higher than the first power again neglected, giving

$$E_t \cong \frac{E_0\sqrt{2}}{2}\left\{1 + \frac{|\Gamma_c|\, m}{2}\cos(\delta)\cos(\omega_2 t)\right\}. \quad (11)$$

The i.f. voltage at the output of the mixer crystal is, therefore, proportional to

$$E \cong \frac{\sqrt{2}\, E_0 |\Gamma_c|\, m}{4}\cos(\delta)\cos(\omega_2 t). \quad (12)$$

This i.f. voltage is proportional to the imaginary part of the reflection coefficient of the cavity if the variable-line length between the tee and the cavity is so set that the i.f. voltage is zero for a real reflection coefficient. Under this condition (1) may be used in

$$\Gamma_c = \frac{1 - Y_c}{1 + Y_c}$$

to show that the i.f. voltage is proportional to

$$E \cong -\frac{\sqrt{2}\,\alpha a E_0 m}{2[(\alpha+1)^2 + a^2]}\cos(\omega_2 t). \quad (13)$$

The dependence of the amplitude of the i.f. voltage on the oscillator frequency is thus the same as that of the output voltage of the microwave discriminator. The greatest rate of change with frequency is obtained for α equal to unity. This means that the cavity is nonreflecting at resonance, and thus Γ_c is very small compared with unity and the approximations are valid for this condition.

The i.f. signal is amplified in an i.f. amplifier and mixed, in a phase mixer, with a signal derived from the same i.f. oscillator that supplies the modulating voltage. This mixer produces a d.c. voltage proportional to the i.f. voltage, and the d.c. voltage reverses in sign as the r.f.-oscillator frequency is changed from one side of the resonant frequency of the cavity to the other. To obtain the proper sense to be applied as a frequency-control voltage, the phase of the i.f.-oscillator voltage injected into the lock-in mixer may be chosen to be the same as that of the output of the i.f. amplifier for an error in frequency of one sign or of the other. The same effect may be obtained by the choice of the length of the variable length of line in the microwave circuit, since opposite senses are obtained at alternate positions a quarter-wavelength apart.

One very important feature of this stabilization system is that zero i.f. signal is fed into the amplifier when the r.f. oscillator is at the desired frequency. This is true even if the reflection coefficient of the cavity is not zero at resonance. Therefore, very large gain can be used in the amplifier without danger of limiting in either the amplifier or the phase mixer, and thus the phase mixer is linear. In practice, at high gain a signal does appear in the amplifier, and this could cause limiting to occur. Such a signal could be produced from a small inequality in the amplitudes of the two sideband signals arriving at the mixer crystal. If the algebra for that situation is carried through, the i.f. signal from such a cause is found to be orthogonal to the useful signal. For proper setting of the phase of the mixing signal in the lock-in mixer, it thus does not contribute to the d.c. output voltage and therefore does not detune the r.f. oscillator. This spurious signal has not been large enough to cause limiting at the gain found adequate in the systems tried.

A circuit diagram of the i.f. amplifier, phase mixer, i.f. oscillator, and buffer amplifiers used in several stabilization systems of this kind is shown in Fig. 12. Care was taken to shield the i.f. oscillator and buffer amplifiers from the other amplifier. The i.f. amplifier had a pass band about 5 Mc. in width at the half-power points, at a center frequency of 30 Mc. Better phase characteristics for this service could probably be obtained from an amplifier having a specially designed tuned-circuit combination. Stability of the feedback loop is obtained by the use of a capacitance from the reflector of the stabilized 2K25 to ground, to obtain the high-frequency cutoff.

This stabilization system has the advantage that the d.c. level of the lock-in mixer can be made, with suitable insulation, anything required to allow the plate voltage to be used directly as the control voltage of the

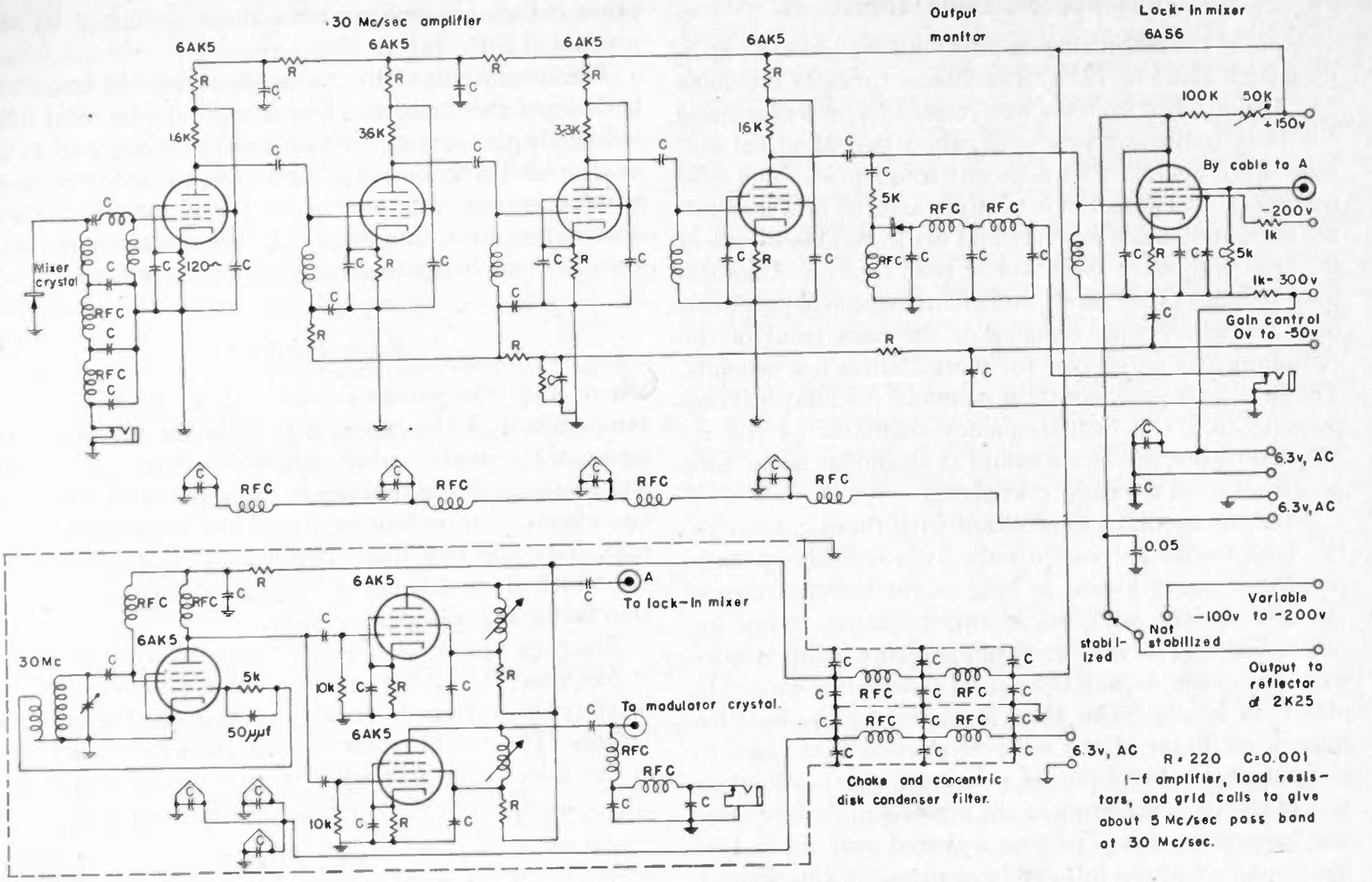

Fig. 12—I.f. amplifier, oscillator, buffers, and phase mixer of the stabilization system.

r.f. oscillator. Thus, oscillators requiring high potentials can easily be stabilized.

If the output voltage is applied to the reflector of a reflex klystron, the cavity frequency controls the oscillator frequency only over the electronic tuning range available at the reflector. As with the d.c. system, however, the output voltage can also be applied to a thermal tuning structure to obtain single-knob tuning over a wider band. Not so wide a band can be accommodated with this system as with the d.c. system, however, because the phase factor δ is determined by the difference in effective length of the two paths to the mixer crystal taken by the direct signal from the oscillator and by the wave reflected from the cavity. Since these two paths cannot be made very nearly the same, the variable-line length must be readjusted to accommodate a large change in frequency.

Degenerative frequency modulation can also be obtained with this system. There are several ways in which the modulating voltage may be introduced. One of particular interest is the application of the modulation voltage as a bias voltage to the mixer crystal through the filter normally used to allow metering of the rectified current. In the absence of the bias voltage the mixer crystal is nonreflecting, but the application of a small bias voltage causes the crystal to reflect. The reflected wave travels, in part, into the modulator crystal, and therefore sideband signals are returned to the mixer crystal. The frequency of the oscillator shifts away from the cavity resonance by an amount sufficient to cancel the i.f. voltage produced by reflection from the mixer crystal.

For the largest frequency shift per unit of bias voltage, the length of line between the mixer crystal and the tee should be chosen to make the i.f. voltage produced by the bias voltage a maximum. A variable length of line could be used here, too, and adjustment of this length could be made to give maximum deviation for a given bias voltage.

The deviation obtained in this way is independent of the amplifier gain and the characteristics of the lock-in mixer and the oscillator. In addition, it is not very dependent on the r.f. power delivered by the oscillator to the stabilization circuit, since the admittance of the mixer crystal is not very dependent on this power at the level of about 1 milliwatt. By measurement of small changes in the deviation produced by a given modulation voltage, small changes in the dissipation of the cavity could be detected. This is the basis of a possible application of this system to the measurement of resonance absorption of microwave energy in certain gases.

Results and Limitations

Most of the stabilizing circuits constructed have operated with 2K25 or 723A/B oscillator tubes in the 9000-Mc. region. The systems were tested by observation of the beat frequency produced when two identical systems were operated on adjacent frequencies. The beat frequency was detected in a mixer and fed into a standard communication receiver and made audible by use of the beat-frequency oscillator of that receiver. Unstabilized oscillators are rarely sufficiently stable to produce a beat frequency that remains in the pass band of the communications receiver for more than a few seconds. The beat frequency contains so many modulation components that the beat-frequency oscillator of the receiver does not produce a sound at all similar to the tone produced from a steady c.w. signal.

With the oscillators stabilized with the d.c. systems, the beat frequency varied only by a few kilocycles in periods of many hours, so long as the temperatures of the two cavities were not changed relative to one another. Cavities having good temperature compensation were not available, and those used changed frequency by about 50 kc./°C. The tone produced by the beat-frequency oscillator of the receiver showed that there remained about 100 c.p.s. of relative frequency modulation of the two oscillators at the power-supply frequency and harmonics of it. The tone wavered over about 1 kc. in a random fashion but rarely changed by this amount in less than a second. This waver is probably caused by the low-frequency noise in the crystal rectifiers. It is found that there is a noise voltage at the output terminals of a crystal detector which is very large compared with the Johnson noise associated with a resistor at room temperature, when the detector is producing a large rectified voltage. This fluctuation is equivalent to a frequency-modulating voltage, and, to account for the waver over about 1 kc., a voltage fluctuation of about 1 millivolt is required. Measurements have shown that this is common with crystals under the conditions of operation in this circuit. The fluctuation is less at higher audio frequencies. It is also less if the rectified voltage is reduced by reduction of the incident power. This means that the most stable oscillator frequency is obtained for the maximum slope in the discriminator characteristic at a given power level. Thus a high-Q cavity and α equal to unity are favored.

With the i.f. systems the waver of the beat frequency was absent, and the tone produced under the best conditions indicated only about 25 c.p.s. of relative frequency modulation at the power-supply frequency and its harmonics. The noise figure of the crystal mixer is, of course, very much less than that of the detectors. There is a limit to the gain which is useful in this system, too, because noise in the output terminals of the crystal mixer and in the i.f. amplifier gives rise to degenerative frequency modulation. Increasing the gain of the system beyond the point at which the residual deviations from other causes are smaller than those produced by this noise is of little value.

The magnitude of the r.m.s. deviations in frequency because of this noise can be calculated. The total noise voltage in the output terminals of the mixer and in the i.f. amplifier may be considered to be caused by a noise-voltage generator connected to the input terminals of a noise-free mixer and amplifier. The mean-square noise voltage of such a generator, when open-circuited, would be

$$\overline{E_n^2} = 4kTNRB$$

where k is Boltzmann's constant, T is the absolute temperature of the laboratory, N is the over-all noise figure of the mixer and i.f. amplifier actually used, R is the characteristic resistance of the wave guide, and B is the effective noise bandwidth of the stabilization circuit. The effective noise bandwidth is approximately the width of the frequency band in which the stabilization factor is greater than unity.

The presence of the noise voltage causes a fluctuation in frequency such that the i.f. voltage developed by the wave reflected from the cavity cancels out the i.f. noise voltage. The combination of r.f. signals on the r.f. side of the mixer crystal is equivalent to an r.f. signal generator having an open-circuit voltage given by

$$E_s = \frac{2\sqrt{2RP_0}\, mQ_0\alpha}{\nu_0(1+\alpha)^2}\, d\nu$$

where P_0 is the power available from the attenuator at the input to the magic tee and $d\nu$ is the difference between the oscillator frequency and the resonant frequency of the cavity. Setting E_s equal to $(\overline{E_n^2})^{1/2}$, the r.m.s. deviation caused by noise is found to be

$$(\overline{d\nu^2})^{1/2} = \left(\frac{kTNB}{2P_0}\right)^{1/2} \frac{(1+\alpha)^2\nu_0}{\alpha mQ_0}. \tag{14}$$

In the experimental systems, N was about 10, B about 10 kc., P_0 about 1 milliwatt, α about unity, m almost equal to unity, Q_0 equal to 25,000, and ν_0 equal to 9000 Mc. These values, used in (14), with kT taken as 4×10^{-21} joules, give 6.5 c.p.s. as the r.m.s. frequency deviation caused by noise. This is somewhat less than the deviations observed at the power-supply frequency and its harmonics. Better filtering of the power-supply voltages and, perhaps, d.c. heater voltages might give some improvement. The power supplies used had about 5 millivolts of ripple per hundred volts.

Applications

There are many uses for stabilized oscillators of this kind. They are very useful for laboratory measurements of very highly resonant circuits because pulling of the frequency of the oscillator by the load circuit is greatly reduced, especially if the oscillator is fed into the stabilization circuit and the test circuit through a magic tee. This reduces the effect of reflections in the test circuit on the power delivered to the stabilization circuit.

The cavity of one of these circuits might be used as a device for measurement of very small thicknesses. A small distortion in the shape of the cavity produces a measurable change in the beat frequency of two stabilized oscillators, and cavities that are very sensitive in this respect could easily be designed. The wavemeters used in the experimental systems changed frequency by 100 c.p.s. for a change in the position of the end plate of 10 angstrom units.

Oscillators having such narrow-band output frequencies as these could be used as carriers for voice communication in narrow frequency channels. If the carrier frequency contains deviations from a discrete frequency of less than 100 c.p.s., as appears to be possible, a frequency modulation producing a deviation of 100 kc. would give a transmitted signal-to-noise ratio of 60 db. A tremendous number of channels wide enough for such signals could be created in a microwave band only a few per cent in width. Another paper by the author will describe a special duplex communication system built up around these frequency-stabilization systems.

These oscillators would also be useful in fundamental research on the interactions between gases and high-frequency fields. There are several gases having quantum mechanical transitions giving rise to resonance absorption in the microwave region. The stabilized oscillators make possible investigation of the details of the structure of these absorption spectra, and ultimately one might use one such absorption line to obtain, with a stabilized oscillator, an absolute standard of frequency.

Synchronization of Oscillators*

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***Summary*—A theory is presented which predicts the behavior of any self-limiting oscillator in the presence of an injected sinusoidal voltage or current of small but constant magnitude. The internal mechanism responsible for synchronization is not needed, and the theory is thus applicable to any source of alternating current. Experimental verification of the theory is presented for the case of a low-power Hartley oscillator operating at 11.5 Mc.**

The theory is extended to include the mutual synchronization of two oscillators of arbitrary properties, and applied to a number of examples to indicate briefly the properties of a synchronized oscillator when used as (a) a linear voltmeter for small voltages, (b) a field-intensity meter, (c) a linear a.m. demodulator for small signals, (d) an f.m. demodulator, and (e) a synchronous amplifier-limiter. The use of a synchronized oscillator is of particular interest because microwave generators can be used in addition to the more conventional triode oscillators.

I. Introduction

THE EARLY EXPERIMENTS of Vincent,[1] followed by Appleton's[2] theoretical treatment, have led to a considerable interest in possible practical applications of the synchronization of oscillators.[3] Since the publication of these early papers, there has been a continually growing literature on the subject, with attention now primarily centered on (a) the use of an oscillator as a synchronous-amplifier limiter for f.m. reception, and (b) the use of a chain of synchronous oscillators to drive a linear accelerator for the production of high-energy atomic particles. There are, of course, numerous other applications, some of which are discussed in the light of the theory which is the subject of this paper.

Following Appleton, theoretical treatments of oscillator synchronization have been concerned with the internal mechanism within a triode oscillator which accounts for synchronization. The phenomenon of synchronization with a disturbance impressed from an external source is not limited to triode oscillators. Rather, any source of alternating e.m.f. whose frequency and amplitude are continuous functions of the load impedance attached to it (the magnetron, for example) will exhibit similar behavior. It should thus be possible to discuss certain general features of synchronization without reference to the internal mechanism which accounts for it. The theory so derived will be generally applicable to all types of oscillators.

In a recent paper Adler[4] has developed a differential equation whose solution accounts for many of the observed phenomena of synchronization. Again, the triode oscillator mechanism has been the basis of the discussion. However, the scheme used by Adler can be extended in a manner which does not involve the particular generator. The result is a differential equation similar to his but more general. In addition, amplitude behavior as well as frequency behavior can be included.

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[1] J. H. Vincent, "On some experiments in which two neighboring maintained oscillatory circuits affect a resonating circuit," *Proc. Roy. Soc.*, vol. 32, part 2, pp. 84–91; 1919–1920.

[2] E. V. Appleton, "The automatic synchronization of triode oscillators," *Proc. Camb. Phil. Soc.*, vol. 21, pp. 231–248; 1922–1923.

[3] The term "oscillator" as used here means a source of harmonic vibration whose steady-state amplitude is limited to a finite value by some internal nonlinear characteristic.

[4] Robert Adler, "A study of locking phenomena in oscillators," Proc. I.R.E., vol. 34, pp. 351–357; June, 1946.

The performance of the oscillator is specified in terms of a set of compliance coefficients which show how amplitude and frequency depend upon the load impedance. The values of the coefficients are not derived here but are assumed to be given as constants of the problem. They may be derived theoretically or measured for the particular oscillator.

The injected voltage is considered as equivalent to the IZ drop on a fictitious increment in the load impedance. The oscillator's frequency and amplitude shift in accordance with its compliance coefficients and the magnitude and phase of the incremental load impedance. If the disturbance due to the injected voltage is small and its frequency is close to that of the oscillator, replacing the actual voltage by a fictitious impedance of varying phase and magnitude is valid and the synchronization behavior can be calculated.

II. Synchronization by an Impressed Voltage

In the discussion to follow, complex quantities will be represented by boldface italic characters; quantities not so designated will denote absolute magnitudes. The factor $e^{j\omega t}$ will usually be omitted.

A. Compliance Coefficients

Let Fig. 1 represent an energy source of the type which converts d.c. energy to a.c. energy, such as a typical triode oscillator or magnetron. We will be interested in two pairs of terminals. Those marked E–E are the output terminals of the device for delivering a.c.

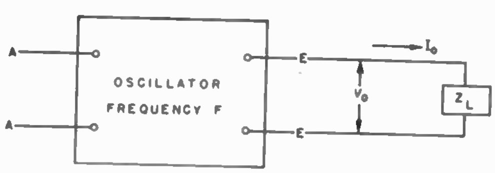

Fig. 1—Oscillator for synchronization studies.

power to a load impedance Z_L. The terminals A–A represent any pair of terminals which give a d.c. or a.c. indication of the amplitude of oscillation, such as grid bias or d.c. plate current.

Assume that there are also available, when necessary, instruments which indicate either the voltage V_0 across the load or current I_0 through it. Let $\boldsymbol{V}_0$ and $\boldsymbol{I}_0$ be the initial values of these quantities when the oscillator is feeding its load circuit. Similarly, let F represent the frequency of the oscillator, and F_0 its undisturbed value. When a small impedance $\boldsymbol{z}$ is added to the load, the frequency and amplitude change. The compliance coefficients are defined in terms of these changes; thus,

$$A_r = \left.\frac{\partial A}{\partial r}\right|_{z=0} \qquad A_x = -\left.\frac{\partial A}{\partial x}\right|_{z=0} \tag{1}$$

$$F_r = \left.\frac{\partial F}{\partial r}\right|_{z=0} \qquad F_x = -\left.\frac{\partial F}{\partial x}\right|_{z=0} \tag{2}$$

where

$$\boldsymbol{z} = r + jx. \tag{3}$$

The negative sign in A_x and F_x is incorporated here for reasons of symmetry in later expressions.

A and F are expanded in a Taylor series about A_0 and F_0, keeping only first-order terms. This gives

$$A - A_0 = rA_r - xA_x \tag{4}$$

$$F - F_0 = rF_r - xF_x. \tag{5}$$

Complex compliance coefficients for amplitude, $\boldsymbol{C}_A$, and frequency, $\boldsymbol{C}_F$, will be needed. These are

$$\boldsymbol{C}_A = C_A e^{j\alpha} = A_r + jA_x = \sqrt{A_r^2 + A_x^2}\, e^{j\alpha} \tag{6}$$

and

$$\boldsymbol{C}_F = C_F e^{j\beta} = F_r + jF_x = \sqrt{F_r^2 + F_x^2}\, e^{j\beta}. \tag{7}$$

B. Synchronization Equation

Let a small voltage be induced in the load circuit from an outside source. Assume the voltage is small enough so that the change in I can be neglected and we can, with sufficient accuracy, represent I by its initial value I_0. We replace the induced voltage $\boldsymbol{v}$ by a small impedance z where

$$\boldsymbol{z} = \frac{v}{I_0} e^{j\phi}. \tag{8}$$

We may thus write, with the aid of (4) and (5) (keeping only real parts),

$$A - A_0 = \boldsymbol{C_A z} = \frac{C_A v}{I_0} \cos(\phi + \alpha) \tag{9}$$

and

$$F - F_0 = \boldsymbol{C_F z} = \frac{C_F v}{I_0} \cos(\phi + \beta) \tag{10}$$

where

$$\tan\alpha = \frac{A_x}{A_r}, \qquad \tan\beta = \frac{F_x}{F_r}.$$

If the injected voltage $\boldsymbol{v}$ has the frequency F' and the instantaneous frequency of the oscillator is F, we can write

$$\frac{1}{2\pi}\frac{d\phi}{dt} = F' - F = (F' - F_0) - (F - F_0) \tag{11}$$

and

$$\boldsymbol{z} = \frac{v}{I_0} e^{j2\pi(F'-F)t} = \frac{v}{I_0} e^{j\phi(t)}. \tag{12}$$

If $F' - F$ is not too large, the oscillator will follow the impedance changes as shown by (9) and (10). In particular, (10) gives

$$\frac{1}{2\pi}\frac{d\phi}{dt} = (F' - F_0) - \frac{C_F v}{I_0}\cos(\phi + \beta), \qquad (13)$$

a differential equation similar to that derived by Adler which shows how the beat frequency, if any, varies with time.

Putting

$$F' - F_0 = f$$

and

$$\frac{C_F v}{I_0} = Kv$$

into (13) yields

$$\frac{1}{2\pi}\frac{d\phi}{dt} = f - Kv\cos(\phi + \beta). \qquad (14)$$

It is immediately evident from (14) that the solution $\phi(t)$ is of a complicated periodic form when

$$f^2 > K^2v^2 \qquad (15)$$

and reduces exponentially to a steady value of ϕ when

$$f^2 < K^2v^2. \qquad (16)$$

Condition (16) corresponds to synchronization between the injected voltage and the oscillator current at a fixed phase angle ϕ. Since we are interested primarily in synchronization, the solution of (14) subject to (16) is needed. It is

$$\frac{\cos\psi - \cos(\phi + \beta)}{1 - \cos(\phi + \beta - \psi)} = \text{const. } e^{-2\pi t\sqrt{K^2v^2 - f^2}} \qquad (17)$$

where

$$\cos\psi = \frac{f}{Kv}\cdot$$

The steady-state value of ϕ for large t is given by

$$\cos(\phi + \beta) = \frac{f}{Kv}\cdot \qquad (18)$$

The equilibrium value is approached in such a manner that the time constant is approximately

$$\tau \cong \frac{1}{2\pi\sqrt{K^2v^2 - f^2}} = \frac{1}{2\pi Kv\sin\psi}\cdot \qquad (19)$$

There are two values of $(\phi+\beta)$ which satisfy (18). One corresponds to stable equilibrium; the other to unstable equilibrium. From (14),

$$\frac{1}{2\pi}\frac{d}{d\phi}\left(\frac{d\phi}{dt}\right) = Kv\sin(\phi + \beta). \qquad (20)$$

For stability,

$$\frac{d}{d\phi}\left(\frac{d\phi}{dt}\right)$$

must be negative. Thus only values of $(\phi+\beta)$ such that $\sin(\phi+\beta)$ is negative lead to stable synchronization.

Equation (18) shows that synchronization can be obtained over a range of f such that

$$-Kv < f < Kv,$$

or, over a band of frequencies,

$$\Delta f = 2Kv. \qquad (21)$$

C. Amplitude Changes

The quantity $a = A - A_0$ expresses the change of some convenient amplitude parameter, such as plate current, in the presence of an injected voltage. It is evident from (9) and (10) that a and f are functionally related through the parameter ϕ. By defining new quantities

$$\begin{aligned}\delta &= (\phi + \beta)\\ \rho &= (\alpha - \beta)\end{aligned} \qquad (22)$$

we can write (9) and (10) in terms of dimensionless variables U, W,

$$U = \frac{fI_0}{C_F v} = \cos\delta \qquad (23)$$

$$W = \frac{aI_0}{C_A v} = \cos(\delta + \rho), \qquad (24)$$

from which it is evident that the form of the functional relation between a and f is independent of I_0, v, C_A, and C_F for small disturbances. Elimination of δ in (23) and (24) leads to the equation for an ellipse in the U, W plane, which degenerates to a line when $\rho = 0$ or π and into a circle when $\rho = \pm\pi/2$.

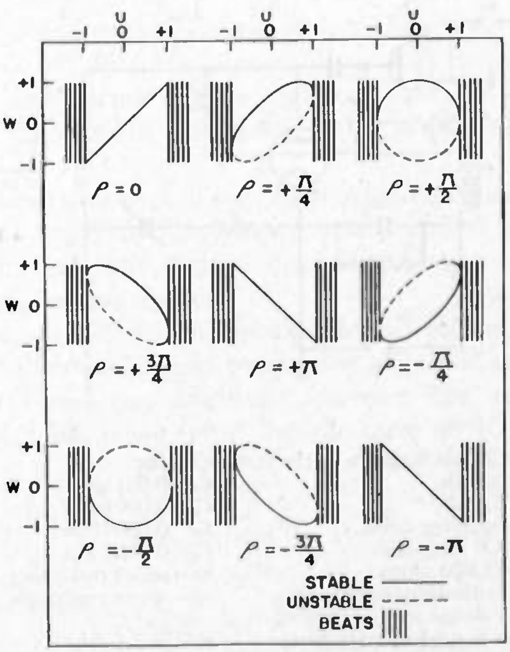

Fig. 2—Forms of the U-W curve in the region of synchronization.

Fig. 2 shows the U–W curves for several typical values of ρ. In most cases the frequency of an oscillator depends more upon the reactance of the load than upon its resistance. Thus ρ will generally be nearly $-\pi/2$ and the U–W curve almost a semicircle. In the figure the broken line shows the condition of unstable equilibrium, the solid line shows stable equilibrium, and the vertical lines indicate the region of beats outside the synchronization band.

It is important to note that, no matter what the value of ρ, the maximum absolute value of a is the same and is given by

$$a_{\max} = \frac{C_A v}{I_0}. \tag{25}$$

Thus, if the frequency of the injected voltage is swept across the synchronization band of the oscillator, there will be a pulse of voltage or current (depending upon the quantity represented by a) whose peak value is independent of ρ.

From (21), (23), (24), and (25), we see that

$$\frac{\Delta f}{a_{\max}} = \frac{2C_F}{C_A}, \tag{26}$$

which shows that the synchronization bandwidth is proportional to $a_{\max}$, or, from (25), to the injected voltage v. It is often convenient to use $a_{\max}$ as a measure of v without measuring v. The bandwidth of synchronization can then be predicted directly from (26).

III. Experimental Measurements

In order to check the foregoing theory, experimental measurements were made on a small Hartley oscillator operating at 11.5 Mc. R.f. voltage for injection was supplied by a push-pull power oscillator operating at ten times the plate voltage of the small oscillator and very loosely coupled to it inductively.

Fig. 3 is a circuit diagram of the test oscillator showing the method of voltage injection and a diode for measuring r.f. plate swing. It will be noted that the plate coil has been used for the load Z_L and that the synchronizing voltage is induced in it.

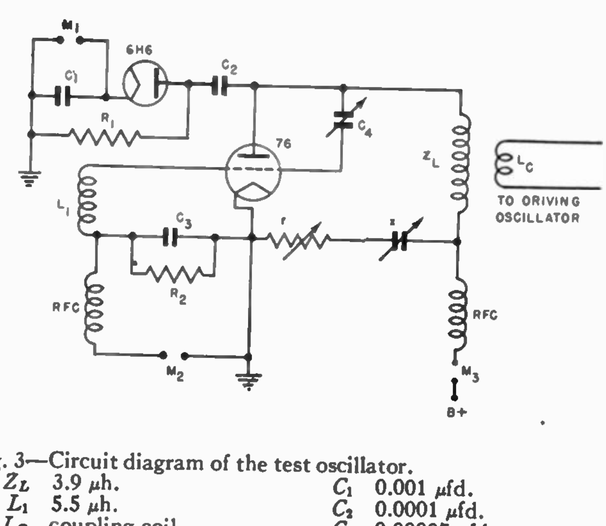

Fig. 3—Circuit diagram of the test oscillator.

Z_L	3.9 μh.	C_1	0.001 μfd.
L_1	5.5 μh.	C_2	0.0001 μfd.
Lc	coupling coil	C_3	0.00007 μfd.
R_1	1.0 megohm	C_4	0.0002 μfd.
R_2	15,000 ohms		r = series resistance
M_1	radio-frequency plate-voltage meter		x = series reactance
M_2	d.c. grid-voltage meter		RFC = 2.5 mh.
M_3	plate-current meter		

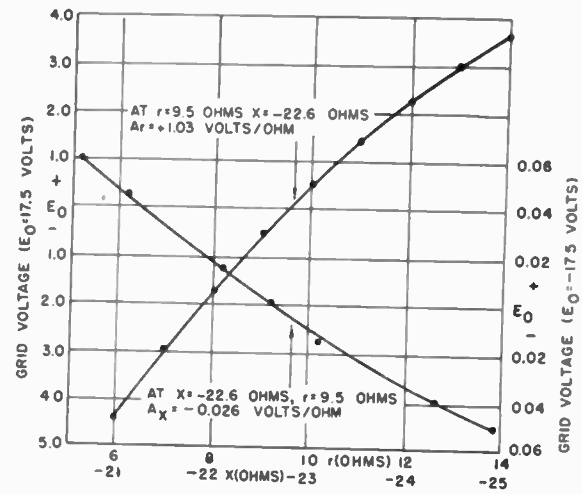

Fig. 4—Experimental curves for evaluation of A_r and A_x. (Use left ordinates for A_r, right ordinates for A_x.)

The compliance coefficients were measured by inserting capacitors x and resistors r in series with the plate tank coil. To allow measurement on both sides of the operating point, this point was specified to be $r=9.5$ ohms, $x=-22.6$ ohms.

Fig. 4 shows the experimental curves from which A_r and A_x can be obtained. From them we observe that the compliance coefficient C_A has the value

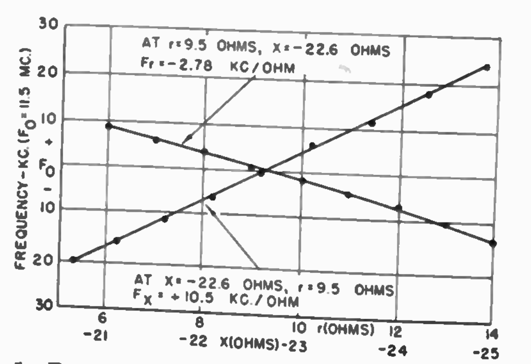

Fig. 5—Experimental curves for evaluation of F_r and F_x.

$$C_A = 1.03 \text{ volts/ohm}$$

$$\alpha = -1.5 \text{ degrees.}$$

Fig. 5 shows similar curves for the evaluation of C_F. The appropriate values are

$$C_F = 10.8 \text{ kc./ohm}$$

$$\beta = +105 \text{ degrees.}$$

Calculation of the expected bandwidth of synchronization from these values gives

$$\frac{\Delta f}{a_{\max}} = \frac{2C_P}{C_A} = 21 \text{ kc./volt.}$$

Fig. 6 shows the experimental curve of bandwidth of synchronization as a function of $a_{\max}$. The slope of the curve at the origin is 20.6 kc./volt, in good agreement with the expected value. Note also that the curve is linear over the range of voltages used.

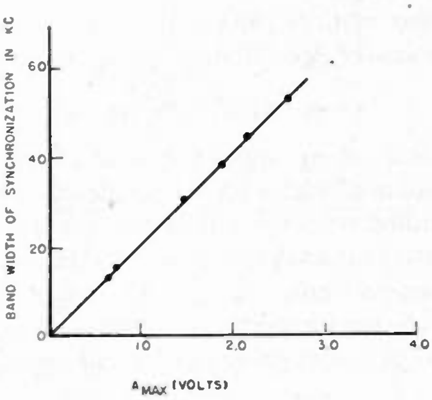

Fig. 6—Experimental determination of bandwidth of synchronization in terms of injected voltage as measured by $a_{\max}$. The slope of the curve is 20.6 kc./volt.

Fig. 7 shows the result of an experimental measurement of the relation between $a_{\max}$ and v. We note again that the relation is linear over the range investigated.

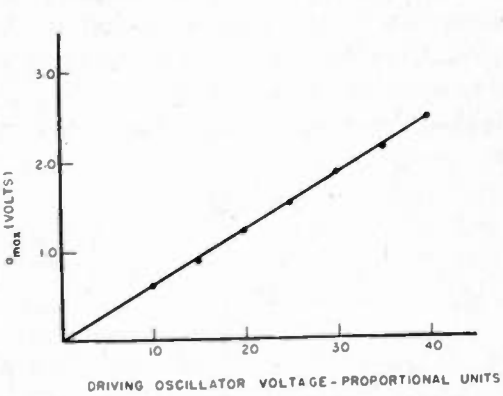

Fig. 7—Relation between $a_{\max}$ and injected voltage.

For this oscillator $\rho = -106.5$ degrees, and the U–W curve should be nearly a semicircle. Fig. 8 shows the exact form of the U–W curve for $\rho = -106.5$ degrees (solid line) and the measured curve when sweeping the power oscillator from high to low frequency across the band (solid dots). To check for possible hysteresis the curve was measured again, sweeping from low to high frequency, with results shown by crosses. The injected signal was increased from $a_{\max} = 0.92$ volts to $a_{\max} = 4.9$ volts and the curves were repeated to observe the effect of a large signal. Results are given by triangles (low to high frequency) and circles (high to low frequency). There is no evidence of hysteresis, although its presence has been mentioned in Appleton's studies.

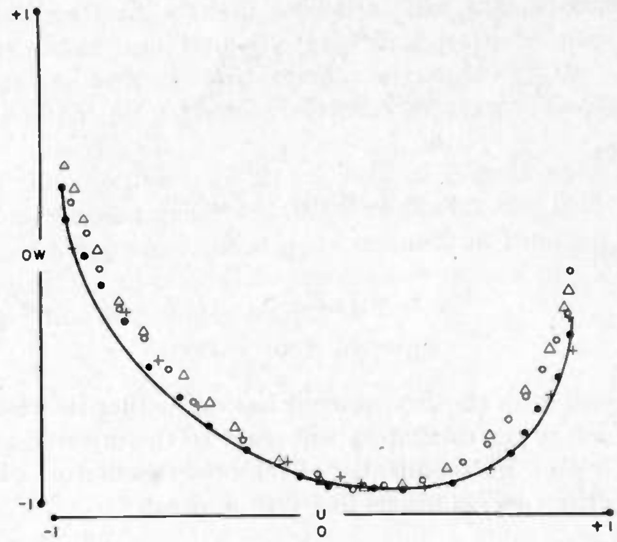

Fig. 8—Experimental and theoretical U-W curves for the test oscillator. Solid line: theoretical curve for $\rho = -106.5°$).

$a_{\max}$ 0.92 volts: Solid dots: experimental values sweeping from high to low frequency. Crosses: experimental values sweeping from low to high frequency.

$a_{\max}$ 4.9 volts: Circles: high to low frequency. Triangles: low to high frequency.

IV. Mutual Synchronization of Two Oscillators

Consider two oscillators of the form shown in Fig. 1 and let them be coupled by a mutual impedance

$$\boldsymbol{Z}_{12} = Z_{12}e^{j\phi_{12}}. \tag{27}$$

Let the two systems to be identified by subscripts 1 and 2. The coupling is assumed to be arranged so that the coupled voltages are induced in the load impedances Z_L of each system.

Both Z_{12} and ϕ_{12} will, in general, be functions of frequency. To simplify the present discussion, we assume that this dependence can be neglected over the narrow range of frequencies covered by the synchronization band.

Since we are interested only in synchronization we assume that both oscillators are synchronized at frequency F, and that their undisturbed frequencies are F_{01} and F_{02}, respectively.

In order to specify phases we refer all phases to the current $\boldsymbol{I}_{01}$ in the load of oscillator 1. We will seek the value of the phase angle θ_{12} between the currents $\boldsymbol{I}_{01}$ and $\boldsymbol{I}_{02}$. We write (omitting the term $e^{j2\pi Ft}$)

$$\begin{aligned} \boldsymbol{I}_{01} &= I_{01}e^{j0} \\ \boldsymbol{I}_{02} &= I_{02}e^{j\theta_{12}} \\ \boldsymbol{v}_1 &= v_1e^{j\phi_1} \\ \boldsymbol{v}_2 &= v_2e^{j(\theta_{12}+\phi_2)} \\ \boldsymbol{Z}_{12} &= Z_{12}e^{j\phi_{12}}. \end{aligned} \tag{28}$$

Now,

$$v_1 = I_{02}Z_{12} = I_{02}Z_{12}e^{j(\theta_{12}+\phi_{12})}, \tag{29}$$

whence

$$\begin{aligned} v_1 &= I_{02}Z_{12} \\ \theta_{12} + \phi_{12} &= \phi_1 + 2n\pi. \end{aligned} \tag{30}$$

Also,

$$v_2 = I_{01}Z_{12} = I_{01}Z_{12}e^{j\phi_{12}}, \tag{31}$$

whence

$$\begin{aligned} v_2 &= I_{01}Z_{12} \\ \phi_{12} &= \theta_{12} + \phi_2 + 2n\pi. \end{aligned} \tag{32}$$

We will drop the $2n\pi$, since it has no further interest.

Each of the oscillators will react to the injected voltage it sees, independently of the other oscillator. Thus, we write two equations like (10) and get

$$F - F_{10} = \frac{C_{F1}v_1}{I_{01}} \cos(\phi_1 + \beta_1) \tag{33}$$

$$F - F_{20} = \frac{C_{F2}v_2}{I_{02}} \cos(\phi_2 + \beta_2). \tag{34}$$

These we can combine, with the aid of (29), (30), (31), and (32), to get

$$F_{20} - F_{10} = \frac{C_{F1}v_1}{I_{01}} \left[\cos(\theta_{12} + \phi_{12} + \beta_1) - \frac{C_{F2}}{C_{F1}}\left(\frac{I_{01}}{I_{02}}\right)^2 \cos(-\theta_{12} + \phi_{12} + \beta_2) \right] \tag{35}$$

which is an equation involving θ_{12} as the only unknown.

We observe immediately from (35) that both oscillators contribute to the bandwidth of synchronization. To see the effect more clearly, we write

$$\begin{aligned} \Phi &= (\theta_{12} + \phi_{12} + \beta_1) \\ \mathcal{E}_1 &= -(B_1 + B_2 + 2\phi_{12}) \\ k &= \frac{C_{F2}}{C_{F1}}\left(\frac{I_{01}}{I_{02}}\right)^2 \end{aligned} \tag{36}$$

and get

$$F_{20} - F_{10} = \frac{C_{F1}v_1}{I_{01}} \left[\sqrt{1 + k^2 - 2k\cos\mathcal{E}_1}\,\cos(\Phi + \mathcal{E}_2)\right] \tag{37}$$

where

$$\tan \mathcal{E}_2 = \frac{-k\sin\mathcal{E}_1}{1 - k\cos\mathcal{E}_1}.$$

From (37) we see that the two oscillators synchronize over a band of frequencies Δf_{12}, given by

$$\Delta f_{12} = \Delta f_1\sqrt{1 + k^2 - 2k\cos\mathcal{E}_1}. \tag{38}$$

If oscillator 2 is much more powerful than oscillator 1 but otherwise identical, k will be very small and Δf_{12} becomes equal to Δf_1.

From this it can be seen that it is important to have the driving oscillator more powerful than the test oscillator when making synchronization measurements. If the two are identical, k will be 1 and the band of synchronization can vary from 0 to $2\Delta f_1$, depending on $\mathcal{E}_1$.

The allowed values of Φ and hence of θ_{12} can be obtained from (35), (36), and (37) when the necessary parameters are given.

In a similar manner the equations can be extended to include the case of N oscillators acting upon one another.

V. Applications

Several interesting applications of the synchronized oscillator, some of which have been described elsewhere, may be studied with the aid of this theory. In the following no attempt has been made to make an exhaustive study of any particular application, but rather to indicate as a basis for further investigation some interesting applications of the synchronized oscillator.

A. Linear R.F. Voltmeter

It has been shown that a_{max} is proportional to v and independent of C_F, α, or β, and therefore a synchronized oscillator can be used as a linear votmeter giving a d.c. indication of the amplitude of the injected a.c. voltage. The use of a synchronized oscillator provides a linear voltmeter for small voltages at any frequency for which an oscillator can be constructed, including the microwave region, since the treatment is not confined to triodes and lumped circuit elements.

If V is the r.f. voltage (peak) on the load impedance Z_L, then

$$V = I_0Z_L$$

and

$$a_{max} = C_AZ_L\frac{v}{V}. \tag{39}$$

Typical values measured on the experimental oscillator are $C_A = 1.03$ volts/ohm, $Z_L = 236$ ohms, and $V = 47$ volts. Whence

$$a_{max} = 5.2v,$$

indicating that this oscillator-voltmeter gives about a five-fold amplification of the voltage to be measured.

If the oscillator is properly designed it will be found that, to a good approximation,

$$C_A = SV$$

where S is a constant of proportionality. We can then write

$$a_{max} = SZ_Lv$$

and note that a_{max} is independent of V, or that the calibration of the voltmeter is independent of the power-supply voltage driving it. To demonstrate this, the test oscillator was used to measure a fixed injected voltage while its own power-supply voltage was varied from 105 to 225 volts. The result is shown in Fig. 9. By careful design the dependence on power-supply voltage can be further reduced. One of our test oscillators showed no measurable change in reading.

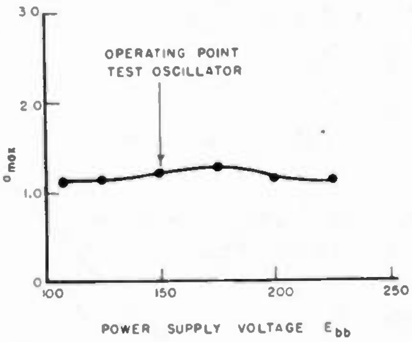

Fig. 9—a_{max} as a function of test oscillator E_{bb} for fixed injected voltage.

While we have been concerned with the behavior of the oscillator under small disturbances, we have seen that the device is linear for a_{max} up to 2.5 volts and therefore for injected voltage of 0.5 volt. Higher voltages can be handled by more powerful oscillators, but it must be remembered (as seen in Section IV) that the source of the voltage to be measured must have a higher power than the test oscillator to avoid complications.

If the frequency of the injected voltage cannot be varied across the synchronization band of the voltmeter, the frequency of the voltmeter can be varied across the synchronization band by a small variable capacitor. The d.c. grid bias or other amplitude indicator can be coupled through a blocking capacitor to a peak voltmeter. As the voltmeter-oscillator is wobbled back and forth across the frequency of the injected voltage to be measured, a pulse will be observed whose peak value is a_{max}. From this pulse the size of the injected voltage can be calculated.

It should also be noted that the synchronized oscillator can be used, as done by Appleton,[2] to measure small voltages by determining the bandwidth of synchronization, which is also linearly related to v by the relation

$$\Delta f = 2C_P \frac{vZ_L}{V} \cdot$$

B. Field-Intensity Meter

The voltmeter properties of the synchronized oscillator lend themselves nicely to the measurement of field intensity at any frequency for which an oscillator is available. Appleton used the synchronization bandwidth of an oscillator to measure field intensities. It is proposed here to use the voltage changes directly, instead of the synchronization band, largely because the power-supply variation no longer enters the calculation and frequency measurements are not needed.

Assume that a small oscillator, like that of Fig. 1, is available and that the grid bias is to be used as indicating voltage. An antenna is coupled to the load Z_L so that its radiation resistance appears as R_S in that load circuit.

If the antenna is in an r.f. field of E peak volts per meter, whose strength is to be measured, the field will induce a voltage v (as already defined) in the load impedance Z_L of which the antenna is now a part. The magnitude of v can be shown to be

$$v = \frac{\lambda}{\pi} E \sqrt{\frac{R_s G}{120}}\, f(\theta) \tag{40}$$

where G is the gain referred to an isotropic radiator and $f(\theta)$ is the normalized electric-field radiation pattern of the antenna.

C_A and S should be measured about an operating load including the R_s of the antenna used. If a tuning capacitor in the oscillator is wobbled back and forth through the synchronization region, a pulse of peak value a_{max} will be observed, as in the case of the voltmeter. From its magnitude the strength of the field E can be calculated. It will be

$$E = \frac{\pi}{\lambda} \frac{a_{max}}{SZ_L} \sqrt{\frac{120}{R_s G}} \cdot \tag{41}$$

The sensitivity of the field-strength meter will decrease with decreasing λ, but at the higher frequencies an increased gain G can be used to compensate for the loss.

C. Linear A.M. Detector

The synchronized oscillator can be used as a linear demodulator for amplitude modulation by taking the intelligence-frequency component from the A terminals. In this application it will be best to use an oscillator which has $\rho \cong \mp \pi/2$ so that the U–W curve is nearly a semicircle. This will be true if the signal is injected into the plate or grid circuit of a class-C oscillator, and the output is read from the d.c. grid bias. It will be necessary to have sufficient signal strength so that the synchronization band will include all the sideband frequencies.

To achieve this it appears reasonable to require that the time constant of the device be short compared to the shortest period of the modulation to be received. We have seen in (19) that the time constant is approximately

$$\tau = \frac{1}{2\pi\sqrt{\left(\frac{\Delta f}{2}\right)^2 - f^2}} = \frac{1}{\pi \Delta f \sin \psi} \cdot \tag{42}$$

Sin ψ is unity near the center of lock-in where f is nearly zero. Thus the requirement that τ be short compared to $1/f_{max}$, where f_{max} is the highest modulation frequency to be reproduced, means that

$$\frac{1}{\pi\Delta f} \ll \frac{1}{f_{max}}$$

or that

$$\pi\Delta f \gg f_{max}. \tag{43}$$

From this we conclude that the signal used at the demodulator must be large enough to give a synchronization bandwidth of at least 30 kc. in order to give faithful reproduction of 10 kc. modulation.

If the synchronized demodulator is used it will have the advantage not only of linearity, but it can also give a demodulation voltage amplification, as shown in (39) *et seq.*

When nearly 100 per cent modulation is used the device will lead to distortion of a peculiar form because synchronization may be lost when the signal is small near the peak of modulation. However, the synchronized oscillator-demodulator appears to present interesting possibilities worthy of further investigation.

D. F.M. Discriminator

If the oscillator circuit is arranged so that $\rho=0$ or π, the synchronized oscillator can be used as an f.m. discriminator-demodulator. Reference to Fig. 2 shows that, under these conditions, the U–W curve is a straight line with $U=0$ at center frequency.

One way of achieving this is to couple an auxiliary resonant circuit to the test oscillator and inject the synchronizing signal into this auxiliary circuit. The output can be taken from the d.c. grid bias of the oscillator or from a diode connected across the resonant circuit. Fig. 10 shows the auxiliary resonant circuit and the coupling to the driving oscillator used in the experimental tests.

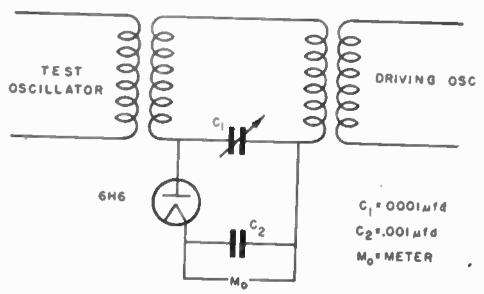

Fig. 10—Auxiliary resonant circuit to obtain behavior characteristic of $\rho=0$.

If the resonant circuit is properly detuned (about 70 per cent of resonance voltage), resistance and/or reactance added in the auxiliary circuit appear as reactance and/or resistance in the oscillator load circuit. If in the original oscillator $\rho=\pm\pi/2$, it will appear to be $\rho=0$ or π when the auxiliary circuit is added and the desired result is attained.

Fig. 11 shows three U–W curves taken from the diode across the resonant circuit; for exact resonance of the diode circuit, 100 per cent; detuned to 70 per cent of resonant voltage, and detuned to an intermediate value, 90 per cent. The linear U–W curve desired was achieved at the 70 per cent detuning adjustment.

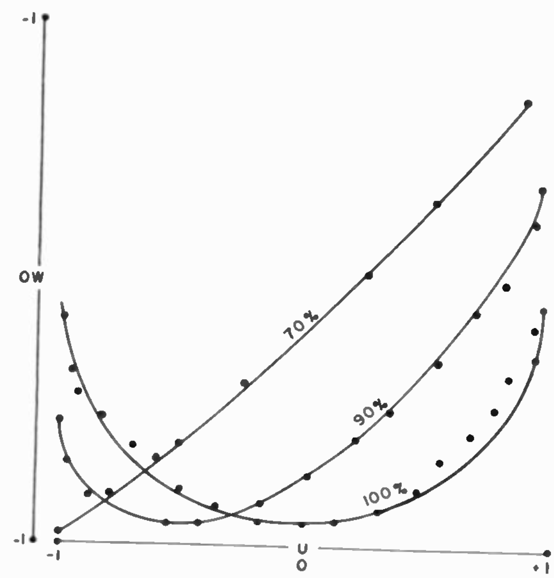

Fig. 11—Experimental U–W curves taken from auxiliary resonant circuit. (Percentages refer to resonant voltage on auxiliary circuit.)

When the arrangement described above is used as a discriminator, a will be zero at the center frequency, and the device is thus insensitive to amplitude modulation in a manner similar to a balanced discriminator.

E. F.M. Synchronous Amplifier-Limiter

In this application the oscillator is locked to an f.m. signal. It follows the frequency variations without serious amplitude change, and hence becomes a combined amplifier and limiter. It has been discussed previously in the literature.[5]

If the synchronized oscillator is capable of following the frequency deviations, the response to an f.m. signal of the form

$$f(t) = f_0 \sin 2\pi f_m t \tag{44}$$

will be a solution of

$$\frac{1}{2\pi}\frac{d\phi}{dt} + \frac{C_F v(t)}{I_0}\cos(\phi+\beta) = f(t) \tag{45}$$

where $v(t)$ represents any amplitude modulation of v that may be present. Direct integration of (45) is compli-

[5] C. W. Carnahan and H. P. Kalmus, "Synchronized oscillators as frequency-modulation receiver limiters," *Electronics*, vol. 17, pp. 108–112; August, 1944.

cated and need not be performed to the approximation needed here. It will be recalled that ϕ responds to changes in f and v with a time constant τ given by (19). If the changes in f or v occur in a time long compared with τ, the oscillator is essentially in equilibrium at each instant, and a succession of steady-state solutions for various fixed f is a good-enough approximation to the actual solution for varying f. If the injected voltage v is always so large that

$$Kv = \eta f_0; \qquad \eta > 1, \tag{46}$$

then

$$\tau \leqq \frac{1}{2\pi f_0 \sqrt{\eta^2 - 1}}: \tag{47}$$

Since it is standard practice to have $f_0 > 5f_m$, it is evident that the time constant is short compared with the frequency-modulation period $1/f_m$ and the equilibrium solution (18) is a reasonable approximation. Similar arguments hold for changes in v, but we are not not interested in amplitude modulation here, and will henceforth assume v to be constant.

We see from (18) that changes in f will produce changes in ϕ, so that an additional phase modulation will be added to the impressed signal. This implies that $d\phi/dt \neq 0$, in contradiction to the original assumptions made in solving (14) to get (18). The correction will be small if the frequency variations are slow, and it can be shown that the distortion will be negligible if $Kv \geqq 2f_0$.

Amplitude changes in v will also lead to phase modulation. However, if Kv is kept somewhat larger than f_0, the phase is relatively insensitive to voltage changes and the distortion arising from amplitude modulation is thereby minimized.

If the criterion $Kv = 2f_0$ is set as a design center, the voltage amplification achieved by the use of the synchronized oscillator will be

$$\frac{V}{v} = \frac{C_F Z_L}{2f_0}. \tag{48}$$

To estimate the order of magnitude of the gain that may safely be used, assume (a) that a single LC circuit is controlling the oscillator, (b) that the voltage V is the one across the entire inductance of the oscillating circuit, and (c) that v is injected into this inductance. Then

$$C_F = \frac{1}{4\pi L}. \tag{49}$$

This gives

$$\frac{V}{v} \quad \frac{F}{4f_0}.$$

If the oscillator frequency is 10 Mc. and f_0 is 100 kc. the maximum voltage amplification of the device can be about 25. Of course, if the voltage across a part of the tank inductance is used, as in the experiments already described, the gain is correspondingly reduced. However, gains of 10 or more should be readily obtainable. Also, the voltage v may be injected into the grid circuit of the oscillator and the gain of the tube used to increase the voltage seen in the tank circuit. Equation (49) refers only to the voltage v injected into the oscillating circuit which controls the frequency.

Unless there is some amplitude-regulating device on the synchronized oscillator, there will also be an amplitude modulation in its output. The magnitude of the effect can be calculated from (9) and (24). It is usually small enough to be neglected.

VI. Conclusion

Although the theory and experiments just described have been discussed in terms of a conventional self-limiting source of alternating e.m.f., the concepts involved are quite general, and with appropriate redefinition of symbols the equations can apply equally well to any source of harmonic disturbance, electrical, electromagnetic, mechanical, or acoustical, singly or in combination. The self-limitation implies that some nonlinear element is present to limit the amplitude of oscillation. A truly linear oscillator will not exhibit synchronization effects. It will, however, not appear in practice.

We may then conclude that any source of harmonic disturbance whose steady-state frequency is a continuous function of the load applied to it, and whose frequency can shift with sufficient rapidity, will exhibit synchronization behavior when a harmonic disturbance is impressed upon it from an external source. If the amplitude of the device is also a function of its load, then it will exhibit a characteristic amplitude variation in the synchronization region. We may also conclude that the source will synchronize with an impressed disturbance, however small, if the frequency of the outside disturbance is close enough to that of the undisturbed source.

It is important to remember that the properties of the external source, supplying the synchronizing signal and the coupling impedance, are important in determining the bandwidth and phase of synchronization. If there is a considerable disparity in the power output of the two sources, the weaker determines the synchronization properties of the system. If the power outputs are nearly equal, both sources contribute almost equally to the synchronization properties.

VII. Acknowledgment

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Reflex Oscillators for Radar Systems*

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Summary—**The advantages to be gained in the operation of radar systems at very high frequencies have led to the use of frequencies of several thousand megacycles. Operation at these frequencies has imposed serious problems in obtaining suitable tube behavior. Because of the difficulty in obtaining amplification at the transmission frequency, the r.f. section of the usual radar receiver consists of a crystal converter driven by a beating oscillator and operating directly into an i.f. amplifier. Since the midband frequency of the latter has commonly been either 30 or 60 Mc., it has been necessary to provide beating oscillators operating at frequencies differing from those of the transmitter by only a few per cent.**

For radar systems intended to operate at approximately 3000 Mc., which were under development in the early days of the war, it was found that triodes then available gave unsatisfactory performance. Attention shifted to the possibility of using velocity-modulated tubes, and the particular form known as the reflex oscillator came into general use.

In this paper the requirements on beating-oscillator tubes for radar systems will be discussed, and the design features which have made the reflex oscillator eminently satisfactory in this application will be pointed out. Problems encountered in such oscillators will be outlined, and the solution in a number of cases is indicated. In some instances military requirements and expediency were in conflict with the optimum performance, and hence certain compromises were necessary.

Requirements on a Beating Oscillator for Radar Systems

Power Output

Vacuum-tube converters, requiring about 100 milliwatts of beating-oscillator power, were early abandoned in favor of crystal converters. Although the power required by a crystal converter for good performance is about 1 milliwatt, attenuation of the order of 13 db is commonly inserted between the oscillator and the crystal to reduce the effect of variations of the load on the oscillator. For this reason it is generally desirable to provide a minimum beating-oscillator power of about 20 milliwatts.

Frequently, tubes designed as beating oscillators are also used as signal generators for field and laboratory test equipment. For these purposes it appeared that a power output of 50 milliwatts would meet most needs.

Low-Voltage Requirements

Early velocity-modulated tubes operated at high voltages, with all the attendant disadvantages. The advantages of being able to operate the beating oscillator from the same voltage sources as the i.f. amplifier were apparent, and accordingly it became the practice whenever practical to design reflex oscillators for 300-volt operation.

Tuning

The problems of frequency stability encountered in radar systems operating at several thousand megacycles are quite different from those in radio communication transmission and reception. In existing radar transmitters a frequency shift may be caused by thermal expansion of the resonant elements of the magnetron cavity. In addition, frequency pulling may result from the presentation to the magnetron of load variations which are caused by imperfect matching of rotating or moving joints. At a frequency of 10,000 Mc., for example, the expected frequency shift caused by temperature effects on the magnetron may be of the order of 20 Mc. per 100° C. change in resonator temperature. To this, 5 or 10 Mc. may be added for frequency pulling by variations in the load. Since the pass band of the i.f. amplifier may be 3 Mc. or less it is evident that, if the received signal is to be held well centered in the i.f. band, automatic frequency control of the beating oscillator is essential. The reflex oscillator is particularly well suited to automatic frequency control, and has been widely adopted.

Simplicity in operation of the reflex oscillator arises from a number of properties. A single resonant element, and the fact that a vernier frequency setting and feedback control are obtained by adjustment of the repeller or reflector voltage, are major features in this simplification. Frequency adjustment by the repeller voltage requires a negligible amount of power and, for practical purposes, is free of inertia effects, permitting extremely rapid frequency correction.

Operation of the Reflex Oscillator

External-Cavity Type

Fig. 1 shows an X-ray view of the Western Electric 707A, which was the first reflex tube used extensively as a beating oscillator in 3000-Mc. radar systems. This tube, designed for the application of an external resonator, has two copper disks extending through the glass envelope, to which the external resonator, sketched in Fig. 1, is attached. Two grids, mounted over holes in the disks, permit electrons to pass through the resonator on their way into the retarding electric field between the resonator and the repeller, the latter electrode being maintained at a negative potential with respect to the cathode.

In operation, electrons leaving the cathode are formed into a cylindrical beam by the beam-forming electrode, held at cathode potential, and by the positive grid G_1 maintained at the positive potential of the resonator. As the electrons pass through the resonator from

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the cathode, the high-frequency field between the grids G_2 and G_3 superimposes a sinusoidally varying component of velocity on the stream velocity, determined by the d.c. potential of the resonator. Since as many electrons are speeded up by the same amount as an equal number are slowed down, the work done on the outgoing stream

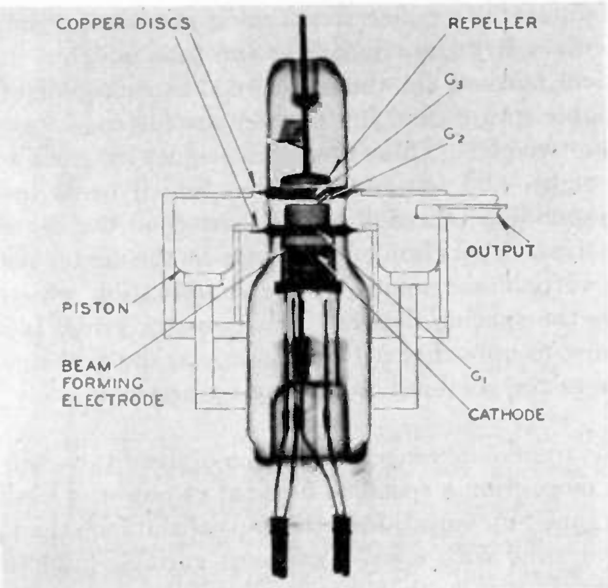


Fig. 1—X-ray view of a W.E. 707A tube. A method of mounting an external resonator on this tube is indicated.

by the field is zero to a first order. During the excursion of the electrons in the retarding field of the repeller space, the slower electrons tend to catch up with the faster ones to produce groups or bunches of charge in the returning stream. The phase at which the bunches return is controlled by an adjustment of the repeller voltage. In general, the returning electron stream produces an admittance across the gap which, depending on the phase of arrival, may be real or complex and with the real component positive or negative. It can be shown[1] that a pure negative conductance will appear when the transit angle θ, which is the phase difference in the r.f. gap voltage between the leaving and returning times of an electron whose velocity is unmodulated by the gap, is given by

$$\theta = (n + 3/4)2\pi \tag{1}$$

where n is 0, 1, 2, 3, etc. It is evident that there can be a number of repeller voltages or modes for which oscillation at the frequency of the resonator will exist, corresponding to the different values of n.

If the repeller voltage is varied from the values which give pure negative conductance across the gap, it is possible to introduce positive or negative susceptance components which decrease or increase the frequency from that of the resonant frequency of the cavity alone. By suitable design the frequency may be varied over a sufficient range to follow the frequency deviations in the radar transmitter, as well as to compensate the changes in frequency in the beating oscillator, caused, for example, by frequency drifts resulting from thermal changes. This frequency control of the reflex oscillator is called electronic tuning. It is frequently specified quantitatively as the frequency change between the two repeller voltages of a given mode which reduce the power output to one-half of the maximum value.

The electronic tuning for the 707A tube working at 3300 Mc. into a typical resonator is 20 Mc. Other typical performance data are as follows:

Resonator voltage	=	300 volts
Repeller voltage	=	-145 to -230 volts
Cathode current	=	35 milliamperes
Drift angle	=	5.5π radians
Power output at 3300 Mc.	=	125 milliwatts.

By the use of an external resonator of the type which has been sketched in Fig. 1, it has been possible to provide a frequency range with the 707A tube extending from 1150 to 3750 Mc. A typical cavity used in a radar application is shown in Fig. 2. Here the resonator frequency is varied by means of plugs screwed into the cavity to change the effective inductance.

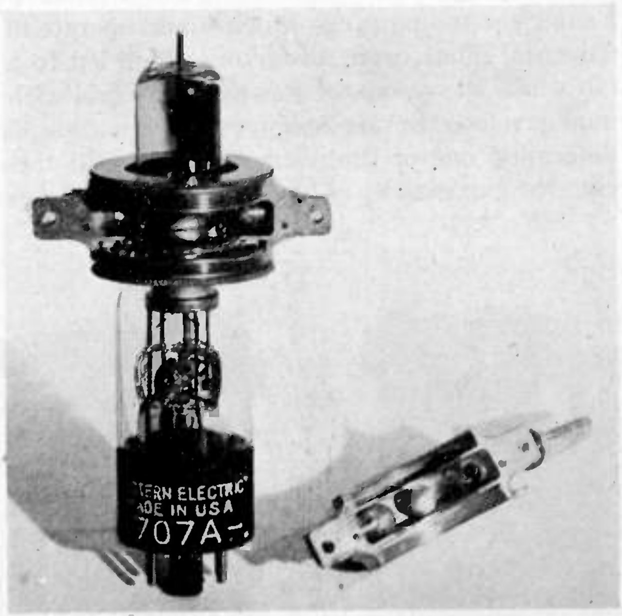


Fig. 2.—W.E. 707A tube with typical resonator. A part of the resonator is shown attached. The other portion, containing the output coupling, is also shown. Frequency adjustment, exclusive of that produced by variations of the repeller voltage, is made by plugs screwed into the cavity.

The power output is taken from the resonator by means of a loop mounted in the part of the resonator external to the tube. This loop can be adjusted within the cavity to afford the optimum coupling at the desired frequency. With proper precautions taken to prevent leakage of radiation, the loop may be insulated from the cavity.

The 707A tube had a very undesirable frequency shift with temperature, caused by relative motion between

[1] J. R. Pierce, "Reflex oscillators," Proc. I.R.E., vol. 33, pp. 112–118; February, 1945.

the two resonator-grid frames. A reshaping of one of the frames minimized this shift, and the modified tube has been given the code number 707B.

Integral-Cavity Reflex Oscillator with External Tuning Control

Early military experience with the 707-type tube disclosed the fact that the externally applied cavity resulted in difficulties with oscillator installation under field conditions. Corrosion of cavities and copper flanges occurred which resulted in poor electrical contact between cavity and flange. The adjustable coupling obtainable with an external cavity, although desirable in laboratory oscillators operating over a wide frequency range, sometimes proved a handicap in the field where personnel was not thoroughly experienced in the techniques of manipulation. There was small need for adjustable coupling over the relatively limited frequency ranges used in any single radar system.

The externally applied cavity-type of structure was not well suited for operation at frequencies of the order of 9000 Mc., where emphasis was next placed. Difficulty in producing the narrow grid-gap spacing with a sufficiently high degree of accuracy, glass losses in the resonator, and the problem of providing an external cavity with sufficient tuning range which would operate in the fundamental mode, were all factors which led to a design in which the resonator was made integral with the vacuum envelope. In this design, tuning is accomplished by deforming one of the resonator walls and thereby varying the capacitance of the resonator gap. In external form this design became the prototype for a series of oscillators which, in combination, cover a large part of the frequency spectrum from 2500 to 10,000 Mc.

Typical of these tubes is the Western Electric 2K29 tube shown in Fig. 3. This tube is designed to cover a frequency range from 3400 to 3960 Mc. A flexible diaphragm supports one of the cavity grids and the housing containing the repeller. The tuning mechanism consists of a tuner strut on one side of the tube which restricts vertical motion. On the opposite side are a pair of deformable spring-steel strips which are fastened together at the two ends. These are spread apart by a combination right- and left-handed screw which turns in two corresponding threaded nuts mounted on the center of the strips. Separation of the strips at the center results in a vertical shortening of the combination which reduces the spacing between the resonator grids. It is of interest to note that a 1-mil change in the grid spacing changes the resonant frequency by approximately 100 Mc.

The frequency range over which a given tube will deliver more than a specified amount of power is less in a tube tuned by variation of the gap capacitance than in a similar tube with a fixed gap and variable-inductance tuning. It can be shown that the power output into a useful load is

$$P = \frac{1}{2}\,\frac{I_0\beta^2\theta}{2V_0}\left[\frac{2J_1\left(\dfrac{\beta V\theta}{2V_0}\right)}{\dfrac{\beta V\theta}{2V_0}}\right]V_2 - \frac{1}{2}G_R V^2 \qquad (2)$$

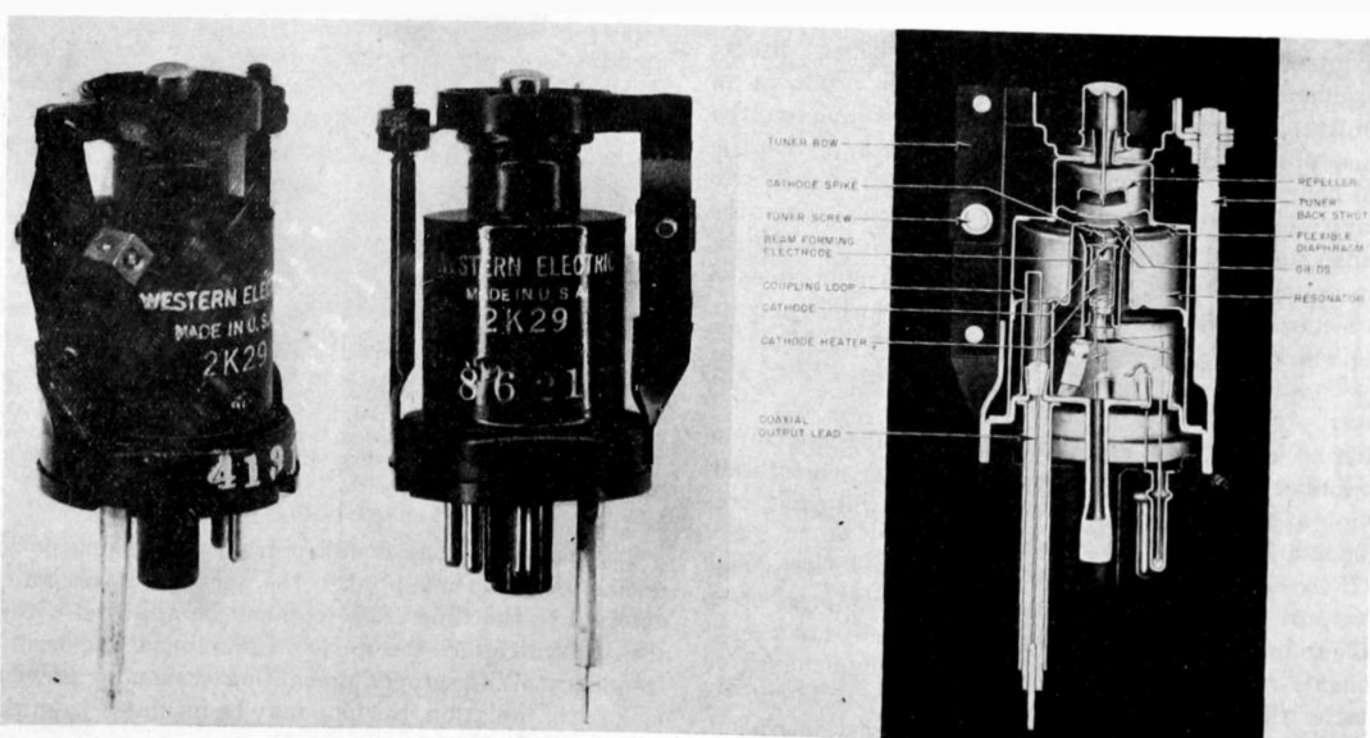


Fig. 3—External and cut-away views of all-metal mechanically tuned beating oscillators. The left-hand views are for a 2K29 showing oscillator, showing the bow construction for mechanical tuning. The frequency range of this oscillator is 3400 to 3960 Mc. The right-hand view shows a cut-away section of the 726B beating oscillator. The frequency range of this oscillator is from 2880 to 3170 Mc.

where

J_1 = the Bessel function of the first order in the argument indicated

I_0 = the direct stream current

θ = the repeller drift angle

V_0 = the d.c. resonator potential in volts

V = the r.f. peak voltage across the gap

β = the beam-coupling coefficient, i.e., the ratio of the peak amplitude of the velocity modulation produced at the gap, expressed in volts, to the r.f. gap voltage

G_R = the conductance representing the resonator losses.

The beam coupling coefficient for a plane-parallel gap is given by

$$\beta = \frac{\sin \theta_g/2}{\theta_g/2} \tag{3}$$

where θ_g is the transit angle across the resonator gap. β has a maximum value of unity for $\theta_g = 0$, and declines monotonically to zero for $\theta_g = 2\pi$. The variation of the bracketed quantity with β is small, so that the principal variation arises from the factor β^2 multiplying this bracket. β, therefore, should be as large as possible, which requires a close gap spacing. However, the resonator losses vary inversely with some power of the gap spacing, so that too close a spacing will increase G_R and thereby reduce the power output. An optimum gap spacing therefore exists at which maximum power output is obtained. The exact value of this gap spacing or transit angle depends upon whether the losses are principally in the inductive or capacitive portion of the resonator. The gap transit angle may be expressed as

$$\theta_g = \frac{1.06 \times 10^{-7}\, d}{\sqrt{V_0}} f \tag{4}$$

where d is the distance in centimeters between the grids, and f is the frequency of oscillation. In order to maintain a fixed beam-coupling coefficient, the gap spacing should be made smaller as the frequency is increased. It is therefore apparent that, whether tuning is accomplished by capacitive or inductive variation, the optimum modulation coefficient cannot be maintained throughout the frequency range. With capacitance tuning the gap spacing increases with frequency, so that the performance drops off rapidly at the high-frequency end because of the declining modulation coefficient. In an oscillator tuned by inductance variation the gap remains fixed, so that this effect is less severe.

The need for maintaining sufficient electronic tuning throughout the useful frequency range imposes special design requirements on the reflex oscillator. A shift of the oscillator frequency requires the introduction by the action of the electron stream of a susceptance component in the electronic admittance which is equal and opposite to the off-resonance susceptance of the resonator. The susceptance near resonance of a simple tuned circuit is given approximately by

$$b = 2C_{eff}\Delta\omega \tag{5}$$

where $\Delta\omega$ is the shift in radian frequency from the value at resonance, and C_{eff} is the effective capacitance. Both the tuned-circuit susceptance and the electronic susceptance decrease with increase in gap spacing. However, the rates of change of the two susceptances differ. Maximum electronic tuning occurs at the gap spacing where the ratio of electronic susceptance to resonator capacitance is a maximum. In general, maximum electronic tuning and maximum power do not occur at the same frequency. The cavity design to produce the best compromise between the variations of electronic tuning and power output is one of the essential features in the development of any reflex beating oscillator. Uniformity of electronic tuning in beating-oscillator applications is more important than the constancy of power output. The compromise is, therefore, weighted in this direction. This will be observed in the characteristics for the 2K29 tube in Fig. 4.

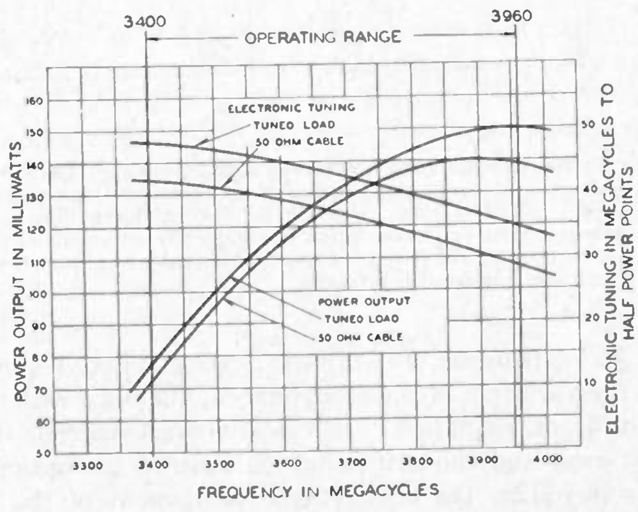


Fig. 4—Power-output and electronic-tuning characteristics for the 2K29 tube. Curves for the tuned-load conditions have been obtained by adjusting a variable load for maximum power output at each measurement. The curves for a 50-ohm-load condition have been obtained by coupling a 50-ohm cable to the tube through a suitable adapter.

The coupling loop and coaxial output line are integral parts of the 2K29-type tube. It will be noted that the coaxial output line is perpendicular to the base, and that installation of a new tube is practically as simple as that of any receiving-type tube. The coupling system has been designed so that the output line may be coupled directly to a 50-ohm line by a suitable adapting fitting. The performance of the tube into a 50-ohm line is compared in Fig. 4 to that into a load optimized at each frequency setting to obtain maximum power output from the tube.

Hysteresis

The term "electronic hysteresis" is used to label a phenomenon which, when present, is evident in the behavior of the power output as a function of the repeller voltage. The simple theory of the reflex oscillator predicts that a curve similar to that of Fig. 5(a) will be obtained. A curve similar to Fig. 5(b) is found when electronic hysteresis occurs. That is, the tracing of the curve in one direction of sweep does not superimpose on that of the other direction over the complete sweep; instead, the power output and frequency are discontinuous functions of the repeller voltage. This effect can be sufficiently serious to cause trouble in utilizing electronic tuning. The discontinuous behavior was at one time thought to result from improper loading of the tube, but it was later established that the effect was electronic in origin.

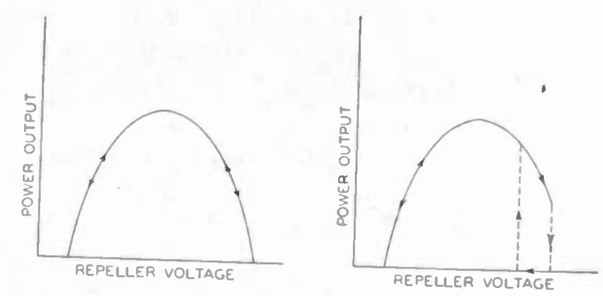


Fig. 5—Power output as a function of repeller voltage. The repeller-voltage range covered is sufficient to show only one operating mode. (a) Ideal curve, showing no discontinuities. The return sweep presents a trace which is completely identical with the forward trace. (b) A type of trace which may be obtained when electronic hysteresis is present.

An explanation of electronic hysteresis may be found in the variation of the electronic conductance with amplitude of oscillation. For cases where hysteresis does not exist and the drift angle is close to the optimum $\theta_0=(n+\frac{3}{4})2\pi$, the conductance component of the admittance may be written with reasonable accuracy as

$$G_e = \left[\frac{I_0\beta^2\theta}{2V_0}\right]\frac{2J_1\left(\dfrac{\beta V\theta}{2V_0}\right)}{\dfrac{\beta V\theta}{2V_0}}\cos\Delta\theta = g_eF(V)\cos\Delta\theta \qquad (6)$$

where $\Delta\theta$, a function of the repeller voltage, is the departure of θ from θ_0, g_e is written for the bracketed factor, and $F(V)$ for

$$\frac{2J_1\left(\dfrac{\beta V\theta}{2V_0}\right)}{\dfrac{\beta V\theta}{2V_0}}.$$

θ can and will be considered constant in these two factors, since its variation does not effect their values to a serious extent over the electronic tuning range. It will be noted that, as V approaches 0, $F(V)$ approaches 1.0. Thus g_e has the physical significance of being the small-signal conductance for $\Delta\theta=0$.

A family of curves of electronic conductance, defined by (6), is shown in Fig. 6(a). For the uppermost curve $\Delta\theta$ is 0, corresponding to a repeller voltage E_0. The lower curves are plotted for various values of $\Delta\theta$, each value of which is assumed to correspond to two values of repeller voltage $E_0\pm\Delta E$. A line is shown representing the negative of the conductance G_L, which is the sum of the resonator conductance and the load conductance. Since stable oscillation requires that $G_e+G_L=0$, the steady-state amplitude of oscillation for each value of $\Delta\theta$ in Fig. 6(a) occurs where the electronic-conductance curve intersects the load conductance. A plot in Fig. 6(b) of

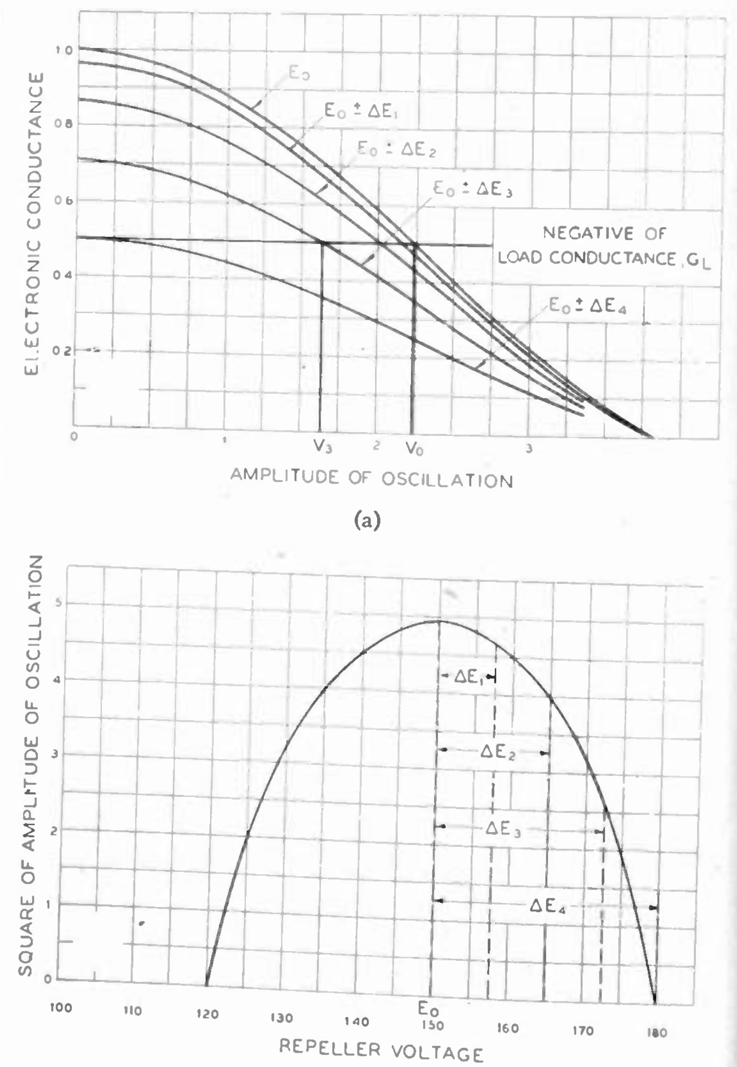


Fig. 6—Electronic-conductance and power-output characteristic for the case where hysteresis is absent. (a) Electronic conductance as a function of amplitude of oscillation for various values of repeller voltage. Also shown is a line representing the negative of the load conductance. (b) Resulting power output versus repeller voltage curve.

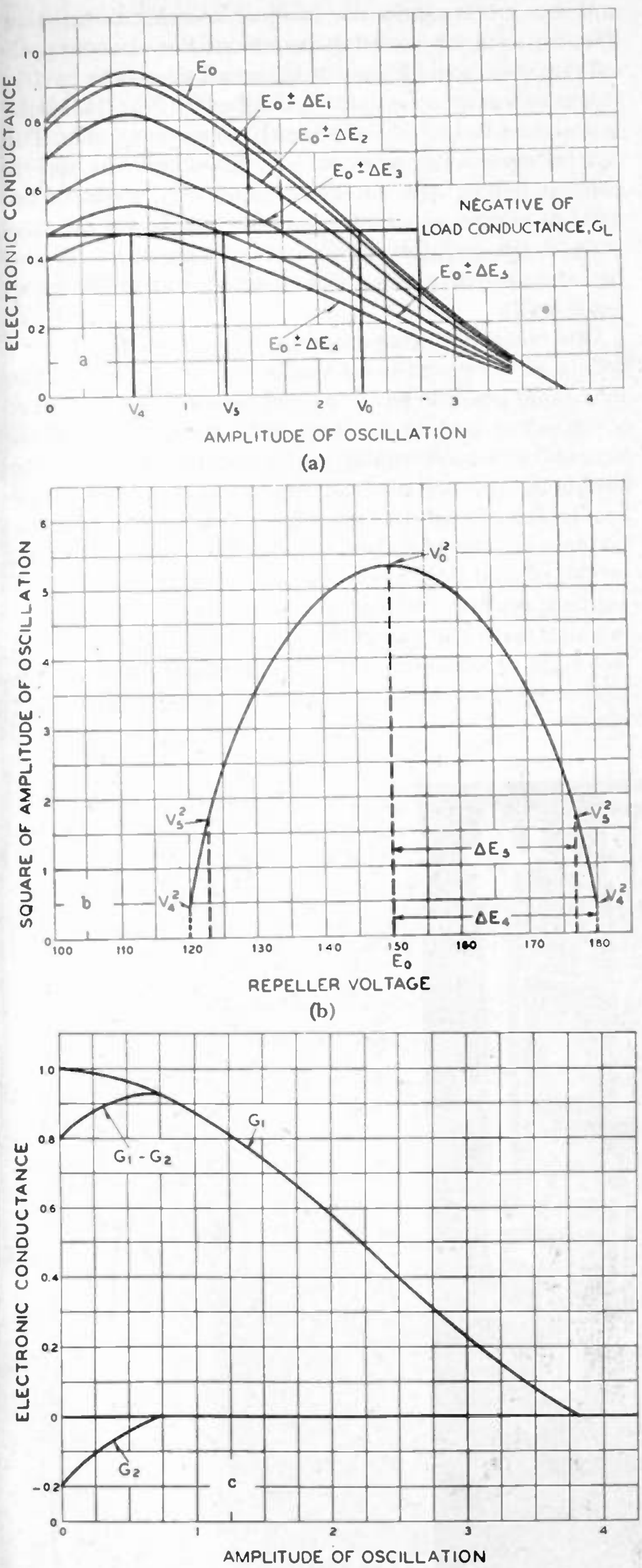


Fig. 7—Type of electronic-conductance and power-output characteristics which may occur where hysteresis is present. (a) Electronic conductance as a function of amplitude of oscillation for various values of repeller voltage. (b) Resulting power output versus repeller voltage curve. (c) Effect of an out-of-phase source of conductance on the total electronic conductance.

the power output derived from these characteristics is similar to the ideal curve of Fig. 5(a).

Electronic hysteresis may be explained by assuming a family of electronic-conductance curves as shown in Fig. 7(a). For certain values of the electronic conductance, the amplitude is double-valued. Starting from the condition of optimum repeller voltage, E_0, and varying the magnitude of the repeller voltage, the oscillation amplitude will decrease smoothly until an amplitude V_4, at repeller voltages $E_0 \pm \Delta E_4$, is reached. A further change in repeller voltage will cause the amplitude to fall to zero. If the repeller voltage is then changed toward E_0, the power output will remain zero until the repeller voltage reaches $E_0 \pm \Delta E_5$, at which point the small-signal electronic conductance becomes equal to the load conductance. At this point any slight disturbance will cause the electronic conductance to exceed the load conductance, and oscillation will start and rise immediately to the amplitude V_5. Further change in repeller voltage toward E_0 will produce a smooth variation of amplitude. For the simplified case illustrated, the hysteresis effect will appear on both ends of the repeller-voltage range (Fig. 7(b)) as a result of the assumed symmetry of variation about the E_0 value. Since the variation in general is not symmetrical, hysteresis may appear only on one end of the characteristics, as shown in Fig. 5(b).

The type of conductance curves shown in Fig. 7(a) can occur if a second source of conductance exists opposing the primary source and varying with amplitude, as illustrated in Fig. 7(c). A second source may be found by considering the electrons, which return from the repeller space, through the gap, and enter the cathode-

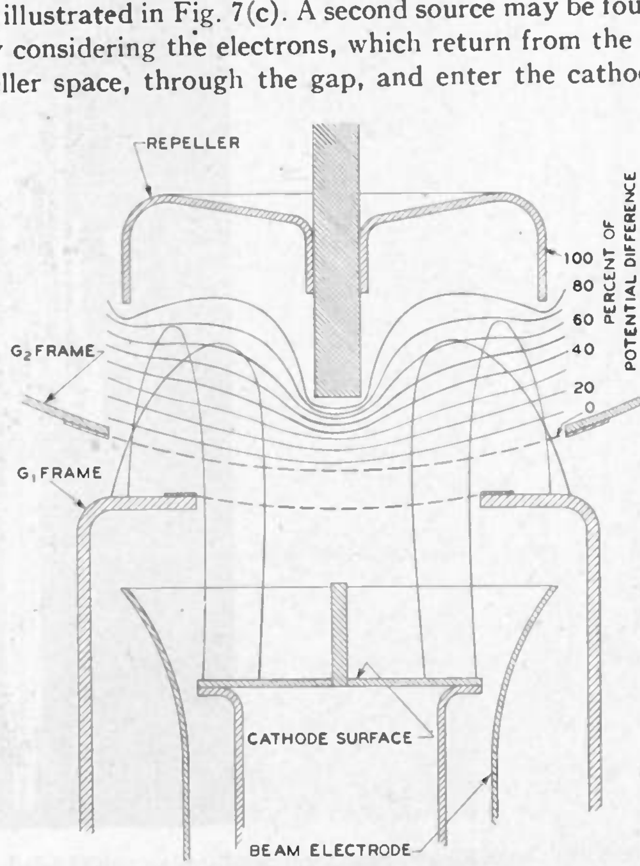


Fig. 8—Typical electrostatic field plot for the repeller space, showing calculated electron paths.

resonator region. These electrons continue to bunch, and some of them will be returned through the gap where they will give rise to electronic conductance.

This source of hysteresis may be eliminated by insuring that the electrons make only one round trip through the gap. The geometrical arrangement adopted in the 2K29 tube to accomplish this result is shown in Fig. 8. The electron stream leaving the cathode is formed into a hollow cylindrical beam by the beam electrode and the central spike projecting from the cathode. The cylindrical beam, following the paths calculated from the equipotentials, returns through the larger secondary-grid opening and is captured on the frame surrounding the first grid.

Other secondary sources of electron admittance may result in hysteresis. The most common in reflex oscillators appears to be that discussed above.

Broad-Band Operation of Reflex Oscillators

The 2K25 tube, a cross-section view of which is shown in Fig. 9, was developed for use in the frequency range 8500 to 9660 Mc. It will be noted that the resonant cavity in this tube is much smaller than in the 2K29 type, and has more nearly the form of a disk transmission line, capacitance-loaded at its center. For electron-optical purposes, and because of the small size of the cavity, it was necessary to employ a third grid. The 2K25 tube is designed to be coupled directly to a wave guide. The central conductor, protected by a polystyrene sleeve, extends beyond the outer conductor of the output coaxial line to form a probe, slightly less than a quarter-wavelength long. Coupling to the wave guide is achieved by extending the probe through an opening in the wave-guide wall.

One of the problems encountered in designing a reflex oscillator for broad-band use is that of delivering the maximum possible power output into a given load over a frequency range consistent with stability of oscillation and electronic-tuning requirements. Generally, the load is the characteristic admittance of a wave guide or a coaxial line. It will be recognized that the various components of the coupling system, such as the output probe, coaxial line, loop, and cavity configuration, will act to transform the load admittance to an admittance in shunt with the gap of the resonator. If the problem were one of obtaining an approximately constant admittance across the gap, it would be a relatively simple one.

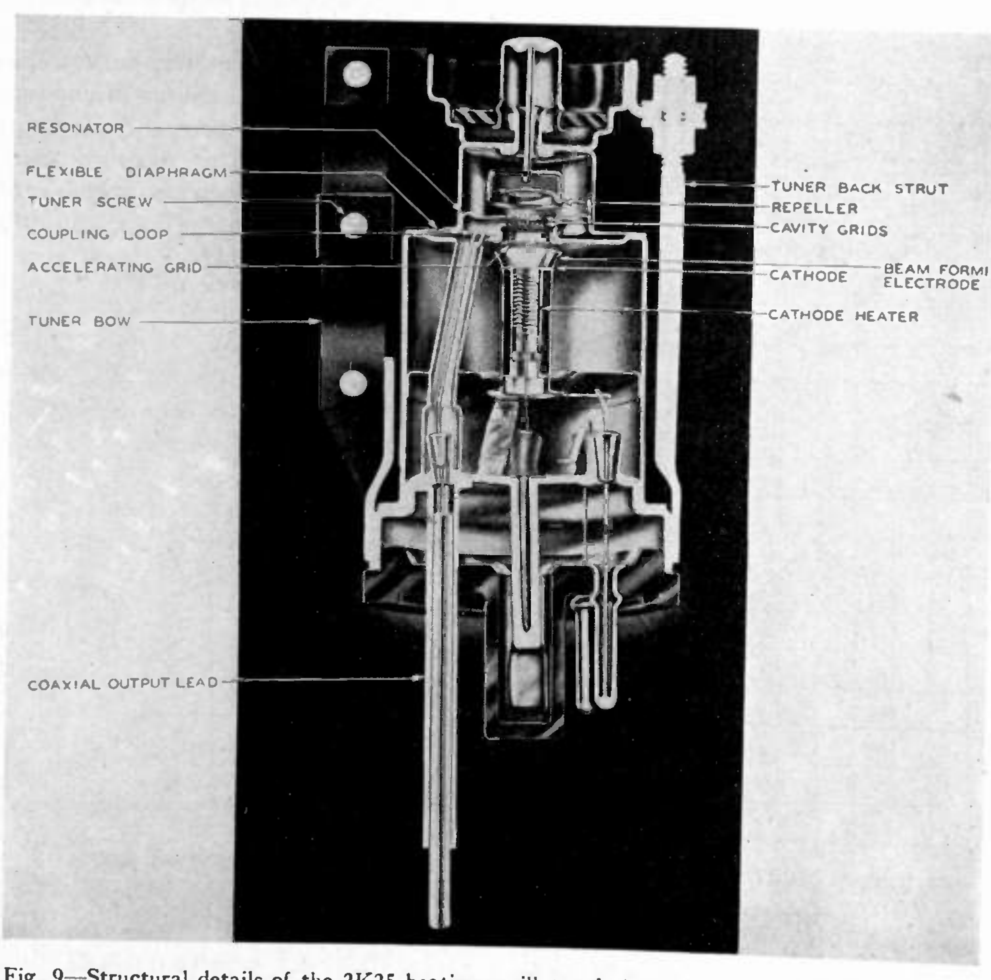


Fig. 9—Structural details of the 2K25 beating oscillator designed to cover a frequency range from 8500 to 9660 Mc. The output coaxial line of this tube is designed to be coupled directly into a wave guide.

In a reflex oscillator the electronic admittance and the cavity losses are not constant over the frequency band. It is necessary to present to the gap a load admittance which varies with frequency in a manner defined by the particular variation in electronic admittance and cavity losses. Sufficiently detailed information concerning the electronic admittance and the resonator conductance would make it possible to determine theoretically the necessary configuration of the coupling system for best performance. Unfortunately, some of the parameters required for calculation of the admittance have not been established adequately, and it is necessary to resort to experimental methods.

The first step in such an experimental method is to obtain a suitable smooth broad-band transducer between the characteristic admittance of the load and the tube output coaxial line. An estimate is then made of a loop size to give a correct mutual inductance with the resonator. The admittance which must be presented to the transducer to obtain optimum power output is then

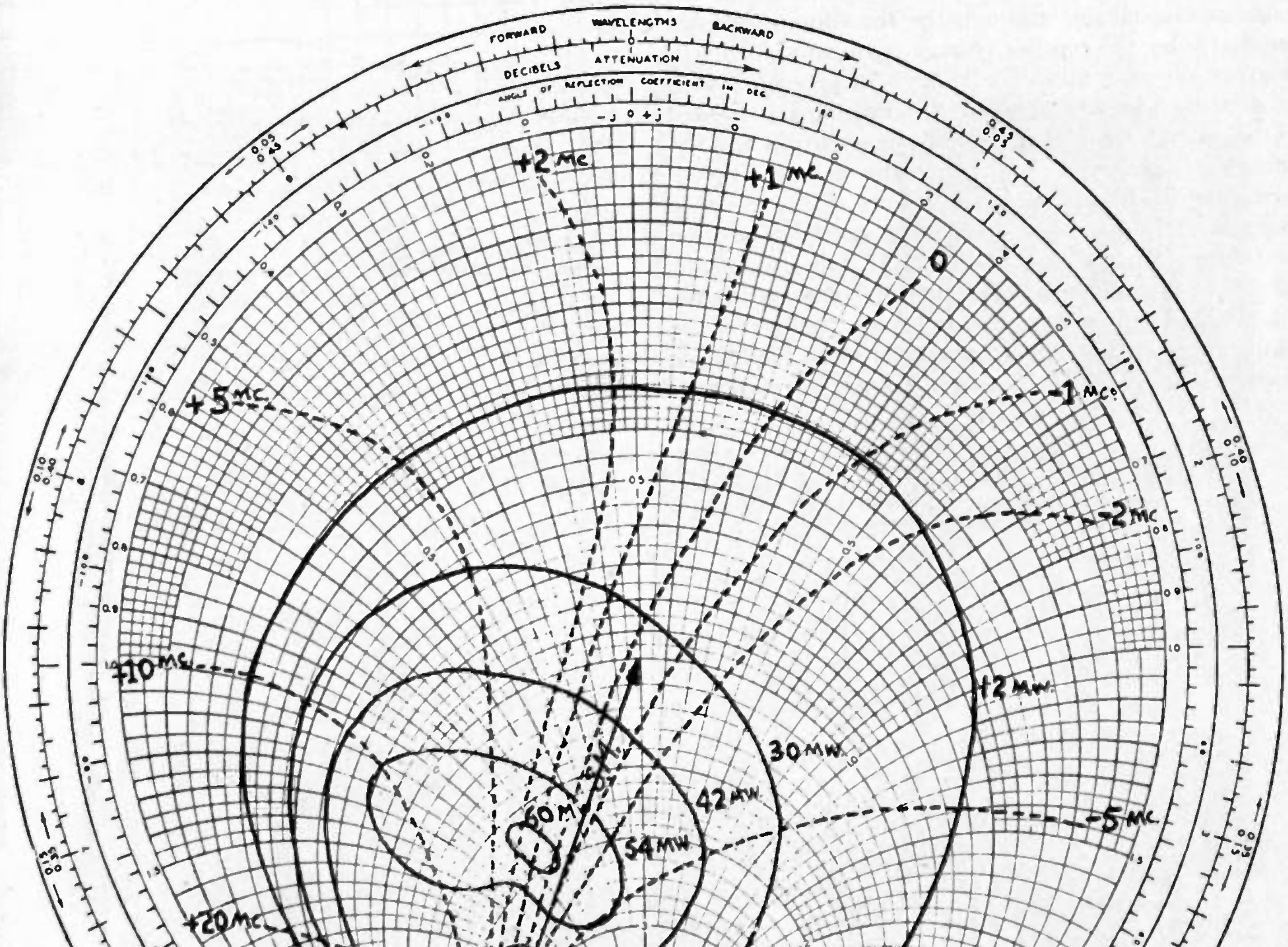


Fig. 10—Impedance performance diagram of a 2K25 tube. Loci of constant power are shown by the solid closed curves. The dashed lines show the loci of constant frequency deviation from the nominal frequency. The tube is operating in mode A, which mode is shown in Fig. 11. Sink margin is indicated by the line SM. This diagram is for mode A, $F=9360$ Mc., and $E_{RBS}=300$ volts.

measured over the band. From these measurements it is possible to compute with reasonable accuracy the correction in loop size and in the transducer characteristics so that, when the characteristic impedance of the line or guide is presented to the transducer, the optimum performance is obtained over the band. A transducer was designed for the 2K25 tube in this manner.

For reasons of frequency stability, it is not always desirable to deliver the maximum available power to the load. The reasons for this are best illustrated by means of an impedance performance plot.[2] With the tube coupled to the characteristic impedance of the guide or coaxial line through the transducer designed for that tube, the repeller voltage is adjusted for maximum power at a given frequency. All operating voltages are then held fixed while the admittance presented to the oscillator is varied and the power output and frequency are observed. The admittances normalized in terms of the characteristic admittance of the guide or line are referred to some convenient reference position on the line. In the case of the 2K25 tube, the plane normal to the guide through the projecting coaxial line was selected and the performance characteristics for the admittances in that plane are shown on an impedance chart in Fig. 10. Loci of constant power are indicated by the solid closed curves, while the dashed lines are the loci of constant frequency deviations from the nominal frequency. It will be noted that for a certain admittance

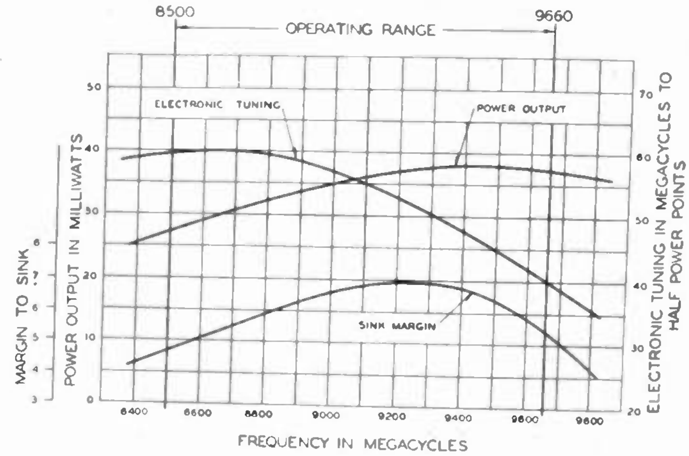


Fig. 11—Power output, electronic tuning, and sink margin for a typical 2K25 tube when coupled to the characteristic impedance of the wave guide through a suitable coupling.

range, oscillation does not exist. This corresponds to a region in which the sum of the load and resonator conductance is in excess of the electronic conductance. The voltage-standing-wave ratio to the nearest point of this

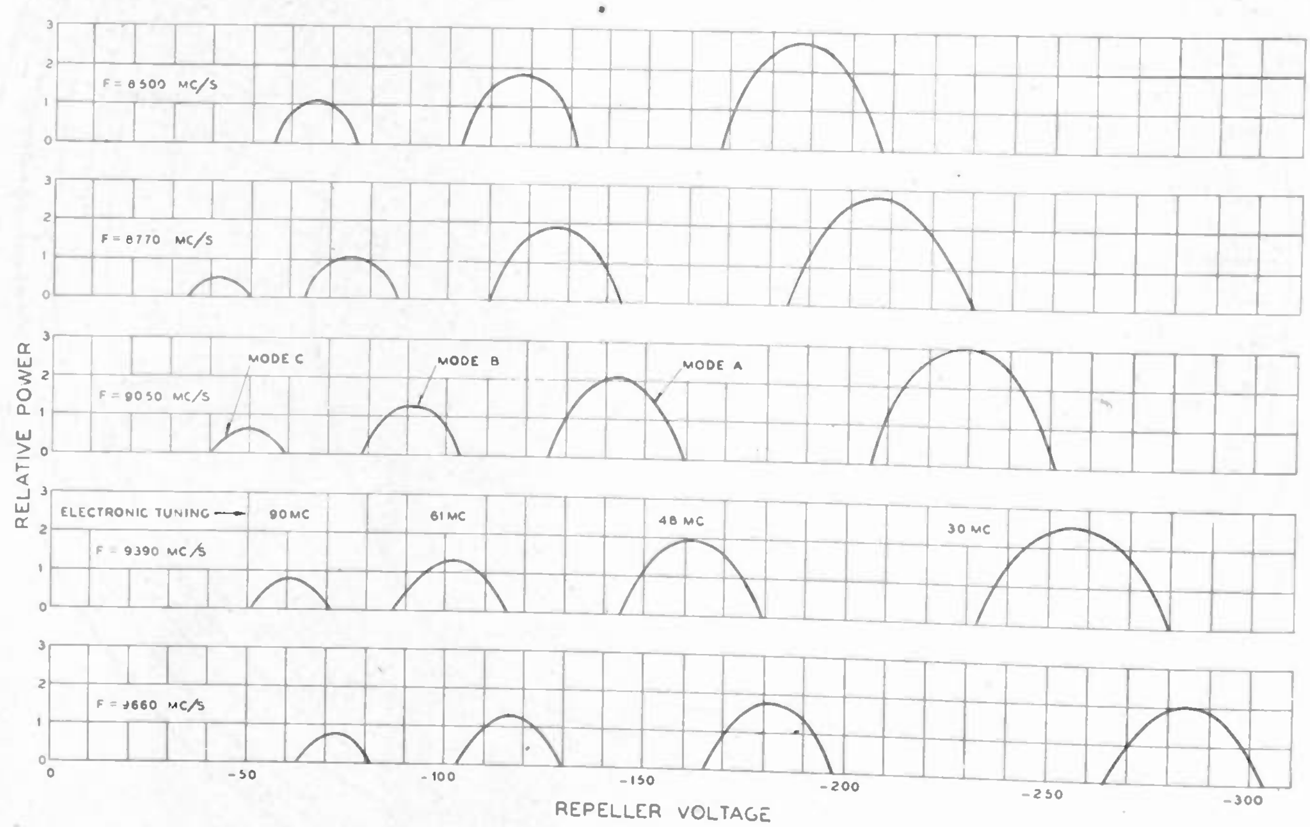
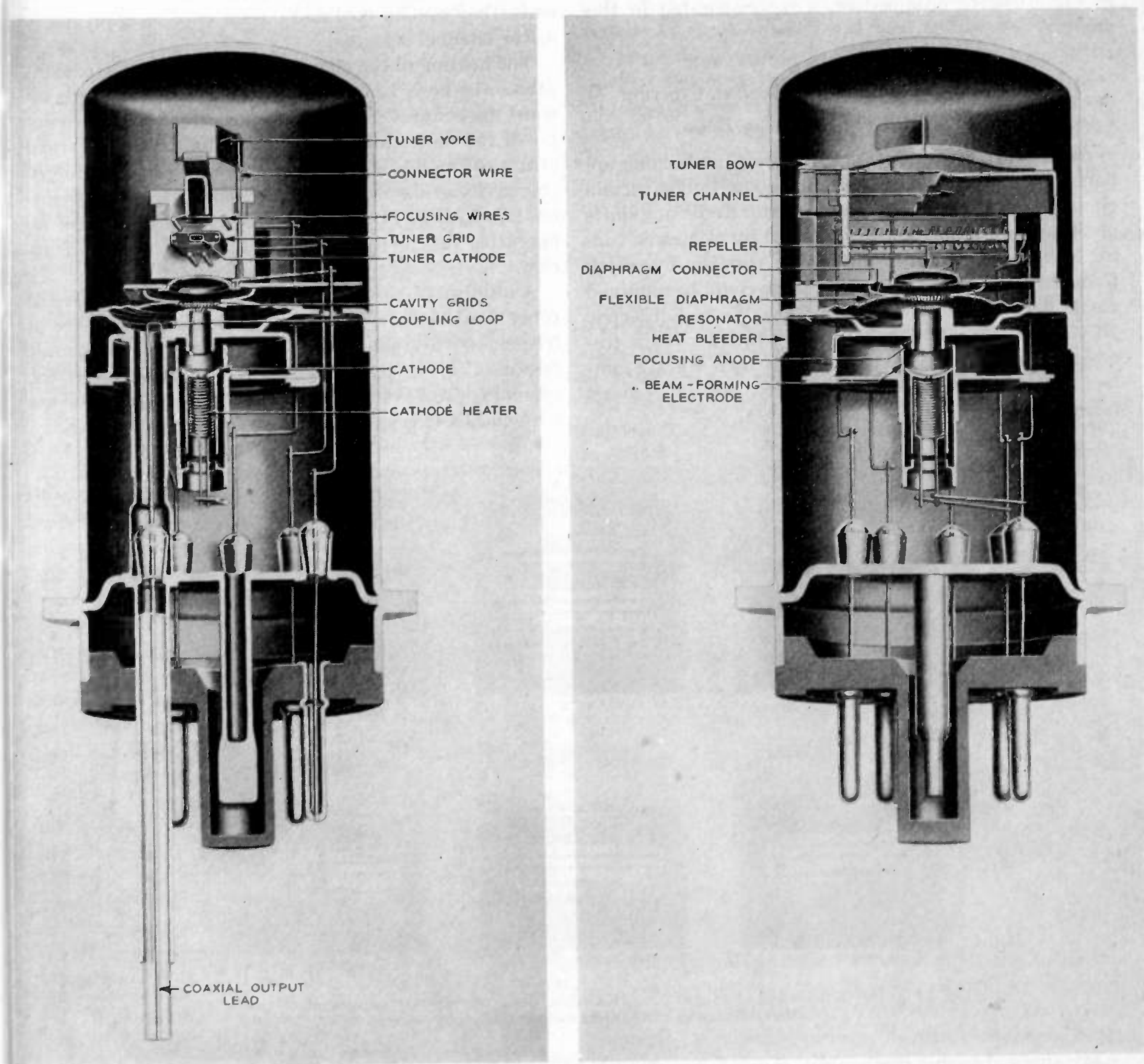


Fig. 12—Power-output performance for a typical 2K25 tube for five different nominal operating frequencies. The performance in the various modes is shown, and the electronic tuning, between half-power points, is indicated for the 9390-Mc. condition. The resonator voltage is 300 volts.

[2] P. H. Smith, "Transmission line calculator," *Electronics*, vol. 12, pp. 29–31; January, 1939.

region from the unity standing-wave point we have chosen to designate as the sink margin, since this is a measure of the factor by which the load admittance may be increased above that of the characteristic admittance of the line without stoppage of oscillation. One may show theoretically and verify experimentally that, if the load admittance for which maximum power is delivered is presented to the oscillator, then approximately doubling the magnitude of this admittance will result in a stoppage of oscillation. Hence, if a sink margin in excess of approximately 2 is desired, it will be necessary to decouple the load by a suitable amount. This entails a sacrifice in power. The coupling system of the 2K25 is such that there is a minimum sink margin of 2.5 at a frequency of 9400 Mc. In order to insure this minimum, a 2K25 tube having average characteristics has a sink margin of approximately 6.7 times. Thus, one compromise in the coupling design is between maximum possible power output and required sink margin.

The broad-band characteristics for a typical 2K25 tube are indicated in Fig. 11. Here the power output, half-power electronic tuning, and sink margin are given as functions of frequency when the tube is operated through the transducer designed for the tube into the



(a) (b)

Fig. 13—Sectional views of the 2K45 thermal-tuned reflex oscillator.

characteristic admittance of a 1- $\times\frac{1}{2}$-inch wave guide. In the design of a reflex oscillator there is some choice in the number of cycles of drift in the repeller space. Neglecting resonator losses, it can be shown that as the number of cycles of drift increases the power output decreases, but the electron tuning increases. The choice of the number is, therefore, determined by a compromise. Fig. 12 gives the relative power output for different numbers of cycles of drift, and for a number of different frequencies, for a typical 2K25 tube. Half-power electronic tuning is indicated for each mode, and the repeller mode indicated as A is the one which has been recommended for general use. The data presented in Fig. 10 were obtained on a tube operated in this mode.

Internal Cavity—Internal Tuning Control

Thermal Tuning

As the tactical use of radar advanced it became apparent that it would be advantageous, under certain circumstances, to produce much greater frequency shifts of the system than could be followed by electronic tuning. An investigation of the various means by which frequency variation over a wide range could be obtained resulted in the development of the Western Electric 2K45 oscillator. This tube can be tuned over a frequency range from 8500 to 9660 Mc. by a voltage control. Fig. 13 illustrates this tube by two sectional views taken at right angles.

The channel of a material having a large coefficient of expansion is heated by electrons from a thermionic cathode. The electron flow is directed into the channel by the focusing wires, and is controlled by a negative grid which draws no appreciable power from the control system. Longitudinal expansion of the channel is permitted by flexible tabs bent down at right angles to the channel axis and fastened to the resonator. These tabs serve as heat bleeders for the channel ends, as well as providing vertical support. The multileaved bow is rigidly fastened to the channel ends. The leaves are made of a material having a low coefficient of expansion and, since they are fastened to the channel only at the ends, they remain cool and do not expand appreciably as the channel is heated.

The heating of the channel by electron bombardment causes the bows to flatten and their center to move toward the channel. A cross member attached to the center of the bows transmits this motion through vertical struts to the flexible diaphragm which supports one of the cavity grids and forms one wall of the cavity resonator. In Fig. 14 a series of X-ray photographs of an operating tube is shown which illustrates the tuner action.

A number of basic design requirements in a thermal tuner must be met, such as speed of tuning, absence of frequency "overshoot," and satisfactory microphonic response. These requirements depend in part on the type of control system used. That contemplated for use with the 2K45 is such that the control adjusts the heat-

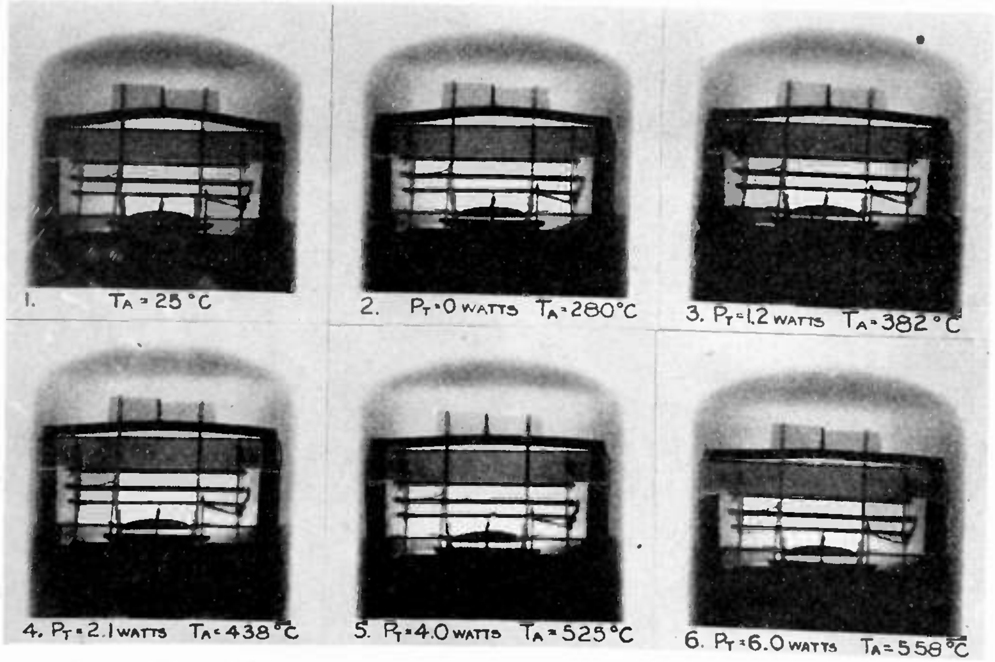


Fig. 14—X-ray views of an operating 2K45 tube, demonstrating the movement of the thermal-tuning mechanism for various input powers.

The thermal tuning mechanism, contained in the upper part of the structure, consists of the bimetallic combination of a U-shaped channel and a multileaved bow. ing of the thermal tuning mechanism on a "full-on" or "full-off" basis.

Speed of tuning at all points in the band requires that two principal conditions should be satisfied. First, the power into the tuner for the full-on condition must be considerably in excess of the power needed to hold the tuner at the limit of the frequency band nearer the full-on condition. This insures that, when operating near this limit, rapid frequency response will be obtained in both directions. The thermionic system must be capable of delivering the excess power, and the tuner must be capable of dissipating it continuously without damage in case the full-on condition accidentally persists. Second, the power to hold the tuner at the limit of the frequency band nearer the full-off condition must be sufficiently in excess of zero so that, near this limit, a rapid response will result when the tuner is switched off.

With the tuner power either full-on or off and the frequency changing toward a stable value, overshoot is said to occur if, on switching the tuner power to the opposite condition, the frequency change continues in the original direction for a period after switching. To prevent overshoot, the heat which activates the tuner must be generated directly in the expanding element, as is done in the 2K45. It is not satisfactory, for example, to activate the tuner by heat radiated to the expanding element by a resistance heater.

Minimization of microphonic response, i.e., frequency shift resulting from vibration and tuning speed, impose conflicting requirements. Stiffness of struts is essential to low microphonic response, whereas tuning speed demands a minimum heat capacity and therefore small mass. A further limitation on the stiffness is a somewhat arbitrary limit on the magnitude of the driving power allowed to produce the necessary motion. Hence, the final design represents a compromise between speed and microphonic performance.

In the 2K45, with the microphonic response reduced to a satisfactory level, an average speed of tuning of approximately 150 Mc. per second is obtained in either direction. This rate, corresponding to a time of 7.7 seconds to tune through a range of 1160 Mc., is based on full-on or off operation. The required frequency band is covered in a power range lying approximately between 2 and 4.3 watts. The tuner can dissipate approximately 7 watts continuously without destruction.

The 2K45 oscillator design is a departure in some respects from those described previously. The concave cathode, cylindrical beam-forming electrode, and focusing anode constitute an electron gun designed according to principles outlined by J. R. Pierce[3] and A. L. Samuel.[4] This gun produces an electron stream converging radially into the focusing anode. The beam has a minimum diameter in the neighborhood of the cavity grids, beyond which it diverges with considerable rapidity. On its return from the repeller region the stream passes through the second grid, which has a larger diameter than the first, and most of the electrons are captured on the surface supporting the first grid, the remainder striking the wall of the focusing anode. This results in virtually complete elimination of electronic hysteresis.

The elimination of the accelerating grid and the improvement in the design of the resonator result in the same power output as for the 2K25 tube, although the cathode current is only about two-thirds that in the latter tube.

The repeller is fixed in the 2K45 tube and the upper grid moves relative to it, while in the structures discussed earlier the repeller was fixed relative to the nearest grid. As a result of the 2K45 arrangement, the repeller space is shortened as the frequency of the cavity is increased. This reduces the repeller-voltage variation necessary for oscillation over the frequency band, since a reduction in the repeller-to-grid spacing tends to provide the required decrease in drift time as the frequency increases.

Scope of Oscillator Development

This paper has discussed the design problems of a few particular reflex-oscillator tubes. In addition to those described, a number of others have been developed, or are currently undergoing development, at the Bell Telephone Laboratories. A chart showing the frequency ranges of these tubes and their place in the frequency spectrum is presented in Fig. 15. All tubes in the lowest two rows are oscillators having the general construction of the 2K29 and 2K25. Development work on the 1413 and 1449 tubes has not been completed, and for that reason they carry laboratory development code numbers.

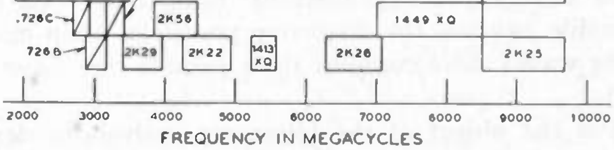


Fig. 15—Frequency coverage of a number of beating oscillators developed in the frequency spectrum of 2000 to 10,000 Mc.

Acknowledgment

So many people have contributed to the work described in this paper that it is impractical to make individual acknowledgment. To these people, physicists, chemists, electrical and mechanical engineers, and laboratory assistants, the authors are greatly indebted.

[3] J. R. Pierce, "Rectilinear electron flow in beams," *Jour. Appl. Phys.*, vol. 2, pp. 548–554; August, 1940.
[4] A. L. Samuel, "Some notes on the design of electron guns," Proc. I.R.E., vol. 33, pp. 233–240; April, 1945.

The Distortion of Frequency-Modulated Waves by Transmission Networks*

A. S. GLADWIN†

Summary—A general solution to the problem of calculating the distortion imposed on the instantaneous frequency of a frequency-modulated wave in passing through a transmission network is obtained by a direct operational method. Approximate formulas for the cases of large and small deviation ratios are derived, and it is shown that a range of overlap exists in practical cases. For very large deviation ratios the distortion is entirely nonlinear in character and depends on the maximum frequency deviation, while for very small deviation ratios the distortion is entirely linear and is independent of the frequency deviation. The nature of the distortion is examined with particular reference to intermodulation distortion. When the modulating wave consists of two sine waves of different amplitudes and frequencies, intermodulation distortion takes the form of a frequency modulation of the small high-frequency component by the large low-frequency one. The application of negative feedback to a frequency-modulation receiver is considered. Numerical examples are worked out.

I. Introduction

THE DISTORTION suffered by a frequency-modulated wave in passing through a transmission network has been investigated by Carson and Fry,[1] who obtained a theoretical solution, and more recently by Jaffe,[2] who calculated the numerical value of the harmonic distortion for the particular cases of sinusoidal modulation with networks consisting of either a single resonant circuit, or a pair of resonant circuits critically coupled and tuned to the carrier frequency.

These analyses apply to the case of a large deviation ratio, but important practical cases exist in which the deviation ratio is small; for example, a superheterodyne receiver in which negative feedback is used to reduce the frequency deviation of the received wave. Moreover, in practice, transmission networks are more complicated than those examined by Jaffe, and it is also desirable to know the distortion produced when modulating waves more complex than a single sine wave are used.

It is the object of the following analysis to derive formulas, suitable for both large and small deviation ratios, from which numerical values of the distortion products can be calculated for any type of transmission network.

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[1] J. R. Carson, and T. C. Fry, "Variable frequency electric circuit theory with application to the theory of frequency modulation," *Bell Sys. Tech. Jour.*, vol. 16, pp. 513–541; October, 1937.

[2] D. L. Jaffe, "A theoretical and experimental investigation of tuned circuit distortion in frequency-modulation systems," Proc. I.R.E., vol. 33, pp. 318–334; May, 1945.

II. List of Symbols

$A(u)$ = amplitude characteristic of the network (nepers)

D = deviation ratio = maximum frequency deviation/highest modulating frequency

$\phi(u)$ = phase characteristic of the network (radians)

p = differential operator = d/dt

$P(u)$ = in-phase component of the network transfer characteristic

$Q(u)$ = quadrature component of the network transfer characteristic

S = the modulating wave

$T(u) = Y(j\omega)$

$u = (\omega - \omega_c)/\omega_B$

v_i = input to a network

v_0 = output from a network

ω_B = semibandwidth of the network (radians/second)

ω_c = carrier frequency (radians/second)

$\Delta\omega$ = maximum frequency deviation (radians/second)

$Y(j\omega)$ = steady-state complex transfer characteristic of the network = output/input.

III. The General Solution

Let the modulating wave be denoted by S, and let the peak value of S be unity. A sinusoidal carrier wave frequency modulated by S can be written

$$v_i = \cos\left(\omega_c t + \Delta\omega \int S dt\right)$$

$$= R \exp j\left(\omega_c t + \Delta\omega \int S dt\right)$$

where ω_c is the carrier frequency, $\Delta\omega$ is the maximum frequency deviation, and R denotes "the real part of." Conventionally, R is omitted from the analysis, but it should always be understood.

When the modulated carrier is applied at the input terminals of a linear transmission network having a transfer characteristic $Y(j\omega)$, the output from the network is

$$v_0 = Y(p)v_i.$$

$Y(p)$ is the transfer characteristic with the differential operator $p = (d/dt)$ written in place of $j\omega$. $Y(p)$ is an operational function, and the output v_0 is obtained by operating on v_i with $Y(p)$. This can be done conveniently by using Murphy's shifting theorem[3] to give the result

$$v_0 = \exp j\left(\omega_c t + \Delta\omega \int S dt\right) Y(p + j\omega_c + j\Delta\omega S). \quad (1)$$

The subject of the operational function is now unity. It is supposed that the modulated carrier has been applied to the network for a very long time, so that the transient solution $Y(p)\cdot 0$ has vanished at the time under consideration.

Since most of the networks to be analyzed are band-pass filters, it is convenient to express a frequency in terms of its difference $\omega - \omega_c$ from the carrier frequency, and also to express this difference as a fraction u of the filter semibandwith ω_B, i.e., $u = (\omega - \omega_c)/\omega_B$. The transfer characteristic may then be specified in terms of a shape function $T(u)$, a scale factor ω_B, and a position factor ω_c, thus:

$$Y(j\omega) = Y(j\omega_c + j\,\overline{\omega - \omega_c}) = Y(j\omega_c + ju\omega_B) = T(u). \quad (2)$$

Applying this transformation to (1) gives

$$Y(p + j\omega_c + j\Delta\omega S) = T\left(\frac{\Delta\omega S - jp}{\omega_B}\right).$$

If it is assumed that this function can be expanded in a power series by Maclaurin's theorem, then

$$T\left(\frac{\Delta\omega S - jp}{\omega_B}\right) = \sum_0^\infty \frac{1}{n!}\left(\frac{\Delta\omega S - jp}{\omega_B}\right)^n T^n(0) \quad (3)$$

where $T^n(0) = d^n/dt^n T(u)\big|_{u=0}$. The expansion is valid only if the series so obtained converges. In particular, the series does not converge if S is a step function, or if there are any discontinuities in $T(u)$ or its derivatives.

The operator $(\Delta\omega S - jp)^n$ denotes $(\Delta\omega S - jp)(\Delta\omega S - jp) \cdots$ to n terms, and may be expanded as follows:

$$(\Delta\omega S - jp) = \Delta\omega S$$

$$(\Delta\omega S - jp)^2 = (\Delta\omega S - jp)\Delta\omega S = (\Delta\omega S)^2 - j\Delta\omega S'$$

and so on:

$$\left(S' = \frac{d}{dt} S\right).$$

It is found that the terms in the series resulting from the expansion of (3) are themselves the Maclaurin series for functions such as $T(\Delta\omega S/\omega_B)$. When all the terms are collected in this way, (1) can be written

[3] A. G. Warren, "Mathematics Applied to Electrical Engineering," Chapman and Hall, London, 1939; p. 169. The theorem states that $Y(p)\exp f(t) = \exp f(t)\, Y(p + f'(t))$ both sides of the identity being operational functions. The analysis can also be carried out, with a little more trouble, by using the well-known Heaviside shifting theorem, which is a particular case of Murphy's theorem. This was the method used by Jaffe.

$$\begin{aligned} v_0 = \Bigg[&\exp j\left(\omega_c t + \Delta\omega \int S dt\right)\Bigg]\Bigg[T\left(\frac{\Delta\omega S}{\omega_B}\right) \\ &- \frac{\Delta\omega}{\omega_B}\left\{\frac{jS'}{2!\omega_B} T''\left(\frac{\Delta\omega S}{\omega_B}\right) + \frac{S''}{3!\omega_B^2} T'''\left(\frac{\Delta\omega S}{\omega_B}\right)\right. \\ &\left. - \frac{jS'''}{4!\omega_B^3} T^{IV}\left(\frac{\Delta\omega S}{\omega_B}\right) - \frac{S^{IV}}{5!\omega_B^4} T^{V}\left(\frac{\Delta\omega S}{\omega_B}\right) + \cdots\right\} \\ &+ j\frac{\Delta\omega^2}{2\omega_B^3} S'\left\{\frac{jS'}{4\omega_B} T^{IV}\left(\frac{\Delta\omega S}{\omega_B}\right)\right. \\ &+ \frac{S''}{3!\omega_B^2} T^{V}\left(\frac{\Delta\omega S}{\omega_B}\right) - \frac{jS'''}{4!\omega_B^3} T^{VI}\left(\frac{\Delta\omega S}{\omega_B}\right) \\ &\left. - \frac{S^{IV}}{5!\omega_B^4} T^{VII}\left(\frac{\Delta\omega S}{\omega_B}\right) + \cdots\right\} \\ &+ \text{etc.}\Bigg] \end{aligned} \quad (4)$$

This is the general solution, but in the form given above it is of little practical use. The desirable form of solution would express the result as a carrier wave modulated in amplitude and frequency. Approximate solutions of this form can be found when the deviation ratio is large or small.

IV. Solution for Large Deviation Ratios

When the deviation ratio is large, S'/ω_B is small, and the terms of the series in (4) are of rapidly decreasing magnitude. Only the first two terms need therefore be considered, and, if D is very large, only the first term.

It is convenient to express the transfer characteristic in polar form: $T(u) = \exp\{A(u) + j\phi(u)\}$, where $A(u)$ is the amplitude characteristic (nepers), and $\phi(u)$ is the phase characteristic (radians). The second derivative $T''(u)$ is easily found and a common factor $T(\Delta\omega S/\omega_B)$ can be removed from the first two terms of (4), which can then be written (neglecting terms beyond the second) as

$$\begin{aligned} v_0 = \Bigg[&\exp A\left(\frac{\Delta\omega S}{\omega_B}\right) \exp j\left\{\omega_c t + \Delta\omega \int S dt\right. \\ &\left. + \phi\left(\frac{\Delta\omega S}{\omega_B}\right)\right\}\Bigg]\Bigg[1 + \frac{\Delta\omega S'}{2\omega_B^2}\left\{\phi''\left(\frac{\Delta\omega S}{\omega_B}\right)\right. \\ &+ 2A'\left(\frac{\Delta\omega S}{\omega_B}\right)\phi'\left(\frac{\Delta\omega S}{\omega_B}\right) - jA''\left(\frac{\Delta\omega S}{\omega_B}\right) \\ &\left. - j\left[A'\left(\frac{\Delta\omega S}{\omega_B}\right)\right]^2 + j\left[\phi'\left(\frac{\Delta\omega S}{\omega_B}\right)\right]^2\right\}\Bigg]. \end{aligned}$$

Of the series of terms in the square brackets, the imaginary part is small compared with 1, and the variable part of the real part is also small compared with 1. The series may therefore be replaced by $K \exp j\alpha$, where K is the real part, and α the imaginary part of the series. Then

$$v_0 = M \exp j\left(\omega_c t + \Delta\omega \int \omega_f dt\right)$$

where

$$M = \left[\exp A\left(\frac{\Delta\omega S}{\omega_B}\right)\right]\left[1 + \frac{\Delta\omega S'}{2\omega_B^2}\left\{\phi''\left(\frac{\Delta\omega S}{\omega_B}\right) + 2A'\left(\frac{\Delta\omega S}{\omega_B}\right)\phi'\left(\frac{\Delta\omega S}{\omega_B}\right)\right\}\right]$$

$$\omega_d = S + \frac{1}{\Delta\omega}\cdot\frac{d}{dt}\cdot\phi\left(\frac{\Delta\omega S}{\omega_B}\right) - \frac{1}{2\omega_B^2}\frac{d}{dt}\left[S'\left\{A''\left(\frac{\Delta\omega S}{\omega_B}\right) + \left[A'\left(\frac{\Delta\omega S}{\omega_B}\right)\right]^2 - \left[\phi'\left(\frac{\Delta\omega S}{\omega_B}\right)\right]^2\right\}\right]. \quad (5)$$

M is the amplitude and $\Delta\omega\omega_j$ the frequency deviation of v_0.

If the transfer characteristics can be represented by simple functions and the modulating wave is also simple, it is sometimes possible to calculate the distortion directly from (5). In general, however, this is not possible, and the transfer functions have to be expressed in a form amenable to computation, e.g., a power series expansion, thus:

$$A(u) = A_0 + uA_1 + \frac{u^2}{2!}A_2 + \cdots$$
$$\phi(u) = \phi_0 + u\phi_1 + \frac{u^2}{2!}\phi_2 + \cdots. \quad (6)$$

The quantity ϕ_1 is the coefficient of the linear part of the phase characteristic, and thus represents time delay for the whole wave. It is advantageous to proceed as if ϕ_1 were zero and to correct the final result, if required, for the time delay corresponding to ϕ_1. This reduces the number of terms to be handled. The coefficients A_0 and ϕ_0, which represent a constant amplitude change and a constant phase shift of the carrier, may also without error be equated to zero.

The series expansion should represent the characteristic accurately over a sufficient range of u; namely, a range corresponding to frequencies slightly beyond the frequency excursion of the modulated carrier wave.

When the series given by (6) are substituted in (5), the distortion terms may be divided into three groups. First, a linear term which is simply a derivative of S. This is

$$-\frac{S''}{2\omega_B^2}(A_1^2 + A_2).$$

Next, even-order nonlinear terms,

$$\frac{1}{\Delta\omega}\cdot\frac{d}{dt}\left\{\frac{\phi_2}{2!}\left(\frac{\Delta\omega S}{\omega_B}\right)^2 + \frac{\phi_4}{4!}\left(\frac{\Delta\omega S}{\omega_B}\right)^4 + \cdots\right\}$$
$$-\frac{1}{2\omega_B^2}\frac{d}{dt}\left[S'\left\{\frac{\Delta\omega S}{\omega_B}(A_3 + 2A_1A_2) + \left(\frac{\Delta\omega S}{\omega_B}\right)^3\left(\frac{1}{6}A_5 + \frac{1}{3}A_4A_1 + A_3A_2 - \phi_3\phi_2\right) + \cdots\right\}\right]. \quad (7)$$

Finally, the odd-order nonlinear terms,

$$\frac{1}{\Delta\omega}\frac{d}{dt}\left\{\frac{\phi_3}{3!}\left(\frac{\Delta\omega S}{\omega_B}\right)^3 + \frac{\phi_5}{5!}\left(\frac{\Delta\omega S}{\omega_B}\right)^5 + \cdots\right\}$$
$$-\frac{1}{2\omega^2_B}\frac{d}{dt}\left[S'\left\{\left(\frac{\Delta\omega S}{\omega_B}\right)^2\left(\frac{1}{2}A_4 + A_3A_1 + A_2^2 - \phi_2^2\right) + \left(\frac{\Delta\omega S}{\omega_B}\right)^4\left(\frac{1}{24}A_6 + \frac{1}{12}A_5A_1 + \frac{1}{3}A_4A_2 + \frac{1}{4}A_3^2 - \frac{1}{3}\phi_4\phi_2 - \frac{1}{4}\phi_3^2\right)\right\}\right]. \quad (8)$$

If the amplitude characteristic is symmetrical and the phase characteristic skew-symmetrical, the even-order terms and some coefficients of the odd-order terms vanish.

V. Solution for Small Deviation Ratios

To obtain the formula for small deviation ratios it is convenient to express the transfer characteristic in Cartesian form: $T(u) = P(u) + jQ(u)$. $P(u)$, and $Q(u)$ are the in-phase and quadrature components, respectively. Equation (4) may then be written

$$v_0 = \left[\exp j\left(\omega_c t + \Delta\omega\int S dt\right)\right][1 + R + jI]$$

$$R = P\left(\frac{\Delta\omega S}{\omega_B}\right) - 1 + \frac{\Delta\omega}{\omega_B}\left\{\frac{S'}{2!\omega_B}Q''\left(\frac{\Delta\omega S}{\omega_B}\right) - \frac{S''}{3!\omega_B^2}P'''\left(\frac{\Delta\omega S}{\omega_B}\right) - \cdots\right\}$$
$$-\frac{\Delta\omega^2 S'}{2\omega_B^3}\left\{\frac{S'}{4\omega_B}P^{IV}\left(\frac{\Delta\omega S}{\omega_B}\right) + \frac{S''}{3!\omega_B^2}Q^{V}\left(\frac{\Delta\omega S}{\omega_B}\right) - \cdots\right\}$$

$$I = Q\left(\frac{\Delta\omega S}{\omega_B}\right) - \frac{\Delta\omega}{\omega_B}\left\{\frac{S'}{2!\omega_B}P''\left(\frac{\Delta\omega S}{\omega_B}\right) + \frac{S''}{3!\omega_B}Q'''\left(\frac{\Delta\omega S}{\omega_B}\right) - \cdots\right\}$$
$$-\frac{\Delta\omega^2}{2\omega_B^3}S'\left\{\frac{S'}{4\omega_B}Q^{IV}\left(\frac{\Delta\omega S}{\omega_B}\right) - \frac{S''}{3!\omega_B^2}P^{V}\left(\frac{\Delta\omega S}{\omega_B}\right) - \cdots\right\}. \quad (9)$$

Since the deviation ratio is small, $\Delta\omega/\omega_B$ is small. Also, from the relations between the polar and Cartesian forms of the transfer characteristic given in the

Appendix, $P(0)=1$ when $A(0)=0$, the condition assumed in the previous section. It follows that both R and I are small, compared with 1.

$$\begin{aligned}\therefore\quad 1+R+jI &= \{(1+R)^2+I^2\}^{1/2}\exp j\{\tan^{-1} I/(1+R)\}\\ &\doteqdot (1+R)\exp jI(1-R).\end{aligned}$$

Equation (4) then becomes

$$v_0=(1+R)\exp j\left(\omega_c t+\Delta\omega\int\omega_d dt\right)$$

where

$$\omega_d = S+\frac{1}{\Delta\omega}\frac{d}{dt}\cdot I(1-R). \qquad (10)$$

It is now assumed that $P(u)$ and $Q(u)$ can be expressed in the form of power series

$$P(u) = P_0+uP_1+\frac{u^2}{2!}P_2+\cdots$$

$$Q(u) = Q_0+uQ_1+\frac{u^2}{2!}Q_2+\cdots. \qquad (11)$$

In the previous section it was shown that it is permissible and advantageous to write $A_0=\phi_0=\phi_1=0$. From the relations given in the Appendix, the corresponding conditions for the Cartesian form are $P_0=1$ $Q_0=Q_1=0$.

On substituting the series of (11) into the expressions for R and I given by (9), the frequency deviation ω_d can be expanded in a series. Since $\Delta\omega/\omega_B$ is small, only terms with coefficients proportional to $\Delta\omega/\omega_B$ and $(\Delta\omega/\omega_B)^2$ need be considered in addition to terms independent of $\Delta\omega$. The linear distortion terms are

$$-\frac{S''P_2}{2!\omega_B^2}-\frac{S'''Q_3}{3!\omega_B^3}+\frac{S^{IV}P_4}{4!\omega_B^4}+\cdots.$$

The even-order nonlinear terms are

$$\begin{aligned}&\frac{\Delta\omega}{\omega_B^2}\frac{d}{dt}\left[S\left\{\frac{SQ_2}{2}-\frac{S'}{2!\omega_B}(P_3-P_2P_1)-\frac{S''}{3!\omega_B^2}(Q_4-Q_3P_1)\right.\right.\\ &\qquad\left.\left.+\frac{S'''}{4!\omega_B^3}(P_5-P_4P_1)+\cdots\right\}\right]\\ &-\frac{\Delta\omega}{2\omega_B^3}\frac{d}{dt}\left[S'\left\{\frac{S'}{2!\omega_B}\left(\frac{1}{2}Q_4-P_2Q_2\right)\right.\right.\\ &\qquad-\frac{S''}{3!\omega_B^2}(P_5+Q_3Q_2-P_3P_2)\\ &\qquad\left.\left.-\frac{S'''}{4!\omega_B^3}(Q_6-P_4Q_2-Q_4P_2)+\cdots\right\}\right] \qquad (12)\end{aligned}$$

and the odd-order nonlinear terms are

$$\begin{aligned}&\frac{1}{2}\frac{\Delta\omega^2}{\omega_B^3}\frac{d}{dt}\left[S^2\left\{S\left(\frac{1}{3}Q_3-Q_2P_1\right)\right.\right.\\ &\qquad-\frac{S'}{2!\omega_B}(P_4-P_2^2-2P_3P_1+Q_2^2)\\ &\qquad-\frac{S''}{3!\omega_B^2}(Q_5-Q_3P_2-2Q_4P_1-P_3Q_2)\\ &\qquad\left.\left.+\frac{S'''}{4!\omega_B^3}(P_6-P_4P_2-2P_5P_1+Q_4Q_2)+\cdots\right\}\right]\\ &-\frac{1}{2}\frac{\Delta\omega^2}{\omega_B^4}\frac{d}{dt}\left[SS'\left\{\frac{S'}{2!\omega_B}\left(\frac{1}{2}Q_5-Q_3P_2\right.\right.\right.\\ &\qquad\left.-\frac{1}{2}Q_4P_1-P_3Q_2\right)\\ &\qquad-\frac{S''}{3!\omega_B^2}(P_6-P_4P_2+Q_3^2-P_5P_1+Q_4Q_2-P_3^2)\\ &\qquad-\frac{S'''}{4!\omega_B^3}(Q_7-Q_5P_2-P_4Q_3\\ &\qquad\left.\left.-Q_6P_1-P_5Q_2-Q_4P_3)+\cdots\right\}\right]. \qquad (13)\end{aligned}$$

The first term in each of the series in (13) has an anomalous value, but all the following terms form a regular sequence.

When the in-phase characteristic is symmetrical and the quadrature characteristic skew-symmetrical, the even-order terms and half of the coefficients of the odd-order terms vanish.

VI. Discussion of Results

For both large and small deviation ratios, the amplitudes of the linear distortion terms are independent of the frequency deviation. These terms which represent phase and frequency distortion of the modulating wave are not usually of interest.

For large deviation ratios, the distortion given by (5) can be divided into two parts. The first and principal part, $(1/\Delta\omega)\cdot(d/dt)\phi(\Delta\omega S/\omega_B)$, depends only on the phase characteristic; the second part is determined mainly by the amplitude characteristic and partly by the phase characteristic. As the deviation ratio is large, S'/ω_B is small, so that the distortion produced by the amplitude characteristic is usually small compared with that produced by the phase characteristic. Moreover, if the modulating wave is the sum of a number of cosine waves, it is clear from (5) that the principal part of the distortion is the sum of a number of sine waves, whereas the second part is the sum of a number of cosine waves. Thus, the distortion products due to the amplitude characteristic are in phase quadrature with the principal distortion products produced by the phase characteristic, and so have little effect on the total distortion until their magnitude equals or exceeds that of the principal part of the distortion.

If the deviation ratio D is increased while $\Delta\omega/\omega_B$ is kept constant, the principal part of the distortion decreases in the ratio $1/D$ and the amplitude characteristic distortion in the ratio $1/D^2$. As the deviation ratio is increased, therefore, the distortion is ultimately entirely nonlinear in character, and is produced by the nonlinear part of the phase characteristic.

For small deviation ratios the limits of the frequency spectrum of a modulated carrier wave are determined by the spectrum of the modulating wave rather than by the maximum frequency deviation. Consequently, ω_B cannot be reduced below a certain minimum value. From the formulas of Section V it is then seen that as $\Delta\omega$ is reduced the distortion is ultimately entirely linear in character, is independent of the maximum frequency deviation, and is produced by the symmetrical part of the in-phase characteristic and the skew-symmetrical part of the quadrature characteristic.

VII. Nature of the Distortion in a Frequency-Modulation Network

If the input to a nonlinear vacuum-tube amplifier with a resistive load is the sum for a number of cosine waves, the output from the amplifier, including the distortion products, is also the sum of a number of cosine waves. In the previous section it was noted that the principal distortion products for large deviation ratios are the sum of a number of sine waves. Thus, in a frequency-modulation network the principal distortion products are in phase quadrature with the corresponding distortion products in a nonlinear amplifier.

A convenient method of specifying and measuring distortion in low-frequency apparatus is the intermodulation method, in which two sine waves, one of low frequency and amplitude nearly equal to the capacitance of the apparatus, the other of high frequency and small amplitude, are applied simultaneously to the apparatus. Nonlinear distortion results in the amplitude of the high-frequency component being modulated by the low-frequency component, the amount of such modulation being a measure of the distortion. In frequency-modulation systems, however, the intermodulation distortion is of an entirely different character.

Let $S=C_1\cos\omega_1 t+C_2\cos\omega_2 t$ where $C_2\ll C_1$ and $\omega_1\ll\omega_2$, and suppose that the deviation ratio is large so that only the principal part of the phase-characteristic distortion is significant. Then, from (5), the frequency deviation is

$$C_1\cos\omega_1 t + C_2\cos\omega_2 t + \frac{1}{\Delta\omega}\frac{d}{dt}\phi\left(\frac{\Delta\omega}{\omega_B}C_1\cos\omega_1 t+\frac{\Delta\omega}{\omega_B}C_2\cos\omega_2 t\right).$$

Since C_2 is small, the last term can be written approximately (because $C_1+C_2=1$) as

$$\frac{1}{\Delta\omega}\frac{d}{dt}\phi\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right) + \frac{C_2}{\omega_B}\frac{d}{dt}\left\{\cos\omega_2 t\,\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right)\right\}.$$

The first term in this expression represents harmonics of the low-frequency component. The second term represents the intermodulation products, and, since $\omega_2\gg\omega_1$, this term is

$$-\frac{C_2\omega_2}{\omega_B}\sin\omega_2 t\,\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right).$$

Adding to this the component of frequency ω_2 in the frequency deviation gives

$$C_2\left[\cos\omega_2 t-\frac{\omega_2}{\omega_B}\sin\omega_2 t\,\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right)\right]$$

$$\doteq C_2\cos\left\{\omega_2 t+\frac{\omega_2}{\omega_B}\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right)\right\}. \tag{14}$$

Nonlinear distortion is manifest as a modulation of the high-frequency component, not in amplitude, but in frequency (or phase) by the low-frequency component. The amount of the modulation depends not only on the amplitude of the low-frequency component, but is also directly proportional to the frequency of the high-frequency component.

Since the intermodulation is of frequency instead of amplitude, the intermodulation products are in phase-quadrature with the corresponding components in the vacuum-tube-amplifier case. Listening tests have shown, as would be expected, that the ear is unable to distinguish this phase difference, provided the distortion is not too great. Accordingly, a sound wave given by (14) produces the same aural effect as a wave given by

$$C_2\left\{1+\frac{\omega_2}{\omega_B}\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right)\right\}\cos\omega_2 t$$

in which the high-frequency component $C_2\cos\omega_2 t$ is modulated in amplitude.

It is, therefore, permissible to specify the distortion in a frequency-modulation network as intermodulation distortion, the magnitude being the maximum deviation of

$$1+\frac{\omega_2}{\omega_B}\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right)$$

from its mean value, but it must be remembered that the modulation is of frequency and cannot be measured by the same methods as are used for amplitude intermodulation. A suitable method is described in the Appendix.

VIII. Methods of Calculation and Examples

If the transfer characteristic can be represented by simple functions and the modulating wave is also simple, it is sometimes possible to calculate the distortion directly from expression (5). An example of such a calculation is given below. In general, however, this is not possible, and the transfer functions have to be expressed in power series form. The distortion is then calculated from (7), (8), (12), and (13).

If the power series for either the polar or Cartesian form of the transfer characteristic are given, the series for the other form may be obtained from the relations between coefficients given in the Appendix.

Finally, quantities of the form S^n have to be evaluated. For the purpose of analysis, a modulating wave which yields a fair amount of information without too much labor is the sum of two cosine (or sine) waves of different or equal amplitudes. A method of expanding the expression $(k_1 \cos \omega_1 t + k_2 \cos \omega_2 t)^n$ in a series of terms of the type $A_{pq} \cos (p\omega_1 \pm q\omega_2)t$ is given in the Appendix.

The intermodulation distortion is found by calculating the maximum or minimum value and the mean value of

$$\frac{\omega_2}{\omega_B}\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right).$$

Example 1

A high-frequency carrier wave, frequency-modulated by a 5-kc. cosine wave, is applied to a network consisting of a single parallel-resonant circuit such that the amplitude response is -3 db at frequencies differing by ± 25 kc. from the carrier frequency. Find the third-harmonic distortion in the frequency deviation of the output wave as the maximum deviation is varied from 10 to 100 kc.

The transfer characteristic of the network is $T(u) = (1+ju)^{-1} \exp ju$, and $\omega_B = 25$ kc. The factor $\exp ju$ is added to satisfy the condition that the phase characteristic should have no linear part. Then $A(u) = -\frac{1}{2}\log h\,(1+u^2)$ and $\phi(u) = u - \tan^{-1} u$.

Let the modulating wave be $\cos \omega_m t$. For large deviation ratios, equation (5) is used. Now $A''(u) + \{A'(u)\}^2 - \{\phi'(u)\}^2 = -(u^2-1)^2(u^2+1)^{-2}$, so that (5) becomes

$$\cos\omega_m t + \frac{1}{\Delta\omega}\cdot\frac{d}{dt}\left\{\frac{\Delta\omega}{\omega_B}\cos\omega_m t - \tan^{-1}\left(\frac{\Delta\omega}{\omega_B}\cos\omega_m t\right)\right\}$$
$$-\frac{\omega_m}{2\omega_B^2}\frac{d}{dt}\left[\sin\omega_m t\left\{\left(\frac{\Delta\omega^2}{\omega_B^2}\cos^2\omega_m t - 1\right)^2\right.\right.$$
$$\left.\left.\cdot\left(\frac{\Delta\omega^2}{\omega_B^2}\cos^2\omega_m t + 1\right)^{-2}\right\}\right]. \qquad (15)$$

Now $\quad \tan^{-1}\left(\dfrac{\Delta\omega}{\omega_B}\cos\omega_m t\right)$

may be expanded in a Fourier series from the formula[4]

$$\tan^{-1}\left(\frac{2a\cos x}{1-a^2}\right) = 2\sum_1^\infty (-1)^{n-1}\frac{a^{2n-1}}{2n-1}\cos(2n-1)x$$

by writing

$$a = \left\{1 + \frac{\omega_B^2}{\Delta\omega^2}\right\}^{1/2} - \frac{\omega_B}{\Delta\omega}.$$

In the third term the factor

$$\left(\frac{\Delta\omega^2}{\omega_B^2}\cos^2\omega_m t + 1\right)^{-2}$$

may be expanded by the formula[5]

$$(1-b^2)^3(1+b^2+2b\cos x)^{-2}$$
$$= 1 + b^2 + 2\sum_1^\infty(-b)^n\{n+1-b^2(n-1)\}\cos nx$$

by writing $b = a^2$. It is then a matter of straightforward trigonometry to find the terms of third-harmonic frequencies in (15). These are

$$-\frac{2\omega_m}{\Delta\omega}b^{3/2}\sin 3\omega_m t$$
$$+\frac{3}{2}\left(\frac{\omega_m}{\omega_B}\right)^2\left\{1+\frac{1}{2}\frac{\Delta\omega^2}{\omega_B^2}\right\}^{-2}(1+b^2)^2(1-b^2)^{-3}$$
$$\cdot\left\{b(2-b)(1+b)^2 + \frac{1}{2}\frac{\Delta\omega^2}{\omega_B^2}(1-2b)(1-b^2)^2\right.$$
$$\left.-\frac{3}{16}\frac{\Delta\omega^4}{\omega_B^4}(1-b)^2(1-b^2)^2\right\}\cos 3\omega_m t. \qquad (16)$$

The amplitude of the resultant is the square root of the sum of the squares of the amplitudes of sine and cosine terms.

For small deviation ratios the Cartesian form of the transfer characteristic is used. If $(1+ju)^{-1}\exp ju$ is expanded in a series of powers of u, then

$$P(u) = 1 - \frac{u^2}{2!} + \frac{9u^4}{4!} - \frac{265u^6}{6!} + \frac{14833}{8!}u^8\cdots$$

$$Q(u) = \frac{2u^3}{3!} - \frac{44u^5}{5!} + \frac{1854}{7!}u^7\cdots.$$

Hence,

$$P_2 = -1 \quad P_4 = 9 \quad P_6 = -265 \quad P_8 = 14833$$
$$Q_3 = 2 \quad Q_5 = -44 \quad Q_7 = 1854.$$

On substituting these values into (13), the terms of third-harmonic frequency are found to be

[4] "Smithsonian Mathematical Formulae and Tables of Elliptic Functions," Smithsonian Institution, 1939; p. 140.

[5] J. Edwards, "The Integral Calculus," Macmillan Co., London, 1922; vol. 2, p. 303.

$$-\frac{\Delta\omega^2\omega_m}{4\omega_B^3}\left[1-\frac{51}{2}\frac{\omega_m^2}{\omega_B^2}+\frac{1077}{8}\frac{\omega_m^4}{\omega_B^4}\right]\sin 3\omega_m t$$

$$+\frac{3}{2}\frac{\Delta\omega^2\omega_m^2}{\omega_B^4}\left[1-\frac{79}{6}\frac{\omega_m^2}{\omega_B^2}+\frac{1409}{40}\frac{\omega_m^4}{\omega_B^4}\right]\cos 3\omega_m t. \qquad (17)$$

Finally, putting $\omega_m = 2\pi 5000$, $\omega_B = 2\pi 25{,}000$, the amplitude of the third-harmonic component is $0.033(\Delta\omega^2/\omega_B^2)$.

On Fig. 1 is shown the amplitude of the third harmonic calculated from (16) and (17). The same figure shows the experimental and theoretical results obtained by Jaffe. Jaffe's theoretical values correspond to the sine term in (16), i.e., to the distortion due to the phase characteristic.

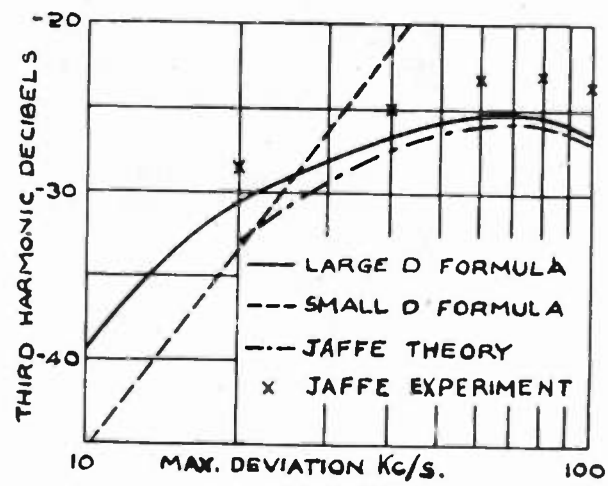

Fig. 1—Third-harmonic distortion in a single resonant circuit.

Example 2

The frequency-selective circuits of an amplifier consist of three identical band-pass filters connected in cascade via amplifier tubes. Each filter is made up of two identical simple resonant circuits critically coupled. The over-all amplitude response of the amplifier is -6 db at frequencies differing by ± 100 kc. from the midband frequency.

Estimate the intermodulation distortion at a frequency of 12 kc. when $\Delta\omega = 75$ kc.,

(a) when the carrier frequency is equal to the midband frequency; and

(b) when the carrier frequency differs by 25 kc. from the midband frequency.

A carrier wave modulated by two cosine waves of equal amplitudes and of frequencies 3 kc. and 5 kc. is applied to the amplifier. The maximum deviation is 75 kc.

(c) Find the amplitude of the distortion product of frequency 11 kc. when the carrier frequency is equal to the midband frequency.

(d) Find the amplitude of the distortion product of frequency 8 kc. when the carrier frequency differs by 25 kc. from the midband frequency.

The transfer characteristic for a pair of identical resonant circuits critically coupled is $(1-\frac{1}{2}\alpha^2u^2+j\alpha u)^{-1}$ $\exp j\alpha u$, where α is a constant depending on the bandwidth. Taking ω_B as 100 kc., the transfer characteristic for the complete amplifier is $T(u)=(1-0.769u^2+j1.24u)^{-3}$ $\exp j3.72u$, from which $\phi(u)=+3.72u-3\tan^{-1}\{1.24u/(1-0.769u^2)\}$. The Gregory series for $\tan^{-1}x$ converges for $|x|\leq 1$, corresponding to $|u|\leq 0.59$. By expressing $\phi(u)$ as an inverse sine, a series can be found which converges for $|u|\leq 1.14$, but the convergence is very slow. However, it is found that for values of u up to 1, $\phi(u)$ can be approximated by the expression $\phi(u)=-0.95u^3+0.51u^5-0.017u^7$.

When the carrier frequency is shifted from the midband frequency of the amplifier by 25 kc., this is equivalent to shifting the working point of the transfer characteristic from 0 to $25/100=\frac{1}{4}$. Consequently, the new phase characteristic is obtained by writing $u+\frac{1}{4}$ in place of u in the above expression for $\phi(u)$. Omitting the linear and constant terms, this is

$$-0.63u^2-0.63u^3+0.63u^4+0.49u^5-0.030u^6-0.017u^7.$$

The amplitude characteristic has a negligible effect on the distortion, and is therefore ignored.

The intermodulation distortion is the maximum deviation of

$$\frac{\omega_2}{\omega_B}\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right)=0.12\phi'(0.75\cos\omega_1 t),$$

from its mean value. In case (a) the mean value is, from (20), -0.061. The maximum value is obviously 0, and the minimum value is easily shown to be -0.096; the intermodulation distortion is, therefore, 6 per cent. In case (b) the mean value is -0.03, and the maximum and minimum values are 0.02 and -0.076. The intermodulation distortion is now 5 per cent.

Cases (c) and (d) are solved from (8) and (7), using the expressions for $\phi(u)$ given above, and evaluating the amplitudes of the distortion products from (20), with $p=2$, $q=1$ in case (c), and $p=q=1$ in case (d). The results are: 0.0038 and 0.0056. In case (c) the distortion products are all of the odd-order type, i.e., of the form $a_{pq}\cos(p\omega_1 \pm q\omega_2)t$ where $p+q$ is odd, but in case (d) the even-order type is predominant.

IX. Distortion in a Receiver with Negative Feedback

Negative feedback may be applied to a superheterodyne receiver by arranging that the detector output operates a modulator, which controls the frequency of the receiver local oscillator, in such a way that the frequency deviation of the received wave is reduced before the wave is amplified at the intermediate frequency. This arrangement was first described by Chaffee.[6]

[6] J. G. Chaffee, "The application of negative feedback to frequency modulation systems," *Bell Sys. Tech. Jour.*, vol. 18, pp. 404–438; July, 1939.

In what follows it is supposed that the detector output is equal to the frequency deviation of the wave applied to it, and that the modulator is also free from distortion. In practice, it is the modulator distortion which sets the limit to reduction in distortion obtainable by feedback.

Let the frequency deviation of the received wave be $\Delta\omega_1 S(t)$, and let the detector output be $O(t)$. The frequency deviation of the local oscillator is $\beta O(t)$, where β is a constant, and the frequency deviation of the wave of intermediate frequency is $\Delta\omega_1 S(t)-\beta O(t)$. The effect of the i.f. amplifier is to delay the modulation by a time T, and to add distortion, so that the frequency deviation of the wave arriving at the detector, and hence the detector output, is $\Delta\omega_1 S(t-T)-\beta O(t-T)+KD(t-T)=O(t)$. $D(t-T)$ is the relative distortion in the frequency deviation, and K is equal to the maximum deviation at intermediate frequency.

By expanding $O(t-T)$ as a series of derivatives of $O(t)$, an equation is obtained expressing $O(t)$ in terms of $S(t-T)$, $D(t-T)$, and derivatives of $O(t)$. On differentiating this equation and eliminating $O'(t)$ between this equation and the first one, a new equation is obtained from which $O'(t)$ is absent. Proceeding in this way, all the derivatives of $O(t)$ can be eliminated and the other terms collected together to give

$$O(t)=\frac{\Delta\omega_1}{1+\beta}\left[S(t-T/(1+\beta))+\frac{1}{1+\beta}D(t-T/(1+\beta))\right]$$

$$+\text{terms of higher order.}$$

The higher-order terms are derivatives of $S(t)$ and $D(t)$ and are always of negligible magnitude. The term $D(t-T/(1+\beta))$ represents the distortion suffered by a wave whose frequency deviation is $\Delta\omega S(t-T/(1+\beta))$ where $\Delta\omega=\Delta\omega_1/(1+\beta)$. Under different conditions which will now be examined, the magnitude of

$$\frac{1}{1+\beta}D(t-T/(1+\beta))$$

is more or less changed when β is varied.

First, let the bandwidth remain constant as β is increased from zero. Then, from (7), (8), (12), and (13), the quadratic distortion (S^2, SS' etc.) is proportional to $\Delta\omega/(1+\beta)$, i.e., to $(1+\beta)^{-2}$, the cubic distortion to $(1+\beta)^{-3}$, and generally the nth order distortion is proportional to $(1+\beta)^{-n}$. Thus, if the application of feedback reduces the deviation of the received wave by N decibels, the quadratic distortion is reduced by $2N$ decibels and the cubic distortion by $3N$ decibels.

Next, suppose that, as the maximum deviation is reduced by feedback, the bandwidth of the i.f. amplifier is reduced in the same ratio, the shape of the transfer characteristic being kept constant. This is possible so long as the reduced bandwidth is greater than twice the highest modulating frequency. From (5) it is seen that the distortion of all orders due mainly to the phase characteristic remains constant, and the distortion due mainly to the amplitude characteristic increases in the ratio $1+\beta$.

The third case to be considered is that in which the deviation ratio is initially large, and the feedback is sufficiently great to reduce the deviation ratio to less than 1. As feedback is applied the bandwidth is reduced from its initial value of $2\Delta\omega_1$ to the limiting value $2\omega_q$. The distortions before and after feedback are not directly comparable on a numerical basis, since the nature of the distortion changes. However, in some cases the distortion for both large and small deviation ratios is due mainly to one particular term. As an example of typical distortion components, the term

$$\frac{1}{\Delta\omega}\frac{d}{dt}\cdot\frac{\phi_3}{3!}\left(\frac{\Delta\omega S}{\omega_B}\right)^3$$

in (8) and the corresponding term

$$\frac{1}{2}\frac{\Delta\omega^2}{\omega_B^2}\frac{d}{dt}S^3(1/3Q_3-Q_2P_1)$$

in (13) may be taken. From the relations between the polar and Cartesian forms of the transfer characteristic given in the Appendix, these two terms are equal.

Initially ($\beta=0\ \omega_B=\Delta\omega_1$), the distortion is

$$\frac{\phi_3}{2}\frac{S^2S'}{\Delta\omega_1},$$

and finally ($\omega_B=\omega_q$), it is

$$\frac{\phi_3}{2}\frac{\Delta\omega_1^2}{\omega_q^3}\frac{S^2S'}{(1+\beta)^3}.$$

The distortion is therefore reduced in the ratio

$$\left\{\frac{\Delta\omega_1}{\omega_q(1+\beta)}\right\}^3=[D/(1+\beta)]^3.$$

Appendix

A. Relations between the polar and Cartesian forms of the transfer characteristic

The steady-state transfer characteristic $T(u)$ may be written $T(u)=exp\{A(u)+j\phi(u)\}=P(u)+jQ(u)$, where $A(u)$ is the amplitude characteristic in nepers, $\phi(u)$ the phase characteristic in radians, and $P(u)$ and $Q(u)$ are the in-phase and quadrature components of the characteristic.

Then

$$P(u)=\exp A(u)\cos\phi(u)$$

$$Q(u)=\exp A(u)\sin\phi(u)$$

$$A(u)=1/2\ \operatorname{log}h[P^2(u)+Q^2(u)]$$

$$\phi(u)=\tan^{-1}[Q(u)/P(u)].$$

It is assumed that $A(u)$, $\phi(u)$, $P(u)$, and $Q(u)$ may be expressed as power series of the form

$$A(u) = A_0 + uA_1 + \frac{u^2}{2!}A_2 + \frac{u^3}{3!}A_3 + \cdots.$$

The coefficient A_0 represents the gain of the network at the reference frequency. Since this is arbitrary, it is convenient to put $A_0=0$. The coefficient ϕ_0 represents a constant phase change of the carrier wave, which is of no interest, and ϕ_1 represents a time delay of the modulation impressed on the carrier which may be allowed for, if necessary, in the final result.

On the assumption that $A_0=\phi_0=\phi_1=0$, the relations between the coefficients in the power series for $A(u)$, $\phi(u)$, $P(u)$, and $Q(u)$, are as follows:

$$P_0 = 1$$
$$P_1 = A_1$$
$$P_2 = A_2 + A_1^2$$
$$P_3 = A_3 + 3A_2A_1 + A_1^3$$
$$P_4 = A_4 + 4A_3A_1 + 3A_2^2 + 6A_2A_1^2 + A_1^4 - 3\phi_2^2$$
$$P_5 = A_5 + 5A_4A_1 + 10A_3A_2 + 10A_3A_1^2 + 15A_2^2A_1 + 10A_2A_1^3 + A_1^5 - 15A_1\phi_2^2 - 10\phi_3\phi_2$$

$$Q_0 = Q_1 = 0$$
$$Q_2 = \phi_2$$
$$Q_3 = \phi_3 + 3\phi_3A_1$$
$$Q_4 = \phi_4 + 4\phi_3A_1 + 6\phi_2A_2 + 6\phi_2A_1^2$$
$$Q_5 = \phi_5 + 5\phi_4A_1 + 10\phi_3A_2 + 10\phi_3A_1^2 + 10\phi_2A_3 + 30\phi_2A_2A_1 + 10\phi_2A_1^3$$

$$A_1 = P_1$$
$$A_2 = P_2 - P_1^2$$
$$A_3 = P_3 - 3P_2P_1 + 2P_1^3$$
$$A_4 = P_4 - 4P_3P_1 - 3P_2^2 + 12P_2P_1^2 - 6P_1^4 + 3Q_2^2$$
$$A_5 = P_5 - 5P_4P_1 - 10P_3P_2 + 20P_3P_1^2 + 30P_2^2P_1 - 60P_2P_1^3 + 24P_1^5 - 30P_1Q_2^2 + 10Q_3Q_2$$

$$\phi_2 = Q_2$$
$$\phi_3 = Q_3 - 3Q_2P_1$$
$$\phi_4 = Q_4 - 4Q_3P_1 - 6Q_2P_2 + 12Q_2P_1^2$$
$$\phi_5 = Q_5 - 5Q_4P_1 - 10Q_3P_2 + 20Q_3P_1^2 - 10Q_2P_3 + 60Q_2P_2P_1 - 60Q_2P_1^3.$$

B. Calculation of harmonics and intermodulation products

If the modulating wave is comprised of two cosine waves of different amplitudes, then in calculating distortion products it is necessary to expand expressions of the form $(k_1 \cos\theta_1 + k_2 \cos\theta_2)^n$ in terms of unit powers of multiple angles, n being a positive integer. This can be done most conveniently by expressing the cosine terms in exponential form and applying the multinomial theorem to expand the result.

Thus, if

$$y = [k_1 \cos\theta_1 + k_2 \cos\theta_2]^n$$
$$= \frac{1}{2^n}[k_1 \exp j\theta_1 + k_1 \exp -j\theta_1 + k_2 \exp j\theta_2 + k_2 \exp -j\theta_2]^n,$$

then, by the multinomial theorem,

$$y = \frac{n!}{2^n}\times \sum \frac{k_1^{(\alpha_1+\alpha_2)}k_2^{(\alpha_3+\alpha_4)}\exp j\{(\alpha_1-\alpha_2)\theta_1+(\alpha_3-\alpha_4)\theta_2\}}{\alpha_1!\alpha_2!\alpha_3!\alpha_4!}. \tag{18}$$

α_1, α_2, α_3 and α_4 are positive integers or zero, and the summation extends over all possible values of α_1, α_2, α_3, and α_4 consistent with the relation

$$\alpha_1 + \alpha_2 + \alpha_3 + \alpha_4 = n. \tag{19}$$

The coefficient of $\frac{1}{2}\exp j(p\theta_1+q\theta_2)$, which is the same as that of $\cos(p\theta_1+q\theta_2)$, is obtained by putting $\alpha_1-\alpha_2=p$, $\alpha_3-\alpha_4=q$. The coefficient of $\cos(p\theta_1-q\theta_2)$ is the same as that of $\cos(p\theta_1+q\theta_2)$, since it can be formed simply by interchanging the values of α_3 and α_4.

The minimum value of α_2 and α_4 is zero. Let $\alpha_2=r$ then

$$\alpha_1 = p + r \quad \alpha_3 = \frac{n-p+q}{2} - r \quad \alpha_4 = \frac{n-p-q}{2} - r.$$

It is clear that r attains a maximum value when $\alpha_4=0$ and this maximum is $\frac{1}{2}(n-p-q)$. Substituting for α_1, α_2, α_3, and α_4, and summing over all possible values of r, the coefficient of $\cos(p\theta_1 \pm q\theta_2)$ given by (18) is

$$\frac{n!}{2^{n-1}}\sum_{r=0}^{1/2(n-p-q)} \frac{k_1^{(p+2r)}\,k_2^{(n-p-2r)}}{r!(p+r)!\left(\frac{n-p-q}{2}-r\right)!\left(\frac{n-p+q}{2}-r\right)!} \tag{20}$$

Only certain values can be taken by p and q. If n is even, $p+q$ must be even, and if n is odd, $p+q$ must be odd. Also, $p+q$ cannot be greater than n. If either k_1 or k_2 is zero, the expression has a value only when either $p+2r=0$ or $n-p-2r=0$. If n is even, y has a mean value which is $\frac{1}{2}$ of the value found by putting $p=q=0$ in (20).

C. Measurement of intermodulation distortion

It was shown in Section VII that, if a carrier wave modulated by a wave $C_1 \cos\omega_1 t + C_2 \cos\omega_2 t$ ($C_2 \ll C_1$ $\omega_1 \ll \omega_2$) is passed through a network having a nonlinear phase characteristic, the h.f. component, $C_2 \cos\omega_2 t$, of the modulating wave becomes modulated in frequency by a function of the l.f. component. The modulated high-frequency component is, from (14),

$$C_2 \cos \left\{\omega_2 t + \frac{\omega_2}{\omega_B}\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right)\right\}. \qquad (21)$$

The intermodulation distortion is defined as the maximum variation of the quantity

$$1 + \frac{\omega_2}{\omega_B}\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right)$$

from its mean value.

Suppose that the modulated carrier emerging from the network is applied to an ideal detector which yields an output proportional to the frequency deviation. From this output the component given by (21) is selected by means of a suitable filter and applied to a differentiating circuit (e.g., a series RC circuit of small time constant with the output taken across R) to produce a wave proportional to

$$-C_2\omega_2\left\{1 - \frac{\Delta\omega\omega_1}{\omega_B^2}\sin\omega_1 t\,\phi''\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right)\right\}$$
$$\cdot\sin\left\{\omega_2 t + \frac{\omega_2}{\omega_B}\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right)\right\}.$$

This wave is applied to an amplitude detector which produces an output proportional to

$$\omega_2\left\{1 - \frac{\Delta\omega\omega_1}{\omega_B^2}\sin\omega_1 t\,\phi''\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right)\right\}.$$

The alternating part of the output is filtered out and applied to an integrating circuit (a series RC circuit of large time constant with the output taken across C) to give an output proportional to

$$\frac{\omega_2}{\omega_B}\phi'\left(\frac{\Delta\omega}{\omega_B}\cos\omega_1 t\right),$$

the peak value of which may be mesaured by a vacuum-tube voltmeter. The voltmeter may be calibrated to read directly the intermodulation distortion.

A Study of Tropospheric Reception At 42.8 Mc. and Meteorological Conditions*

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Summary—From February, 1945, during its hours of operation, station W2XMN at Alpine, N. J., has been recorded at Needham, Mass., at a distance of 167 miles. W2XMN operates on a frequency of 42.8 Mc., at a power of 50 kw., and its daily schedule is from 1600 to 2300, E.D.S.T. in summer and E.S.T. in winter. Analysis of the Alpine recording has shown that no part of the ionosphere is involved in the transmission, which is purely tropospheric.

The Alpine fields show a marked seasonal change, being much higher in the summer than in the winter, and this has been found to be principally due to the seasonal changes in surface refraction along the transmission path. A controlling factor in the seasonal change of refraction is water-vapor pressure, which is at a maximum in the summer.

All types of frontal passage are found to lower transmission, and, presumably because of wave-guide effects, the amount of field depression caused by the passage of the front varies with the angle made by the front with the path. When the front is parallel with the path, the field is least depressed, but is lowest when the front makes a considerable angle with the path.

High fields at Needham are usually followed by an increase in surface temperature along the path, the temperature reaching a maximum about 30 hours after the field maximum. Conversely, low fields are generally followed by falling temperatures, which reach a minimum some 30 hours after the field minimum.

The best transmission along the Alpine-Needham path occurs when the wind velocity on the path is lowest, and the worst transmission accompanies high winds, probably because of turbulence which breaks up favorable stratification in the lower atmosphere.

Finally, the direction of air movement with respect to the path is related to transmission, Needham fields being higher when the wind is parallel with the path. The principal conditions favorable for transmission over this path are therefore summer, high surface refraction, rising temperatures, low wind velocities, winds parallel with the path, and an absence of frontal passages.

Introduction

TRANSMISSION from the f.m. station W2XMN, operating on 42.8 Mc., at Alpine, N. J., is received at Needham, Mass., distant 167 miles, on a half-wave dipole with reflector 50 feet above ground. In a conventional receiver a variable diode load is utilized to operate a Micromax single-pen recorder. The circuit is periodically calibrated with a Ferris Microvolter at 42.8 Mc. when W2XMN is off the air. Because of the large fading amplitudes, the fields are transcribed from the recorder charts as log microvolts at the receiver. One microvolt at the receiver equals approximately 0.7 microvolt per meter at the antenna. The charts are scaled for the median value of 20-minute intervals from which are derived hourly means and the mean nightly field for the seven hours during which Alpine is on the air each day. In the day-by-day comparisons given below with the tropospheric elements, a general use has been made of the ratio of the daily log field to a

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running mean of 27 days, to eliminate the large seasonal changes of field, surface refraction, etc. It is realized that the use herein of a simple ratio between a daily logarithmic value and a 27-day mean of logarithmic values gives a distorted ordinate scale. The use of this ratio is justified by the fact that the correlations involved are qualitative, rather than quantitative.

The analysis employed throughout this paper, with the exception of seasonal and monthly plots of field and refraction, is the well-known method of taking a maximum or minimum of one of the variables as an epoch, and determining the distribution of the other variable with respect to this epoch. If a sufficient number of cases is taken, the probable error of the mean values thus obtained is low enough to show any significant correlation of the variables.

It is appreciated by the authors, and will be immediately recognized by meteorologists, that the tropospheric elements here treated as independent variables are, in fact, rather closely interrelated, e.g., frontal passages and temperature changes. A more complete study than can here be given would therefore involve partial correlation between three or more variables.

Seasonal Variation

For the reason that Needham is some seven miles below the geometric line-of-sight from Alpine, as indicated by the path profile of Fig. 1 (and about half this value, or some 3.5 miles, for average or yearly mean

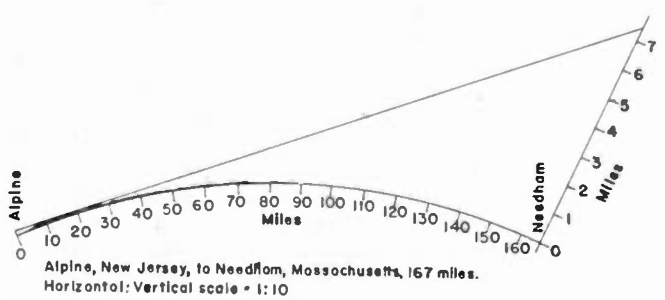

Fig. 1—Profile of Alpine-Needham path.

value of surface refraction on the path), the recorded fields are in general quite low, averaging only a few microvolts per meter and frequently falling below measurable values (under 1 microvolt) for periods ranging from seconds to hours. During periods of good transmission, peak values of the order of 20 microvolts per meter are frequently observed.

The seasonal changes in Alpine fields at Needham are shown in the lower curve of Fig. 2, covering the period February, 1945, to June, 1946, the daily values of log field being smoothed by a running mean of 27. A maximum appears in August, 1945, and a minimum in February, 1946. As the ordinates of this curve are logarithmic, the actual smoothed mean field values range from 0.63 in February to 5.4 microvolts in August, or a range of nearly 1 to 9. The upper curve of Fig. 2 is surface refraction, and will be considered below.

Field Intensity and Atmospheric Refraction

On long nonoptical paths it may be assumed that one of the principal controls is refractive bending of the

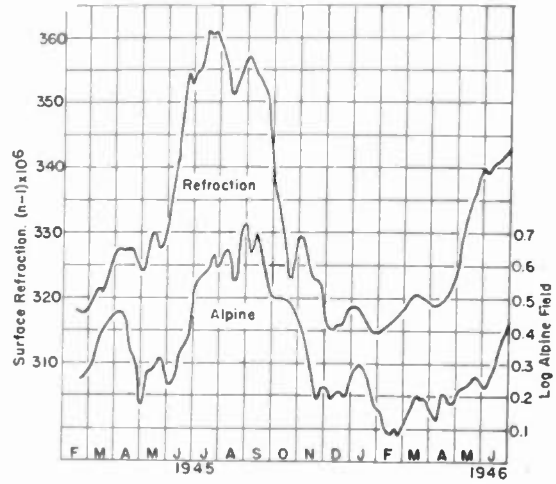

Fig. 2—Field reception of W2XMN on 42.8 Mc. at Needham, Mass., 1945–1946.

wave. The magnitude of this downward bending depends upon the lapse rate of atmospheric refractive index, and an exact evaluation of this would call for a detailed knowledge of the temperature, barometric pressure, and vapor pressure everywhere in a vertical strip of the troposphere including the entire transmission path. While such measurements are readily made by radiosondes, such as RAOB, they were not available over the Alpine-Needham path, along which only surface observations were at hand. The Weather Bureau stations at New York, Hartford, and Boston gave daily surface readings of temperature, barometer, and dew point, from the mean values of which the upper curve of Fig. 2 has been constructed by the formula.

$$(n - 1) \times 10^6 = 79/T(p + 4800\, e/T)$$

where

n = refractive index
p = pressure in millibars
e = vapor pressure in millibars
T = temperature in °K.

A comparison of the two curves of this figure shows good agreement as a seasonal matter, and it will also be seen that some of the finer detail appears in both refraction and field. It may be assumed that the differences found are largely due to the imperfect character of surface refraction as an index of refractive lapse rate, and that if radio-sounding balloon observations for this path had been available a closer detail correspondence would have been found.

In Fig. 3 daily values of surface refraction are given for July, 1945, and January, 1946. Here, in addition to the general difference in level between winter and summer months, the magnitude of the day-by-day change in refraction is shown. It will be seen that these changes are relatively large, so that the lowest refraction in July fell below the highest value of January.

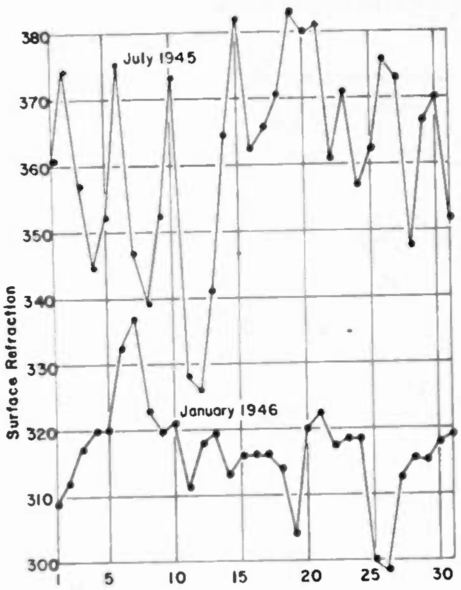

Fig. 3—Calculated surface refraction of the atmosphere for Alpine-Needham path for July, 1945, and for January, 1946.

For the entire period investigated, February, 1945, to July, 1946, it was found that the index ranged from a low of 1.000296 on March 3, 1945, to a high of 1.000388 on September 15, 1945. If it is assumed[1] that there is a uniform lapse rate from the surface value to unity at a height of 10 km., then

$$r/(r+10) = 1/n$$

where

r = radius of curvature in kilometers
n = surface refraction
10 = height of atmosphere for a continuous density gradient,

and from this formula the radii of curvature for the minimum and maximum surface-refraction values given above are 33,800 and 25,800 km., respectively, or 5.31 and 4.05 times the earth's radius.

But, as is well known from sounding-balloon observations, the lapse rate is not at all uniform in the troposphere as a whole, the greater part of the refractive bending taking place in the first two or three kilometers above the earth's surface. The effect of this is naturally to increase the amount of wave bending, or, what is equivalent, to decrease the radius of curvature from that computed on the assumption of a uniform lapse rate. This may be illustrated by RAOB soundings at Albany, N. Y., which were first transformed into refractive index at all levels in an air column some 10 km. high. Computing the amount of bending at various altitudes in this column, it was found that this was greatest in the first 3 km. above the surface; so an air column 3 km. high was taken, and separated into two sections, a top section from 2 to 3 km., and a bottom section from 0 to 1 km. Mean values of refraction for these sections were computed, and the radius of curvature determined for two months, July, 1945, and January, 1946, which are plotted in Fig. 4. It will be seen from

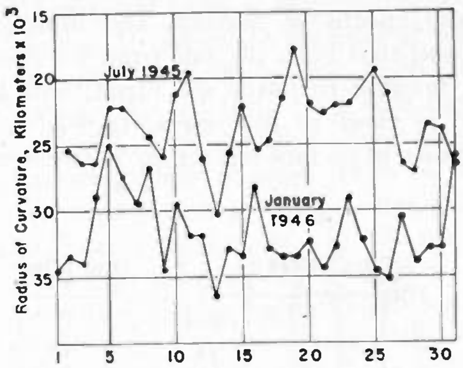

Fig. 4—Calculated radius of curvature of bending by atmosphere from RAOB observations above Albany, N. Y., for July, 1945, and for January, 1946.

this figure that the highest and lowest values for bending are 17,700 and 36,500 km., corresponding to 2.8 and 5.7 times the earth's radius, respectively, or a range of over 2 to 1. We have seen above that the high and low values for bending for these two months, computed on the basis of a uniform lapse rate, are 25,800 and 33,800 km., or a ratio of only 1 to 1.31.

Although Albany is only some hundred miles northwest of the center of the Alpine-Needham path, comparison of Figs. 3 and 4 shows there is only a general resemblance between the two pairs of monthly curves. This is, of course, to be expected for, in any day-to-day comparison, points 100 miles apart would often have both different surface conditions, as well as different lapse rates.

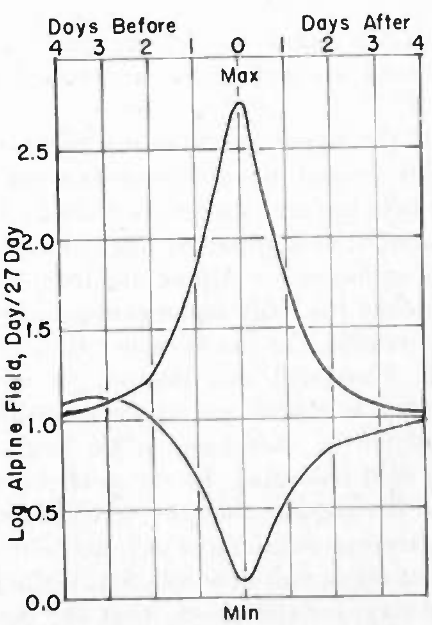

Fig. 5—W2XMN field intensity grouped about epoch of maxima and minima fields in log microvolts. Maxima = 2.0; minima = 0.2.

[1] W. J. Humphreys, "Physics of the Air," McGraw-Hill Publishing Co., New York, N. Y.; 3rd ed., 1940; pp. 468–469.

As a preliminary study of field intensities around days of maximum and minimum fields, the quantities involved were first separately examined to find if their distribution around maxima and minima was favorable for day-to-day correlation. In Fig. 5 is given the distribution of Alpine log fields around maxima of $\geqq 2.0$ and minima $\leqq 0.2$. It will be seen that the field distribution around epochs of maxima and minima is symmetrical, and that both the fall from maxima and the rise from minima to unity are rapid, and are nearly completed in three or four days. In Fig. 6 the same study is shown of surface refraction, with essentially the same result. In other words, the daily maxima and minima chosen are well above and below surrounding values.

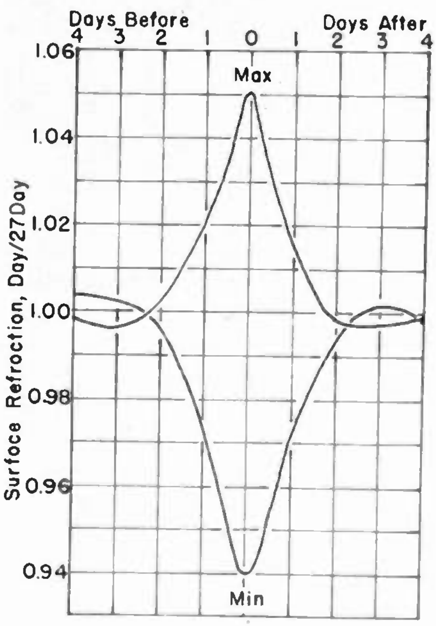

Fig. 6—Change of atmospheric surface refraction grouped around epochs of maximum and minimum fields.

In Fig. 7 the upper curve shows the distribution of Alpine fields around maxima of surface refraction. The maximum field apparently occurs one day before maximum refraction; this apparent displacement is because the recording hours for Alpine are from 1600 to 2300 E.S.T., so that the daily mean centers on 1930, while the surface readings of the Weather Bureau stations at New York, Hartford, and Boston, on which the refractive index is based, are taken at 0130. The next day's refraction is, therefore, only 6 hours away from the day of field recording. In the lower curve of Fig. 7 the process is reversed, and the distribution of surface refraction around maxima of Alpine field is given. A maximum of refraction now follows maxima of field by less than a day, for the reason that the morning readings of refraction are 18 hours from the field readings of the same day.

As a check on the relation shown by Fig. 7, the process was reversed, and minima of refraction and field were taken as epochs. Fig. 8 shows, in the upper curve, the distribution of fields around refraction minima, and in

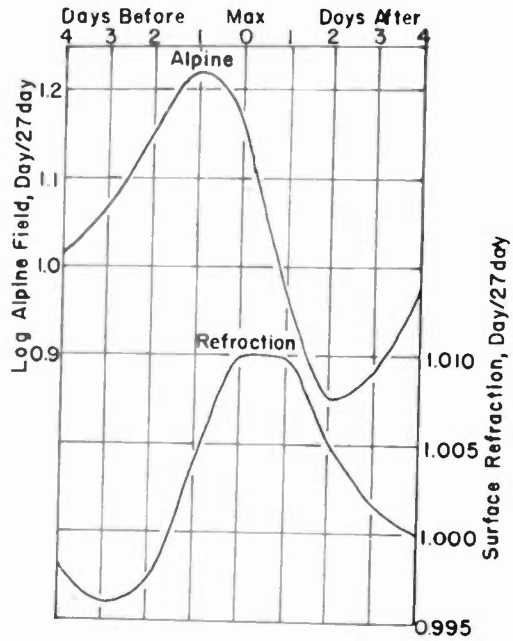

Fig. 7—Upper curve: W2XMN field intensities around epochs of greatest refraction. Lower curve: atmospheric surface refraction around epochs of maximum fields of W2XMN.

the lower, refraction around field minima. While the upper curve, as in Fig. 7, shows a displacement of field to one day before, refraction in the lower curve shows a minimum on the day of the field minimum. But the values used are not smoothed; if they were, the lower curve would show the same half-day displacement to the right as we find in Fig. 7, and for the same reason.

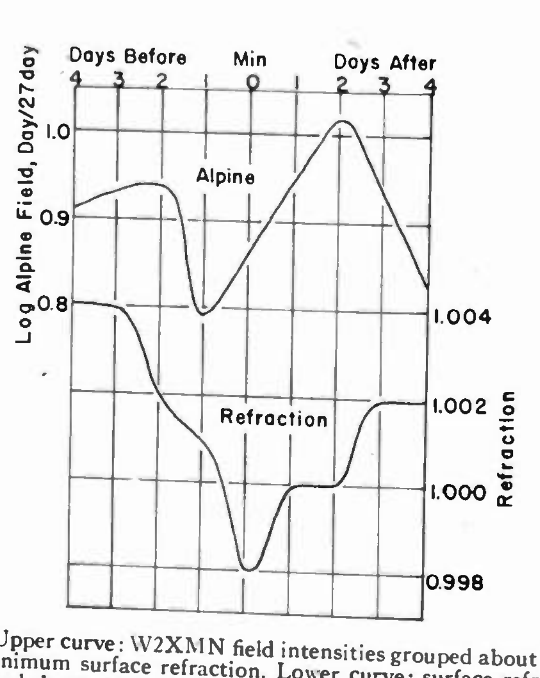

Fig. 8—Upper curve: W2XMN field intensities grouped about epoch of minimum surface refraction. Lower curve: surface refraction grouped about epoch of minimum W2XMN reception.

Variation of Field with Passage of Fronts

Another frequent tropospheric event is the passage of a cold front. In Fig. 9, in which the frontal passage is the epoch, curve A is the resultant effect upon Alpine field of the passage of seventy cold fronts, in the period February, 1945, to July, 1946, showing a lowering of field to 79 per cent of normal. Inasmuch as a front

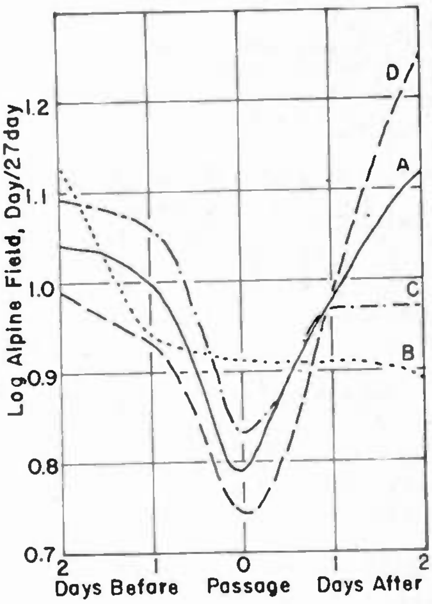

Fig. 9—(A) Alpine fields in log microvolts versus passage of cold fronts. (B) Alpine fields versus passage of cold fronts parallel to transmission path. (C) Alpine fields versus passage of cold fronts making angle with transmission path between 0 and 30 degrees. (D) Alpine fields versus passage of cold fronts making angle with transmission path greater than 30 degrees.

which is parallel with the path can improve transmission by forming a wave guide, such fronts have been separated out in curve B, where it will be seen that the passage of such parallel fronts depresses the field to 91 per cent of normal. In curve C is shown the effect of frontal passages making an angle of over 0 degrees with the path, but less than 30 degrees; here the field is

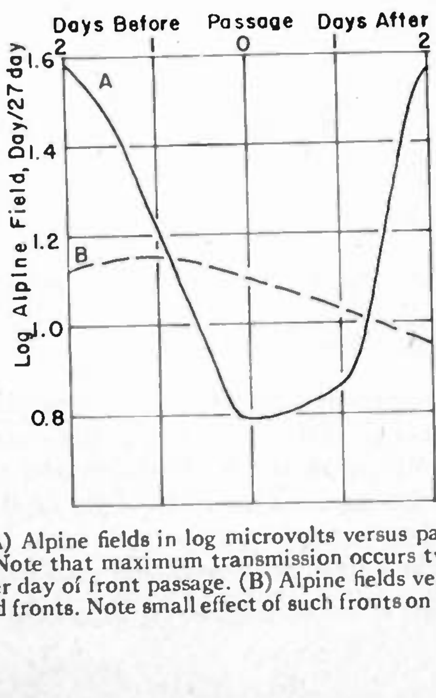

Fig. 10—(A) Alpine fields in log microvolts versus passage of warm fronts. Note that maximum transmission occurs two days before and after day of front passage. (B) Alpine fields versus passage of occluded fronts. Note small effect of such fronts on field strengths.

lowered to 84 per cent of normal. Finally, in curve D, which includes only fronts making an angle of over 30 degrees with the path, we find the greatest depression, to 74.5 per cent. It will be seen from this figure that, while all frontal passages lower reception, the amount of the depression is proportional to the angle made by the front with the path.

There are two other types of surface fronts, the warm and the occluded. Although the number of warm fronts definitely crossing the transmission path during the period February, 1945, to July, 1946, was limited to 8, curve A in Fig. 10 shows a well-marked depression in field accompanying their passage. It is of interest to note that warm-front passages occur principally during periods of high transmission, as will be seen from the values reached two days before and after the front passage. Curve B, which is for occluded-front passage, shows that this type of front has relatively small effect upon Alpine fields.

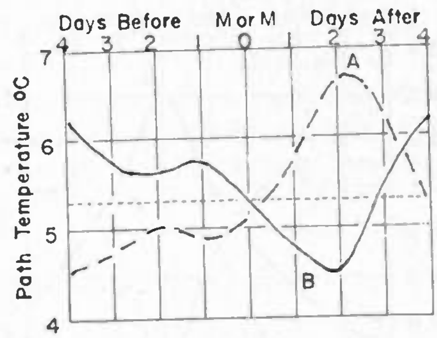

Fig. 11—Correlation of temperature in degrees centigrade with Alpine fields before and after days of maximum reception (A), and of minimum reception (B).

It has been observed, from the very start of Alpine recording at Needham, that abnormally high fields are followed the next day by a rise in temperature. Fig. 11, which takes field maxima and minima as epochs, shows in curve A the relation of field maxima to temperature, the temperature reaching its highest value over two

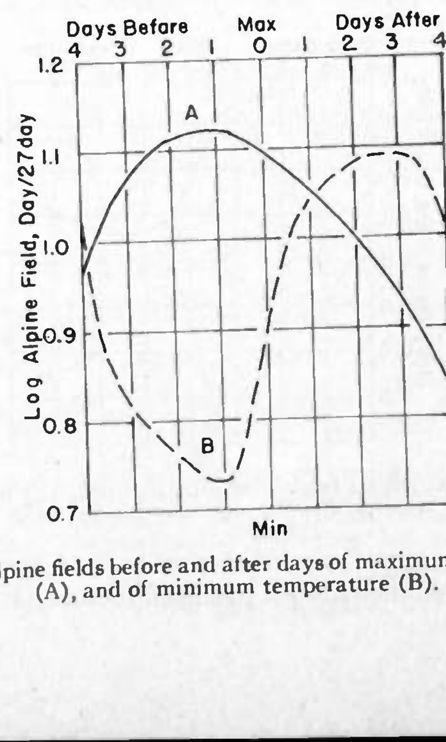

Fig. 12—Alpine fields before and after days of maximum temperature (A), and of minimum temperature (B).

days after the field maximum. Actually, because of the different hours of recording fields and temperatures, the highest temperature is reached in something over 30 hours from the highest field. Similarly, curve B shows that low fields are followed by low temperatures.

As a check on the showing of Fig. 11, the process was reversed, and the distribution of fields around temperature maxima and minima was investigated, with the result shown in Fig. 12. Here, as shown by curves A and B, field maxima and minima occur before temperature maxima and minima, which is exactly the showing of the previous figure.

Variations with Wind Velocity

A well-defined relation between wind velocity on the transmission path and field has been found, and is shown in Fig. 13. In curve A the wind force, in Beaufort units, accompanying field minima shows a maximum one day after (actually, only 6 hours later), while

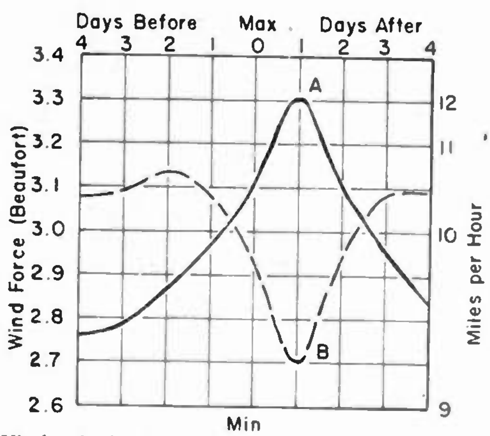

Fig. 13—Wind velocity before and after days of maximum (A) and minimum (B) Alpine fields.

curve B shows a similar relation between maximum field and wind velocity. This is checked in Fig. 14 by reversing the process, and now, of course, the field maxima and minima appear one day (actually, 18 hours), before the wind minima and maxima.

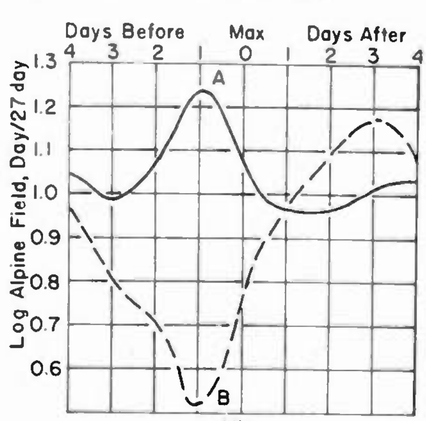

Fig. 14—Alpine fields before and after days of (A) maximum wind velocity, and (B) minimum wind velocity.

The relation of wind direction to field has been preliminarily studied, and it has been found that air movement parallel with the transmission path is less of a disturbing element than winds moving at an angle with the path. In curve A, Fig. 15, wind parallel with the path is taken as the epoch, and a maximum field is found one day before; actually, 6 hours before. Similarly, taking maximum field as the epoch, curve B shows a minimum wind angle with the path one day after, which is in agreement with curve A.

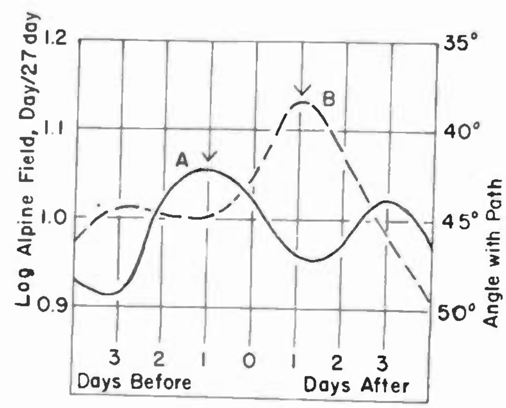

Fig. 15—Alpine fields versus angle of wind direction with respect to transmission path. (A) epoch (0) is for wind parallel with path; (B) shows angle of wind direction with path before and after day of minimum field.

Finally, 2555 hourly averages of Alpine field, covering the period February, 1945, to January, 1946, inclusive, have been analyzed for per cent of time field distribution, with the result shown in Fig. 16. This graph is plotted on arithmetic probability paper, and shows a nearly pure Gaussian distribution; the dotted straight line representing the normal law of error is closely approximated by the full-line curve derived from the data.

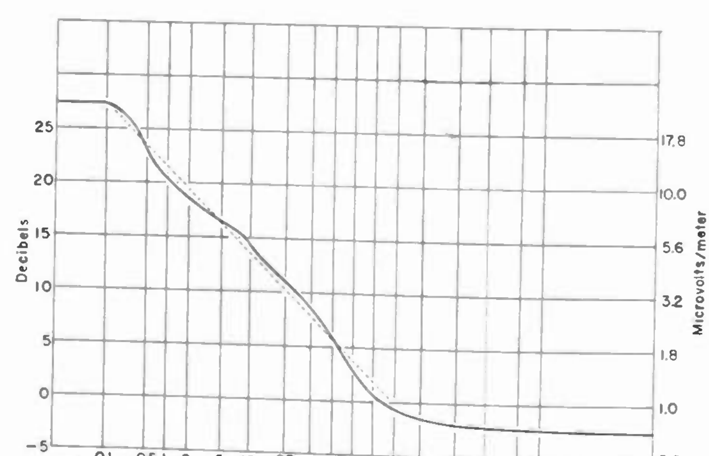

Fig. 16—Percentage distribution of Alpine fields, February, 1945, to January, 1946; 2555 hours.

The field exceeded 50 per cent of the time, or the median value, is 7.5 db above 1 microvolt per meter, or 2.4 microvolts per meter, while the 90 and 10 per cent exceeded fields are -2 and $+13$ db, or 0.8 and 4.8 microvolts per meter, respectively. It must be borne in mind that the unit of measured field used here is a one-hour average.

Measurement of Aircraft-Antenna Patterns Using Models*

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Summary—**Methods of measuring the patterns of airborne antennas using models have been investigated. The conditions which have to be satisfied in a model for accurate simulation are well known. However, in a practical model it is generally impossible to satisfy these conditions exactly, so it is necessary to consider the approximations which are permissible.**

Methods for measuring directly the patterns of transmitting and receiving antennas are described. For low frequencies it has been found advantageous to operate in a vertical direction instead of horizontal when making such measurements, in order to control ground reflections. The equipment which has been used for measuring patterns over a wide frequency range is discussed.

A new method for measuring the patterns of antennas which makes use of the energy reradiated from a receiving antenna when excited by a plane wave has been developed. The reradiated field is distinguished from the exciting field by its modulation, which results from varying the impedance of the receiver periodically. The method has been found to be useful in determining the right- or left-handedness of elliptically polarized fields.

The accuracy of model-antenna-pattern measurements is discussed. Short radial antennas mounted on cylinders have been found to be very useful in evaluating the accuracy of measurements, since their patterns can be calculated.

Models have been used for measuring the patterns of a wide variety of antennas, including simple arrays. Propeller modulation of patterns can be studied with models.

Introduction

THE GREAT increase in number of aeronautical uses of radio in recent years, and the increase in frequency due to the advancement of the art, have created a pressing need for more accurate design information on aircraft antennas. Prior to World War II most aircraft radio installations employed relatively low frequencies and simple antennas. Since the performance of these antennas was comparatively satisfactory, little attention was given to determining the factors which influence the radiation patterns. A few measurements of aircraft-antenna patterns had been made, and these showed the existence of a small number of lobes.

Much of the data needed to solve aircraft-antenna design problems can be obtained from full-scale measurements on the actual aircraft, either in flight or on the ground. The immense amount of time, equipment, and personnel required to perform even the simplest pattern measurement in a flight test makes it impractical thus to attack the problem. Since aircraft for test use are often not available, it becomes important to devise other methods for investigating the patterns of airborne antennas.

Models have long been used to study properties of antennas.[1–11] Methods for measuring the patterns of aircraft antennas using models have been described by Haller.[11] Most of the measurements made on antenna models prior to World War II were limited to an upper frequency (in the model) of about 500 Mc. due to lack of suitable oscillators for higher frequencies. The oscillators used were battery-operated units small enough for installation in the model.

Modeling an Electromagnetic System

Model measurements in electromagnetic systems are based on the principle of electrodynamic similitude—a direct consequence of the linearity of Maxwell's equations.[12–14] Consider an electromagnetic system M (model system) which is derived from another system F (full-scale or prototype system) by dividing all dimen-

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sions of F by a constant factor n. Then it may be shown that systems M and F have geometrically similar fields, provided that the parameters which are characteristic of the media comprising the two systems are related as follows:

$$\epsilon_M \mu_M f_M{}^2 = n^2 \epsilon_F \mu_F f_F{}^2 \tag{1}$$

$$\sigma_M \mu_M f_M = n^2 \sigma_F \mu_F f_F, \tag{2}$$

where

ϵ = dielectric constant
σ = conductivity
μ = permeability
f = frequency
n = an arbitrary constant which determines the size of the model.

The subscript M refers to the parameters of the model, and F to the parameters of the full-scale system. It is assumed that all media are linear, and that a consistent system of units is used.

The quantities ϵ_M, σ_M, μ_M, and n for the model may be chosen at will, provided only that (1) and (2) are satisfied and the media are linear. However, there is only one choice of values of practical interest in antenna models. It is most convenient to make the model measurements in air and, since the air forms part of the prototype as well as of the model, it is necessary to choose $\epsilon_M = \epsilon_F$ (neglecting the possibility of a small conductivity in the air). It is also necessary to choose $\mu_M = \mu_F$ so that it follows at once from (1) that $f_M = nf_F$, and from (2) that $\sigma_M = n\sigma_F$. The conditions to be satisfied for a model in air are summarized in Table I.

TABLE I

Quality	Full-Scale System	Model System
Length	L_F	$L_M = L_F/n$
Frequency	f_F	$f_M = nf_F$
Dielectric Constant	ϵ_F	$\epsilon_M = \epsilon_F$
Conductivity	σ_F	$\sigma_M = n\sigma_F$
Permeability	μ_F	$\mu_M = \mu_F$

Simulation of Dielectrics

For the accurate simulation of an insulating material the requirements in Table I must be satisfied for both dielectric constant and conductivity of the material. However, if the insulator is a high-quality dielectric of negligible loss, its conductivity usually can be neglected in designing the model. Since the only remaining condition on the material is that it have the same dielectric constant as the corresponding material in the prototype, the insulation for the model can be of the same material as in the prototype.

It is sometimes necessary to model the conductivity of a lossy dielectric. Some aircraft are built of plywood with appreciable conductivity, so the error made in neglecting this factor must be considered. At low frequencies the plywood is thin enough to have negligible effect on the propagation of waves, and at very high frequencies the conductivity is unimportant, so that for both these ranges the loss in the plywood can be ignored with little error. However, there is a middle range of frequencies where the conductivity appreciably influences the pattern. The construction of an accurate model for this range is difficult.

Simulation of Metallic Structures

It is impossible to satisfy the requirements on conductivity when good conductors like copper and aluminum are used in the full-scale system and n has a large value. However, the large areas of metal forming airplane surfaces are essentially perfect reflectors for radio waves at all frequencies, so that if a good conductor (copper) be used in the model, the error in simulation will be small.

The effects of inaccurate simulation of metals are most prominent with thin wires. If the metal used to model a wire has too low a conductivity, there is a change in the current distribution, and the pattern will be distorted. Since the high-frequency resistance of a wire is inversely proportional to the square-root of the conductivity, the error is most important with large values of n.

Simulation of Plane Reflectors by Images

It is often desirable to know the pattern of an antenna when mounted on an infinite plane reflecting surface. Use of a flat sheet of finite size will not serve, because there is much distortion of the pattern by the discontinuity at the edges of the sheet. The principle of images is used in computing the patterns of antennas located on or above an infinite plane reflector, which suggests the actual construction of an image to replace a plane reflector. A mirror image of the model and a system to feed the correct currents to the model and its image are used. This method has been very useful in studying the fundamental properties of antennas.

Co-ordinate System used in Measuring the Patterns of Model Antennas

The radiation from most airborne antennas is elliptically polarized, regardless of the antenna employed, so that in measurements of patterns the polarization as well as the magnitude of the radiated signal must be known. The field at a given point in space is resolved into two components in a spherical co-ordinate system with the origin at the antenna. For an elliptically polarized wave these two components at a given point are not in time phase, so that for a complete specification of the field, measurements should be made of this phase angle.

Aircraft may assume arbitrary orientations with respect to the earth, so a co-ordinate system which is fixed to the airplane itself rather than to the earth is desirable. A spherical co-ordinate system is commonly used (see Fig. 1). The field radiated from the aircraft antenna (supposed located at or near the origin) may be resolved into two components E_θ and E_ϕ at any point in space.

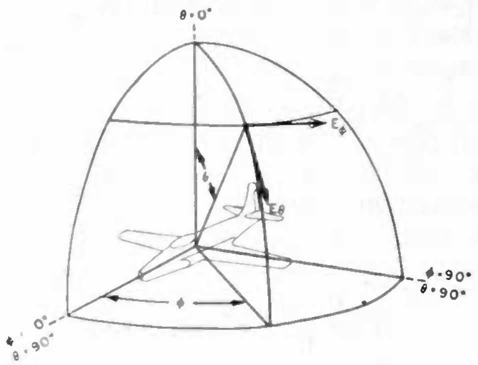

Fig. 1—Spherical co-ordinate system showing orientation of field components relative to aircraft.

The component E_ϕ is always horizontal when the plane is in level flight. The component E_θ is purely vertical only when $\theta=90°$ (that is, for horizontal directions from the antenna). At the zenith $\theta=0°$ the component E_θ is entirely horizontal. Therefore, the use of the terms "horizontally" and "vertically" polarized components as designations cannot be recommended. It should be noted that the angles ϕ and θ may vary over the following ranges:

$$0° \leqq \phi < 360° \tag{3a}$$

$$0° \leqq \theta \leqq 180°. \tag{3b}$$

The reciprocity theorem is of considerable importance in model measurements, since it permits measurements to be made of the pattern under either transmitting or receiving conditions (whatever the function of the full-scale antenna).

Measuring the Pattern as a Transmitting Antenna

In this method a transmitter is used to excite the model antenna and the radiated field is explored with a receiver whose antenna is oriented to measure the desired field component. The method is very convenient when the transmitter can be battery-operated and contained within the model airplane. The receiving equipment can be located at the observing position with no need to relay information from the model.

Avoiding distortion of the test-antenna pattern by stray reflections from the ground and near-by objects is a problem. In the past, a solution has been obtained by locating model and measuring equipment on a high platform (to avoid ground reflections).

A better solution is to operate in a vertical direction. The earth forms a reflector for the receiving antenna located directly under the model antenna. The model is supported on a tall wooden pole in an area clear of objects which might produce stray reflections.

In Fig. 2 a model is shown in position for measurements. The model is held on a wooden frame carried on a cart on wooden tracks attached to the pole. The model transmitter is a self-excited oscillator with tone modulation from an audio oscillator, all battery-operated. The receiver on the ground is conventional, with its antenna arranged to measure the desired field component. A horizontal dipole or a shielded vertical loop may be used. The height of the model above ground is chosen on the basis that the field produced by the antenna on the ground, if transmitting, would be essentially a plane wave in the region occupied by the model.

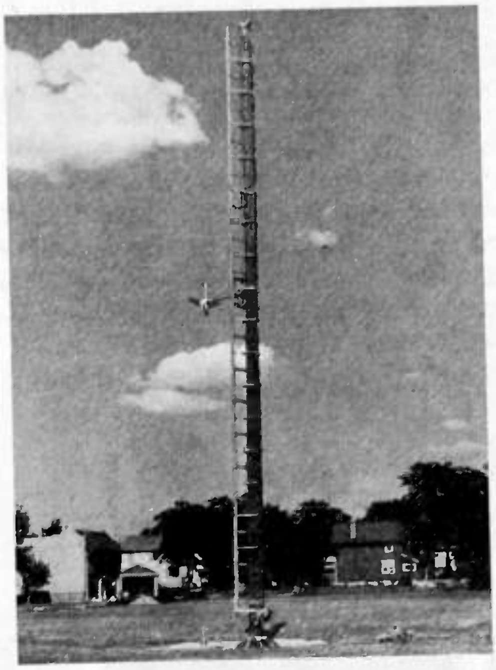

Fig. 2—The vertical pole used to support models for measurements of antenna patterns at low frequencies.

Measuring the Pattern as a Receiving Antenna

The pattern of an antenna when receiving plane waves may be determined by connecting a suitable receiver to it and measuring the receiver output as the direction and polarization of the incident wave are varied. The simplicity and compactness of receivers constructed for high frequencies make this method very convenient for investigating the patterns of aircraft-antenna models. The receiver usually consists of a tuned detector, adequately shielded, no external power supply being needed. Since the transmitter is not inside the model, restrictions on its physical size and power-supply requirements are removed.

The receivers generally used have consisted either of a bolometer or crystal detector with a single tuner. The incident field is tone-modulated, so that the detector output is an audio frequency.

A schematic diagram of typical equipment used at ultra-high frequencies is shown in Fig. 3, and the outdoor installation in Fig. 4.

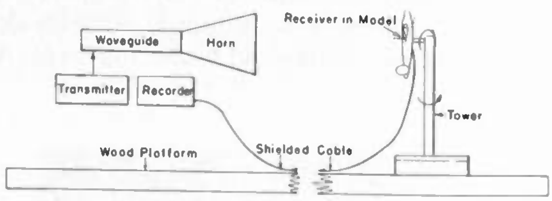

Fig. 3—Diagram showing equipment used for measuring model antenna patterns at high frequencies.

The main difficulty with the method is that of obtaining a remote indication of the output of the receiver. For the frequency ranges where this method is

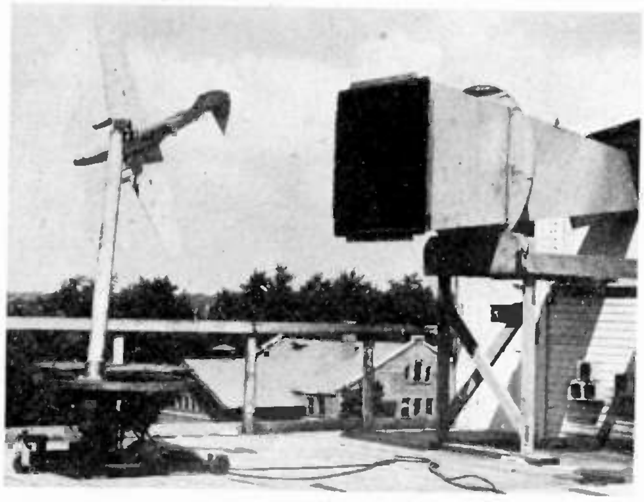

Fig. 4—The electromagnetic horn and the model supporting structure. When measuring patterns, the separation between the model and the horn aperture is usually greater than shown.

of most interest (model frequencies 500 to 10,000 Mc. approximately) the airplane model often provides enough shielding to permit the use of a cable to connect the receiver and the recording equipment. If the cable is carefully located with respect to the antenna on the model, the pattern distortion it causes can be minimized. The cable comes out of the model at such a point as to make maximum use of the metallic portions of the model as a shield. The line leaves the model in such a direction that the wires are normal to the direction of the incident electric vector. A simple test for pattern distortion is the observation of the receiver output when the wires are moved. Variation of output indicates distortion of the pattern.

Two general types of oscillators have been used in the measurements. For the frequency range 600 to about 3500 Mc. triode oscillators with type-GL-2C40 or -GL-2C43 tubes have been employed. Above 3000 Mc. klystron oscillators have proved to be very satisfactory. The oscillators are tone-modulated, usually with a rectangular wave shape.

A horn type of radiator produces a field which is a linearly polarized and nearly plane wave over the region occupied by the model. The design of the horn is a compromise between two conflicting requirements: high directivity to reduce the effects of stray reflections, and low directivity so that the model can be fairly near the horn and still be in an almost uniform field. The horn can be rotated 90° about its long axis to change polarization. The TE_{01} mode is used in the waveguide to excite the horn.

The Supporting Tower

The design of the model supporting structure is one of the most difficult problems encountered, and its solution now leaves much to be desired. The structure must permit the model to be rotated in any arbitrary orientation with respect to the horn, must have a very low echoing area, and must be stable enough mechanically to support a 20- or 30-pound model under ordinary weather conditions. Some of these requirements can be met to a high degree, but combining all the requirements in one structure is another matter. Structures in which models are supported on threads may be used for some tests, but are not flexible enough in operation for routine measurements.

The type of structure used for most measurements may be seen in Fig. 4. The tower is a plywood tube some three inches in diameter with a small metallic head which has a horizontal shaft to hold the model. This shaft is rotated by gearing driven through a small fiber shaft inside the main post. The fiber shaft is motor driven and has a selsyn generator geared for transmitting the angular position of the model to the observing position. The whole structure rotates about its vertical axis to provide the other degree of freedom required, and another selsyn generator is used here.

The supporting post is slanted to clear the tail structures of the model airplanes, and is offset to place the model near the axis of the incident beam. The metal horizontal table provides a definite reflection surface in place of the random reflecting surfaces below the table. The reflection from the table sometimes interferes with measurements. It may be reduced by an absorbing layer of 377-ohm-per-square conducting cloth one-quarter wavelength above the metal table.

A New Method of Measuring Antenna Patterns

Sometimes a transmission line from the model to the observing position cannot be used without distortion of the field. In order to eliminate the transmission line, a new method has been devised and developed. W. L. Everitt suggested that the energy to feed the model antenna could be obtained from a remote transmitting antenna by electromagnetic coupling. Part of the energy induced in the model antenna reradiates, so that measurements of the reradiated field give the desired pattern. To distinguish the reradiated field from the unmodulated primary field of the transmitting antenna, the former is modulated by connecting a periodically varying impedance to the terminals of the model antenna. The radiation pattern of the model antenna is found by measuring the audio output of a suitably located receiver. A schematic diagram of the equipment used is shown in Fig. 5.

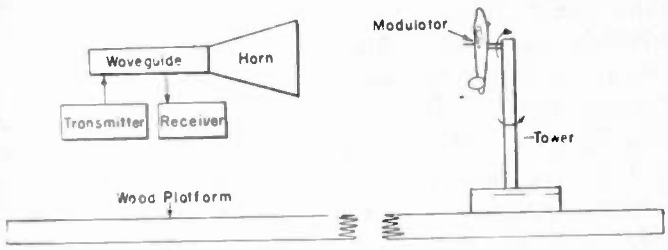

Fig. 5—Diagram of equipment for a new method of measuring antenna patterns.

The current which flows in the model antenna when it is placed in a uniform field is dependent on the pattern of the model antenna for reception. The magnitude of the reradiated field in any direction is fixed both by the antenna current and the directional pattern of the model antenna for transmission. The reradiated field is thus proportional to the product of the two patterns. In general, the desired pattern cannot be found from the data unless the equipment is arranged to make the measured pattern equal to the square of the desired pattern; that is, by making the paths traversed by the transmitted and reflected rays coincident. This is done by mounting the transmitting and receiving antennas in the same wave guide.

Because of the extremely close coupling between the antennas in the wave guide, a large amount of unmodulated signal is applied directly to the receiver. This voltage, rectified by the crystal detector in the receiver, provides bias for the detector and also makes sure that the sidebands of the reradiated signal are rectified linearly. The important components reradiated by the model antenna are the sidebands, since the carrier component is very small compared with the direct field from the transmitting antenna. When the weak modulated signal from the model antenna is combined with the strong unmodulated carrier received directly from the transmitting antenna, the resultant signal has a carrier which is nearly independent of the carrier of the modulated signal. This resultant signal has a very low percentage of modulation due to the modulation on the carrier from the model. Detectors are essentially linear for such signals, which makes calibration unnecessary when only relative indications of field strength are required.

The carrier and sidebands must be combined in the proper phase. If the sideband vectors are combined with the carrier vector to make the angle between them 90°, phase modulation is obtained instead of amplitude modulation, with nearly zero audio output from a detector. If the phase angle between the vectors is other than 90°, a combination of phase and amplitude modulation results, except when the angle is 0° or 180°; in which case pure amplitude modulation is then obtained. This gives correct operation, and is had by varying the phase angle until the audio output is a maximum. The phase angle is varied by changing the distance to the model. This phasing adjustment is important, and must be made for each recorded point on the pattern.

Methods for Modulating the Reradiated Signal

The modulator used in the model consisted of a battery-driven reed vibrator with contacts which open- and short-circuit the end of a transmission line of suitable length, connected to the model antenna. The wave shape of the resultant modulation is roughly rectangular. By using stub lines in shunt with the main transmission line, the system can be tuned to obtain maximum energy in the sidebands. The reed vibrator was designed to minimize discontinuities in the transmission line, and polystyrene insulation was used throughout. The large size of the contacts in these units limited the amount of modulation which could be produced at high frequencies, the upper frequency limit of operation being about 2000 Mc.

An alternative system of modulation was proposed by S. Bertram in which a nonlinear impedance (such as a crystal rectifier) is used. This method has been tried using a type 1N21 crystal biased with an audio-frequency voltage. This type of modulator may permit extension of the method to much higher frequencies.

The auxiliary equipment employed was essentially the same as that described in the previous method.

Design and Construction of the Model

The model airplanes used are simply scale models of the prototype aircraft. However, it is obviously impossible accurately to model every detail of the airplane, so that some basis must be found for determining the degree of detail required. In general, portions of the aircraft structure which are very small in terms of wavelength have a negligible effect on the pattern. An exception occurs in thin wires which carry appreciable currents; these must be modeled. Larger components must be constructed with an accuracy which depends on the extent to which they carry currents and influence the pattern.

The most satisfactory models for ease in handling and convenience in the measurements have been those formed of sheet-copper. These are made by hard-forming copper over a white pine shape. The copper used is about 0.022 inch thick. A simpler method of making models is to use wood (pattern-maker's white pine) made conducting by spraying the completed model with metallic copper (the so-called metallizing process); this is best done over a zinc undercoat.

Modeling the Antenna System

The problems involved in modeling most types of antenna systems are those of eliminating portions of the antenna system which are unimportant as far as the pattern is concerned. The model should be designed to be as simple as possible and yet preserve the principal characteristics of the current distribution on the radiating portions of the system. The patterns of most antennas are insensitive to the impedances of the devices connected to their terminals. It is usually permissible to ignore the nature of the structure inside the airplane and model only the external portion. Important exceptions occur in the cases of antenna arrays and parasitically excited antennas, since their patterns may depend on the impedances connected to their terminals. If the relative pattern of an antenna may be expected to be independent of the impedances of circuits inside the aircraft, then only the external portions of the antenna need to be modeled; otherwise it is necessary to model the interior portions also. If only the external structure is modeled, it is unnecessary to make a model of the connector at the base of the antenna. The requirements of Table I should be observed in designing external portions of the model antenna system to obtain accurate simulation.

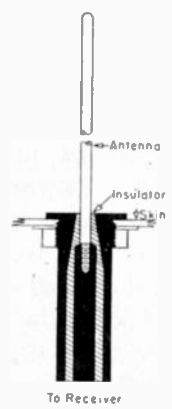

Fig. 6—A cross section of a typical model antenna.

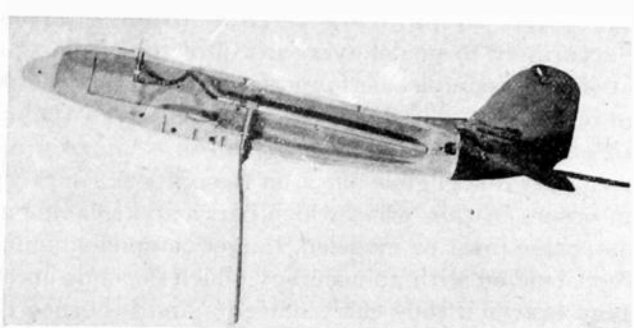

Fig. 7—One-half of a model with antenna and receiver in position. A fiber rod for adjusting the receiver tuning projects from the tail.

Fig. 6 shows a section of a typical model of a whip antenna. The taper section at the base of the antenna should be noted. If the opening in the skin is too large, radiation from the open end of the transmission line may distort the pattern. The opening in the skin is generally made less than 0.125 inch in diameter. Fig. 7 shows an installation in an airplane model.

Antennas with Balanced Feed Systems

In the measurements it is convenient to use coaxial transmission lines wherever possible, since suitable balanced transmission lines, connectors, and tuners are difficult to design. In order to avoid balanced transmission lines, balanced-to-unbalanced transmission-line converter units are required. The best method for performing the conversion has been found to be the use of the so-called "balun"[15] type of circuit. Quarter-wave-type skirt converters have been used to a limited extent.

Another method for feeding a balanced type of antenna from coaxial transmission line is shown in Fig. 8. This method is employed to feed models of loop antennas used in studying the errors of radio-compass loops.

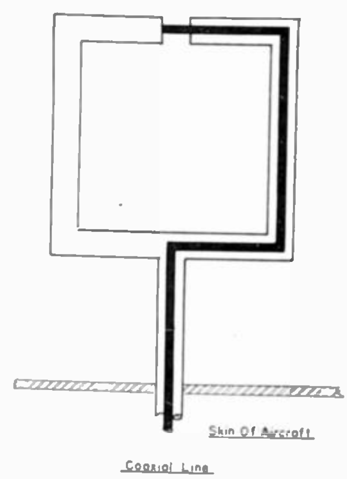

Fig. 8—Method of connecting a balanced loop antenna to a coaxial line.

Modeling Antenna Arrays

The modeling of antenna arrays and other directional systems presents special problems. The patterns of such antennas often depend on the impedances of the circuits attached to their terminals inside the airplane. It is much more difficult to make an accurate model for simulating impedances than for patterns. Two types of two-element antenna arrays in which power is fed to both elements have been successfully modeled with both elements in-phase and with elements out-of-phase while

[15] It is believed that the word "balun" was coined by A. Alford to signify a device suitable for coupling a balanced load to an unbalanced transmission line. This device was originally termed a "bazooka."

carrying equal currents. The method used is shown schematically in Fig. 9. No attempt is made to simulate the impedances of the elements, the only precaution taken being to construct the elements as nearly alike as possible. The actual feed system of the array is replaced with the single-wire-type transmission line, as shown. The length of the line between the elements is chosen so that the distance along the line from the feed point to each element is an odd number of quarter-wavelengths. In quarter-wavelength lines the current through the load is approximately equal to the input voltage divided by the characteristic impedance of the line, and this current is nearly independent of the magnitude of the load impedance. For arrays fed in-phase with equal currents, the feed point is at the center of the line. For arrays fed with equal currents out-of-phase, the line feeding one element is made a half-wavelength longer than that feeding the other. This design of feeding system makes it possible to compensate for small errors in the system by merely sliding the feed tap along the line. If additional adjustment is required, the length of one of the antennas may be changed slightly.

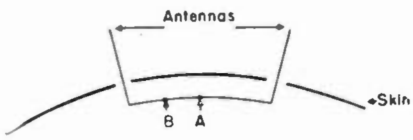

Fig. 9—Method of feeding simple two-element antenna arrays. Feed connection is at *A* for antenna currents in phase, and at *B* for phase opposition.

The accuracy of the phasing obtained by the above method usually can be estimated from the pattern. Arrays on aircraft are usually located in such a position as to have at least one plane in which the pattern is symmetrical. The model array can then be adjusted until a symmetrical pattern is obtained.

Calibration of Equipment

The measurements generally made on model antennas yield only patterns on a relative basis of magnitude. If it is desired to obtain the pattern in absolute terms (for example, field intensity in millivolts per meter at 1 mile for 1 kilowatt radiated), it is necessary to have a means for calibration. Two methods are available for making such calibrations: comparison with an antenna of known performance such as a dipole, or by integration of the Poynting Vector.[10,16] The latter method was employed in the present investigations. This method of calibration yields the field intensity in terms of the power radiated only, since it neglects any internal losses in the antenna. As the internal losses in most aircraft antennas at high frequency are small, the approximation is a good one. Receiving antennas may be calibrated in terms of open-circuit voltage produced per unit of field intensity of an incident plane wave of specified polarization.

Accuracy of Measurements

The accuracy of a specific model measuring apparatus is difficult to determine, since it depends to a large extent on the particular antenna under test. If the model is small in terms of wavelength, it is generally easy to obtain uniformity of field over the model, but it is difficult to prevent distortion of the pattern due to lead wires. On the other hand, if the model is large in terms of wavelength, distortion due to lead wires can be kept to a small amount, but uniformity of field over the region of the model is much more difficult to achieve.

A highly directional antenna is used at the observing position to minimize stray reflections from the ground and from surrounding objects. This means that the model must be placed a considerable distance from the observing position to be in a uniform field. However, the actual separation which can be used is limited by a number of factors, principally the sensitivity of the equipment and stray reflections.

The amount of inaccuracy which nonuniformity of the beam causes in measurements on a specific antenna is difficult to evaluate. Measurements can readily be made of the variation of field intensity over a given region. The interpretation of this variation in terms of errors in measuring a given antenna pattern is extremely difficult. The most satisfactory evaluation, so far, comes from measurements on antennas with known patterns.

The simplest antenna to construct whose pattern is known is a dipole in free space. If such an antenna is placed in the field at any point, properly oriented, and rotated about its own center, the well-known figure 8 pattern is obtained. However, such a procedure merely probes the field in a small region, and an excellent approximation to the theoretical pattern is usually obtained even when the field is known to be quite nonuniform. If the dipole is made to traverse a circle a few wavelengths in diameter, some indication of the nonuniformity of the field is obtained. However, the interpretation of such data (in terms of distortion of the pattern of an aircraft-antenna pattern) is difficult because of the unknown effect of the reflection and shielding from aircraft surfaces.

The test antenna should include metallic surfaces and preferably should approximate the situation with aircraft antennas. At first thought an antenna mounted in the center of a large disk would seem suitable, on the assumption that the pattern is essentially that of an antenna on an infinite plane. Measurements show, however, that a finite disk is a very poor simulation of an infinite plane. The discontinuity represented by the edge of the disk is by no means negligible, and standing waves are set up on the disk to produce considerable distortion

[16] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Co., New York, N. Y., 1943; pp. 782–784.

of the pattern. Increasing the size of the disk merely increases the number of spurious lobes in the pattern, without appreciably decreasing their intensity. Attempts to suppress the standing waves by terminating the edges of a disk have been partially successful. However, the diffraction of the waves over the edge of the disk remains to produce considerable differences between the measured and calculated patterns.

A method for calculating the patterns of dipole and loop antennas mounted on or near infinitely long cylinders has recently been devised.[17] The patterns of an antenna mounted on a cylinder which is finite in length will not be greatly different from those calculated for an infinitely long cylinder, especially if the antenna is located midway between the ends of a long finite cylinder and only the patterns in the plane through the antenna perpendicular to the axis of the cylinder are used.

Fig. 10[18] shows a comparison of the measured curve and the calculated points for the pattern of a short radial dipole antenna projecting from the surface of a cylinder one-half wavelength in diameter. If the axis of the cylinder coincides with the polar axis ($\theta=0°$ in Fig. 1) with the antenna at the origin, the patterns are for the E_ϕ component in the plane $\theta=90°$.

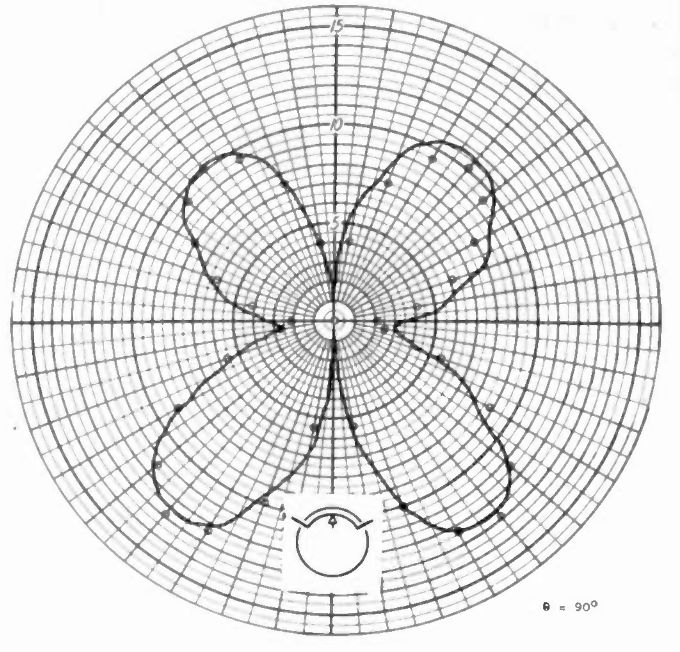

Fig. 11—Patterns of array of two radial antennas on cylinder one-half wavelength in diameter. The antennas were mounted 120° apart on the circumference and fed in phase. Points computed. Curve measured at 1000 Mc.

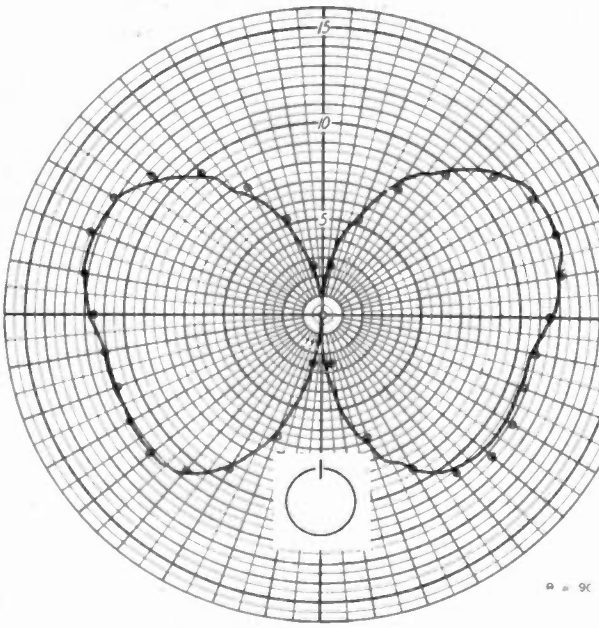

Fig. 10—Pattern of short radial antenna on a cylinder one-half wavelength in diameter. Points computed. Curve measured at 1000 Mc.

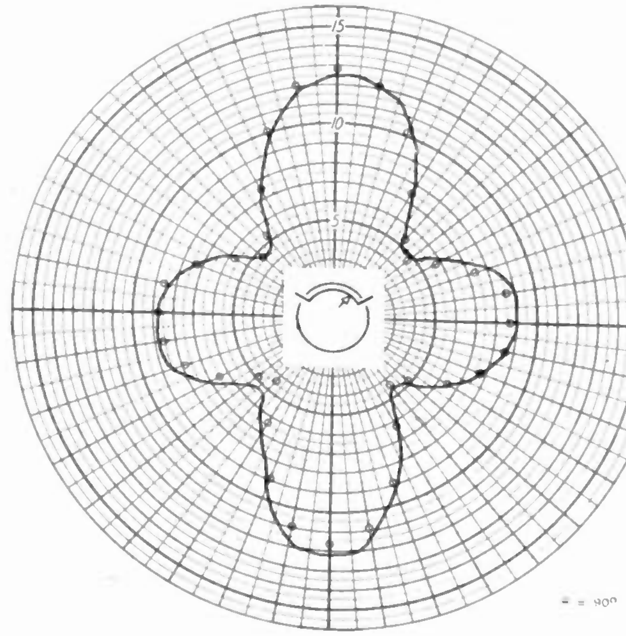

Fig. 12—Same as Fig. 11, but antennas fed in phase opposition.

Figs. 11 and 12 illustrate the agreement which is obtained between measurements and calculations for the patterns of simple antenna arrays mounted on cylinders. The arrays consisted of short radial dipole antennas mounted 120° apart on a circumferential circle of a cylinder one-half wavelength in diameter. The E_ϕ component in the plane $\theta=90°$ was measured.

Approximate calculations of the patterns in certain planes for slot antennas mounted on cylinders can be made. The method of calculation and the assumptions involved are described in the Appendix. Fig. 13 shows the measured and calculated patterns for an axial slot antenna in a cylinder 1.25 wavelengths in diameter.

[17] P. S. Carter, "Antenna arrays around cylinders," Proc. I.R.E., vol. 31, pp. 671–693; December, 1943.

[18] The patterns of Figs. 10 to 16 were measured by Robert A. Fouty.

Fig. 14 is the same for a transverse slot in a cylinder 1.25 wavelengths in diameter. This pattern shows that the assumption of sinusoidal distribution of field intensity in the slot is approximately correct.

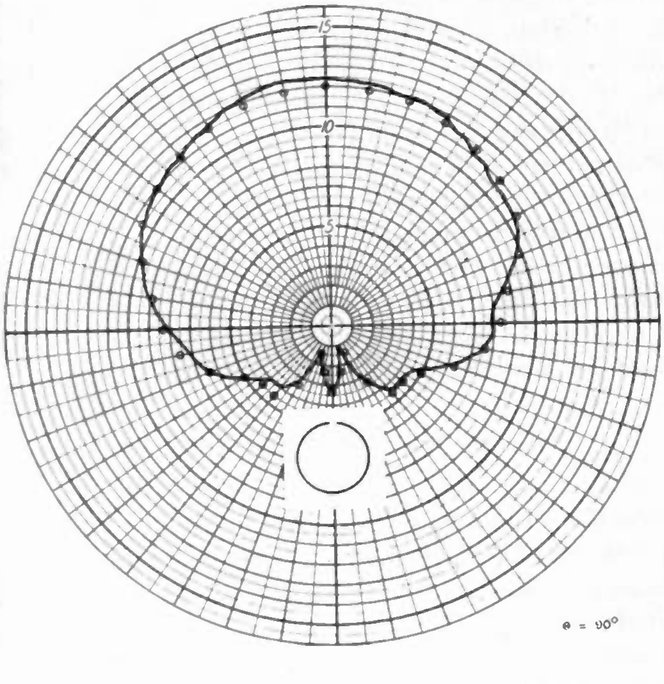

Fig. 13—Pattern of a narrow axial slot three-quarters wavelength long, in a cylinder one and one-quarter wavelength in diameter. Points computed. Curve measured at 3000 Mc.

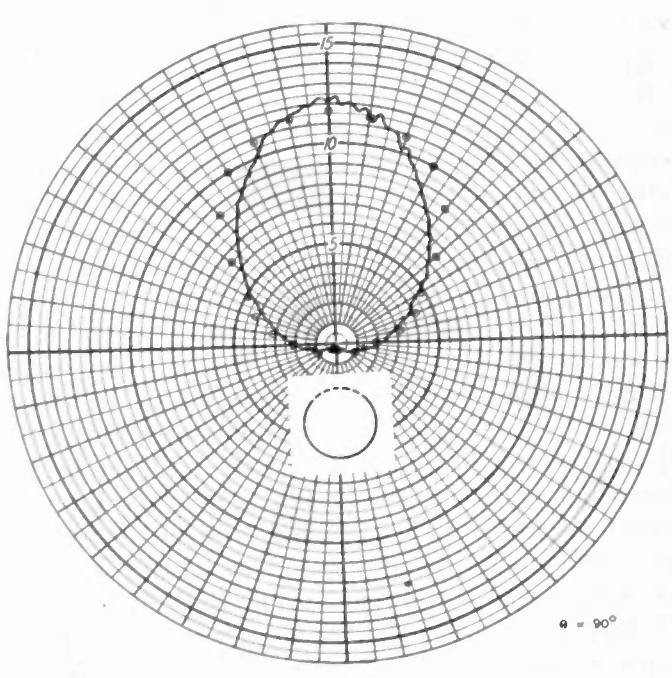

Fig. 14—Pattern of a narrow transverse slot three-quarters wavelength long, in a cylinder one and one-quarter wavelengths in diameter. Points computed. Curve measured at 3000 Mc.

Many similar measurements have been made to test the agreement between measured and calculated patterns. Such a procedure produces confidence in the accuracy of the measurements but does not yield a complete proof of the accuracy of the measurements on aircraft model antennas. The best proof so far has been the measurement of the patterns of a given antenna, using various scales in the modeling. Fig. 15 compares the patterns of a simple quarter-wave antenna mounted on 1/20- and 1/40-scale models of a B-17. The measurements were obtained by choosing locations in the beams from the horns where previous measurements had shown that accurate patterns of antennas on cylinders could be obtained.

It is known that some of the differences in the patterns are due to differences in the models. The remaining discrepancies are due to nonuniformity of the beams, mechanical inaccuracies, etc. Bolometer detectors were used for the measurements.

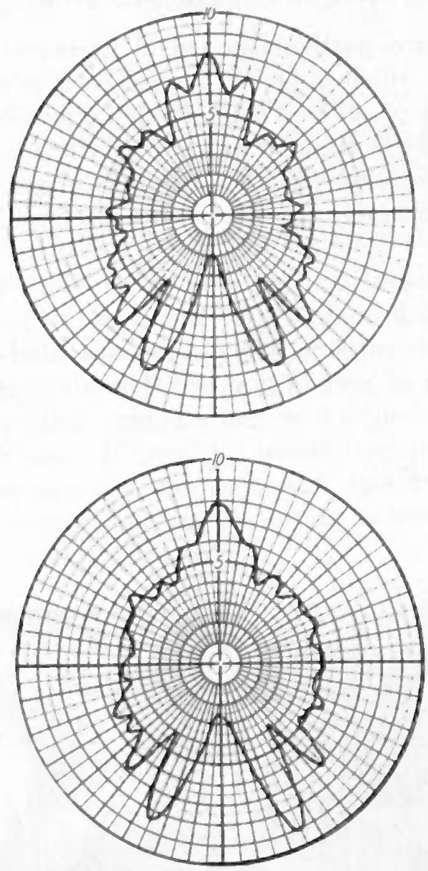

Fig. 15—Patterns of the E_θ component in the horizontal plane ($\theta = 90°$) for a vertical stub antenna mounted in a B-17. Full scale frequency, 150 Mc. Above: 1/20-scale model. Below: 1/40-scale model.

Comparison of patterns of aircraft antennas obtained using models with full-scale measurements have been made by other laboratories. In most cases the agreement is reasonably good. The difficulties in making full-scale pattern measurements make it hard to decide whether the discrepancies which do exist are in the full-scale or in the model measurements.

Miscellaneous Inaccuracies

At the high frequencies used in the model measurements, it is feasible to use antennas having sufficiently high directivity to eliminate most of the effects of platform reflections. Reflections have been noticed when the model frequency has been reduced below 300 Mc. At these frequencies it is better to use a vertically oriented measuring setup (as described earlier) which is not affected by ground reflections.

There are many other possible sources of inaccuracy, such as deviations of detectors from assumed law of operation, nonlinearity in amplifying and recording equipment, sluggishness of recorder, noise and hum, frequency and amplitude instability in oscillators, etc. These sources of error are subject to test and their effects can be definitely evaluated and minimized.

Special Application of Models

Models are particularly useful in investigations of the factors which influence the patterns of aircraft antennas. It is possible to make modifications of the structure of a model airplane which would be impractical in full-scale tests; for example, the empennage of a model may be completely removed to observe its effect on the pattern.

Three-dimensional patterns of antennas are readily obtained with models. Fig. 16 shows the patterns of a vertical stub antenna which were measured on a 1/20-scale model of a B-24 bomber. The antenna was one-eighth wavelength long and was mounted on top of the fuselage, centered above the wing. It should be noted that, even though a vertical antenna was used, a considerable portion of the energy is radiated in the E_ϕ (horizontally polarized) component. This change in polarization is due to currents which flow on the surface of the aircraft.

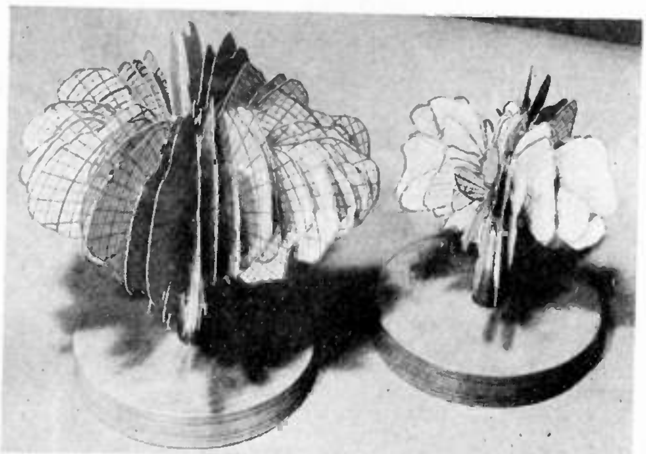

Fig. 16—Three-dimensional patterns of a one-eighth wavelength vertical antenna on top of fuselage of a B-24. Right-hand pattern is the E_θ component and left-hand is E_ϕ component.

Propeller Modulation

In choosing locations for antennas on propeller-driven aircraft, consideration usually must be given to the amount of modulation the rotation of the propeller may introduce into the signal. The propeller blades may have currents induced in them and act as parasitic radiators. The disturbance of the pattern of the antenna varies with the rotation of the propeller to produce a modulation of the signal. Much useful information on the magnitude and wave shape of the modulation of the signal due to propellers may be obtained using models. For a given direction of propagation of the signal, it is merely necessary to observe the variation of signal when the orientation of the propeller is varied. Tests have shown that the percentage of propeller modulation predicted from model tests is in substantial agreement with actual full-scale measurements.

Measurements of Ellipticity of Polarization

The fields radiated from most aircraft antennas have been observed to be elliptically polarized at the higher frequencies. For complete information on the field radiated from a given antenna, it is necessary to measure the orentation of the ellipse of polarization and also the direction of rotation of the electric intensity vector around the ellipse. This information can be found by model measurements.

The major and minor axes of the ellipse of polarization at a given point in the field can be determined by rotating the horn about its long axis to determine the directions and magnitudes of maximum and minimum signals. This may be done using any of the methods of measuring patterns which have been described. The direction of rotation around the ellipse can be determined by observing the changes in the phasing adjustment required in the reflection method for measuring patterns, described above.

The rotation of the polarization is said to be right-handed if the electric vector rotates clockwise when looking toward the source of the field. The reverse direction of rotation is called left-handed. If the transmitting antenna is oriented to measure the major axis of the ellipse with correct phasing adjustment, and then rotated clockwise (when looking toward the model), right-handed polarization will require a decrease in the separation between the model and the transmitter to maintain correct phasing. Left-handed polarization will require an increase in separation.

Conclusions

Models have become a very powerful tool in the design and development of aircraft antennas of all types. Techniques are available for modeling the important pattern characteristics of practically every antenna likely to be employed on aircraft. In studies of the factors which may affect the pattern of an aircraft antenna, model tests provide a much larger amount of data than is obtainable by any other method.

Acknowledgment

It is a pleasure to acknowledge the inspiration and guidance of W. L. Everitt, who directed the research in its early stages. Sidney Bertram and Paul H. Nelson designed the equipment used in the early phases. The co-operation of the personnel of the Special Projects Laboratory of the Aircraft Radio Laboratory, Wright Field, and particularly that of Col. G. L. Haller and Capt. Allen S. Meier, is gratefully acknowledged. Much credit is due our associates of the Antenna Laboratory for their contributions to particular phases of the research.

APPENDIX

Axial Slot Antenna on a Cylinder

It is possible to calculate approximately the pattern of a rectangular slot antenna on a cylinder by assuming a distribution for the field across the slot. The problem is simplified by assuming the cylinder to be infinitely long. In general, the pattern of greatest interest is that in a plane normal to the axis of the cylinder. This pattern is essentially independent of the axial distribution assumed for the slot, so the slot may be infinitely long with no axial variation of the field.

Consider a perfectly conducting cylinder of radius a and of infinite length, with a slot of angular width ϕ_0. The slot is parallel to the axis of the cylinder and infinitely long. Assume that the exciting electric field is uniformly distributed in the slot and polarized such that there is only a circumferential component of electric intensity E_ϕ (in cylindrical co-ordinates). Then the boundary conditions on the surface of the cylinder $\rho=a$ are

$$E_\phi\big|_{\rho=a} = E_0 e^{i\omega t} \qquad -\frac{\phi_0}{2} < \phi < \frac{\phi_0}{2} \tag{4}$$

$$E_\phi\big|_{\rho=a} = 0 \qquad |\phi| > \frac{\phi_0}{2}. \tag{5}$$

This field distribution may be resolved in a Fourier series of the form

$$E_\phi\big|_{\rho=a} = \sum_{n=-\infty}^{\infty} c_n e^{in\phi+i\omega t}. \tag{6}$$

The coefficients C_n are readily found to be

$$c_n = \frac{E_0}{n\pi}\sin\left(\frac{n\phi_0}{2}\right). \tag{7}$$

The field outside the cylinder may be represented by an infinite series of Hankel functions.[19]

[19] See page 525 of footnote reference 14.

$$E_\phi = ik\mu\omega \sum_{n=-\infty}^{\infty} a_n H_n^{(2)\prime}(k_\rho) e^{in\phi+i\omega t} \tag{8}$$

where

$H_n^2(z)=$ Hankel function of the second kind

$$H_n^{(2)\prime}(z) = \frac{d}{dz} H_n^{(2)}(z)$$

$k = 2\pi/\lambda$
$\mu =$ permeability
$\lambda =$ wavelength.

M.k.s. units are used. When $\rho=a$, (8) must be identical with (6). Hence, equating corresponding coefficients, it is found that

$$a_n = \frac{E_0 \sin\left(\frac{n\phi_0}{2}\right)}{ik\mu\omega n\pi H_n^{(2)\prime}(ka)} \tag{9}$$

and the external field is determined.

The pattern is obtained by evaluating (8) at distances from the cylinder which are large compared to the diameter of the cylinder and to the wavelength. Inserting the asymptotic expansions for the Hankel functions and assuming that only a few terms of the series are needed (say $|n| \leqq n_0$) and that

$$\frac{\sin\frac{n\phi_0}{2}}{n} \approx \frac{\phi_0}{2} \quad \text{for} \quad |n| \leqq n_0, \tag{10}$$

there results

$$E_\phi = \frac{A}{2\pi i}\sum_{n=-n_0}^{n_0} \frac{e^{i(n\phi+n\pi/2+\omega t)}}{H_n^{(2)\prime}(ka)} \tag{11}$$

where

$$A = E_0\phi_0\sqrt{\frac{2}{\pi k\rho}}\, e^{-i(k\rho-\pi/4)}. \tag{12}$$

Equation (11) was used to compute the points in Fig. 13.

Transverse Slot Antenna on a Cylinder

The pattern for a rectangular slot antenna whose longest dimension is circumferential to the cylinder cannot be calculated as accurately as that for an axial slot, since the circumferential distribution of the electric field in the slot is known only approximately. An approximation to the pattern may be obtained by assuming a sinusoidal distribution. Since the pattern of most interest is that in a plane normal to the axis of the cylinder, and is not greatly affected by the distribution axially, the slot may be assumed infinitely wide axially.

The field at the surface of the cylinder $\rho = a$ is assumed to be

$$E_z|_{\rho=a} = E_0 \cos\left(\frac{\pi\phi}{\phi_0}\right)e^{i\omega t} \qquad -\frac{\phi_0}{2} \leqq \phi < \frac{\phi_0}{2} \tag{13}$$

$$E_z|_{\rho=a} = 0 \qquad |\phi| > \frac{\phi_0}{2}. \tag{14}$$

This may be resolved in the Fourier series

$$E_z|_{\rho=a} = \sum_{n=-\infty}^{\infty} b_n e^{in\phi+i\omega t} \tag{15}$$

where

$$b_n = \frac{E_0\phi_0 \cos\left(\frac{n\phi_0}{2}\right)}{\pi^2 - n^2\phi_0^2} \qquad |n\phi_0| \neq \pi. \tag{16}$$

Assuming the external field is represented by Hankel functions[17]

$$E_z = \sum_{n=-\infty}^{\infty} d_n H_n^{(2)}(k\rho)e^{in\phi+i\omega t} \tag{17}$$

and comparing coefficients of corresponding terms in (15) and (17) when $\rho = a$, it is found that

$$d_n = \frac{b_n}{H_n^{(2)}(ka)}. \tag{18}$$

Inserting this in (17) and using the asymptotic expansions for the Hankel functions, the field at large distances from the cylinder is obtained

$$E_z = A \sum_{n=-\infty}^{\infty} \frac{\cos\left(\frac{n\phi_0}{2}\right)e^{i(n\phi+n\pi/2+\omega t)}}{(\pi^2 - n^2\phi_0^2)H_n^{(2)}(ka)} \tag{19}$$

where A is given by (12).

Equation (19) was used to compute the points in Fig. 14.

Microwave Antenna Measurements*

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Summary—**A description is given of the techniques involved in measuring the properties of microwave antennas. The measuring methods which are peculiar to these frequencies are discussed, and include the measurement of gain, beam width, minor lobes, wide-angle radiation, mutual coupling between antennas, phase, and polarization. The requirements of the antenna testing site are taken up, and components of a complete measuring system are briefly described.**

I. Introduction

THE RAPID progress in the art of microwave radio during the past few years has produced equally great advances in the development of microwave antennas. At these extremely short wavelengths it becomes feasible to construct compact radiating systems whose dimensions may be very large in comparison to the operating wavelength. Usually these structures are small enough to be placed in a rotatable mount, so that the antenna beam may be steered or pointed over a range of angles. In fact, it is generally more convenient to measure the radiation pattern by rotation of the antenna, instead of by the more conventional method of exploring the surrounding stationary field. These antennas are also distinguished by their relatively high gain and their ability to confine the radiant energy in a sharply defined beam. In many cases these structures are novel in form. Their designs are often based upon the principles of geometric and physical optics, and the associated circuits usually employ wave guides. It is to be expected, therefore, that the study of these newer types of antennas should introduce radically different methods of measurement. It is the purpose of this paper to consider the new problems involved and to describe the measuring techniques employed.

Unless otherwise stated, the techniques described herein are those which were gradually evolved and are at present in use at the Deal and Holmdel Laboratories of the Bell Telephone Laboratories. Many of them originated during the early development of wave-guide techniques and microwaves. Further improvements were made during the extensive radar antenna investigations conducted more recently both at Deal and at Holmdel.

General

In the investigation of antenna characteristics the four main factors of interest are gain, beam width, spurious radiation, and impedance. These factors are frequently interdependent and, in a general way, any one of them can be altered only at the expense of the others. Their relative evaluation is largely dependent upon the purpose for which the antenna is to be used. Since the antenna is a link in an energy transmission system, the antenna gain is nearly always an important factor. In an elementary communication system, for example, the beam width, spurious radiation, and impedance are of interest only as they may affect the gain. However, when the system involves more than one

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† Bell Telephone Laboratories, Inc., New York, N. Y.

transmission path, the spurious radiation (minor lobes and radiation to the rear) becomes important, especially when communication facilities are congested in a given area, and it becomes necessary to avoid interference or cross talk. Furthermore, if the antennas are closely spaced, their mutual coupling becomes important, and if it is necessary to identify or select different transmission paths in nearly the same direction, as occurs in radar target identification and may happen in communication circuits, beam width, symmetry of the beam, and the adjacent minor lobes may be the most important considerations. Thus, in order to be completely cognizant of the properties of a microwave antenna, it is necessary to study the radiation intensity in all directions. There are some applications in which the relative phase and polarization of the radiated wave must also be appraised. In addition to a study of the radiation characteristics, the impedance of the antenna must be controlled and measured, especially when used in conjunction with band-pass circuits, in order that the system perform properly over the desired band. Since the measurement of impedance is, in general, no different than for other wave-guide structures, it will not be considered in this paper.

Generally, an antenna may be classified as either end-fire or broadside, depending on whether the directivity is determined primarily by its length or its area perpendicular to the direction of propagation. Many measurement problems are more easily analyzed in terms of a broadside type of radiator, and in general, only this type will be considered in this paper. The results arrived at will be applicable for the most part to end-fire antennas of directivity comparable to the given broadside structure.

According to the law of reciprocity, equivalent antenna characteristics will be exhibited whether the antenna under test is used as a transmitter or a receiver. Consequently, in the measuring techniques to be described, the antenna under test is sometimes considered as transmitting and at other times as receiving.

II. Measurement of Gain

The gain of an antenna is defined as the ratio of its maximum radiation intensity (power flow per unit area) to the maximum radiation intensity of a standard antenna, both antennas being equally energized. In the past, this standard antenna has been a half-wave dipole, but in microwave measurements it has been replaced by a hypothetical antenna which radiates uniformly in all directions, i.e., an isotropic[1] radiator. When the gain is compared to that of this isotropic radiator, it is defined as the absolute gain of the antenna.

At wavelengths of several meters or more it is necessary to distinguish between two gain definitions, namely, directivity gain, which is a measure of the receiving antenna's ability to discriminate against atmospheric noise by reason of its sharp directional properties, and signal gain, which takes into account heat and other losses in the antenna structure. These losses are important in transmitting antennas as they reduce the power radiated, whereas in the receiver, sensitivity is limited by atmospheric noise, and directivity gain is more important. At microwaves, however, receiver sensitivity is limited by set noise (first-circuit noise) rather than by atmospheric noise, and losses in the antenna impair the sensitivity of a receiver just as seriously as they impair the efficiency of a transmitter. For this reason the concept of directivity gain is less important at microwaves, and in what follows the term "gain" will imply "signal gain."

An antenna of given area exhibits maximum gain when the energy distribution, phase, and polarization are uniform across its aperture; the gain of such a "uniphase, uniamplitude" antenna is

$$\frac{4\pi A}{\lambda^2} \tag{1}$$

where A is the aperture area. This "perfect" antenna is, however, difficult to realize in practice, and to indicate how closely a given antenna approaches perfection the term "effective area" has come into use. This is the percentage of its actual area that an antenna would occupy if it were uniphase, uniamplitude. Thus, an antenna which has an actual area A' and a measured gain $G' = \frac{1}{2}(4\pi A'/\lambda^2)$ has an effective area of 50 per cent.

Two general methods of determining the gain of microwave antennas are possible: absolute-gain measurements, and relative-gain measurements. The former is usually difficult to perform, so that the one more commonly employed is that of measuring the gain relative to some accurately calibrated secondary standard. The absolute gain of this standard antenna, however, must be accurately known, and it is therefore generally determined by the absolute-gain-measuring method described later.

A. Comparison Method

The comparison method of measuring gain involves comparing the signal received by a secondary gain

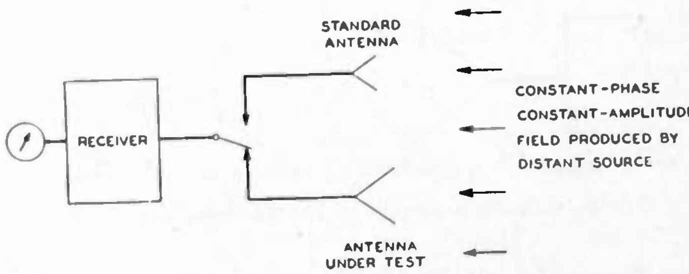

Fig. 1—Gain measurement by the substitution method.

standard to that received by the antenna under test, by substitution (Fig. 1). This standard may be

[1] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., Inc., New York, N. Y., 1943; p. 335.

another antenna which has been accurately calibrated, and is usually a simple radiating structure whose gain can also be calculated. At lower radio frequencies a doublet or a loop antenna is often used for this purpose, but at the higher frequencies, where we are concerned with greater directivity, it is desirable to use a standard of higher gain. The electromagnetic horn[2,3] is commonly used for this purpose because of its basic simplicity, reliability, and desirable broad-band impedance characteristics, and because its gain can be calculated from the physical dimensions.

In making the gain comparison, the safest method is direct substitution whereby the secondary gain standard is physically interchanged with the antenna under measurement. Since it is usually undesirable to disturb the antenna under test, the standard is placed near the antenna and the receiver is switched from one to the other to make the comparison. In this case it is important to be sure that the field strength is identical at the two apertures, or that any differences are accounted for.

In making the comparison, the sensitivity of the receiver must be constant. This requires that the impedance of the antenna and gain standard be accurately matched to the line or that they be adequately isolated by an attenuating pad, so as not to react on the receiver. Furthermore, since any impedance-mismatch loss between the antenna and transmission line will subtract from its gain, maximum gain will be realized only when the mismatch is eliminated. As previously mentioned, one of the virtues of a horn as a substandard is its good impedance match to a wave guide over a broad band of frequencies.

The precision of the comparison method depends upon how accurately the gain of the secondary standard is known. Although the gain of a horn which is used as a standard may be calculated, it is satisfying to be able to check the calculated result by an absolute gain measurement.

B. Absolute-Gain Measurement

In the transmission method of measuring gain, two identical antennas are placed a distance apart r (Fig. 2) and the loss in the transmission path is measured by

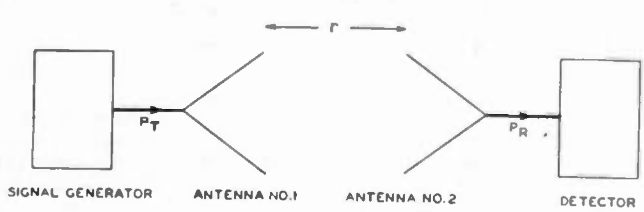

Fig. 2—Gain measurement by the transmission method.

comparing the received power P_R in the setup illustrated with the transmitted power P_T. The gain G of the antennas under test is then determined from the relation[4]

$$\frac{P_R}{P_T} = \left(\frac{G\lambda}{4\pi r}\right)^2 \tag{2}$$

where λ, the wavelength, and r are in like units. The gain as determined by this equation is the absolute power gain and is usually expressed in decibels:

$$G_1 = 10 \log_{10} G \text{ decibels.} \tag{3}$$

For an accurate determination of gain, it is advantageous to take several measurements at different values of r. An example of this type of measurement is shown in Fig. 3. Here the ratio P_R/P_T is plotted versus distance r for two identical antennas. It is seen that, for r large, the curve is a straight line of slope $\frac{1}{2}$, indicating that it is obeying (2) in that P_R/P_T falls off with inverse

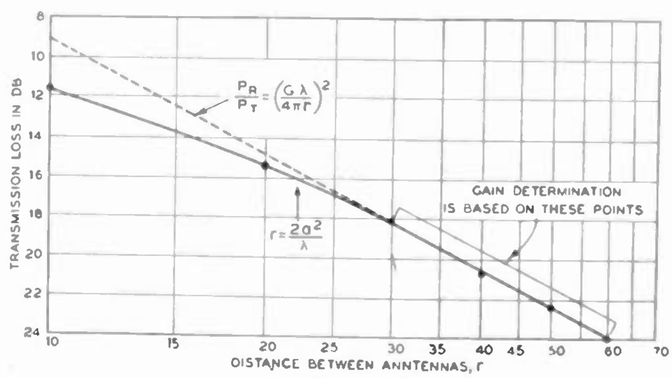

Fig. 3—Absolute-gain determination of two identical optimum horns having square apertures with sides of length a.

distance squared. For r short the curve is no longer a straight line and the gain calculated from points in this range would be in error. This condition arises from the fact that at short distances the antenna does not exhibit the same properties in the way of beam width, gain, and directional pattern that it does at great distances, and this fact imposes a limiting value on the distance between source and receiver in all gain and pattern measurements. This distance limitation will be discussed later.

The gain of an antenna may also be determined by measuring the attenuation (α) of a transmission link (Fig. 2) involving two dissimilar antennas and an intermediate transmission path, and then obtaining the ratio of the gains of the antennas (η) by the comparison method (Fig. 1). The attenuation obtained in the first measurement gives the following relationship:

$$\alpha = \frac{P_R}{P_T} = \left(\frac{\lambda}{4\pi r}\right)^2 G_1G_2. \tag{4}$$

[2] See page 360 of footnote reference 1.
[3] Forthcoming paper by A. P. King.

[4] H. T. Friis, "A note on a simple transmission formula," Proc. I.R.E., vol. 34, pp. 254–256; May, 1946.

The comparison measurement gives

$$\eta = \frac{G_1}{G_2}. \tag{5}$$

Eliminating G_2,

$$G_1 = 4\pi \frac{r}{\lambda} \sqrt{\alpha\eta} \tag{6}$$

and

$$G_2 = 4\pi \frac{r}{\lambda} \sqrt{\alpha/\eta}. \tag{7}$$

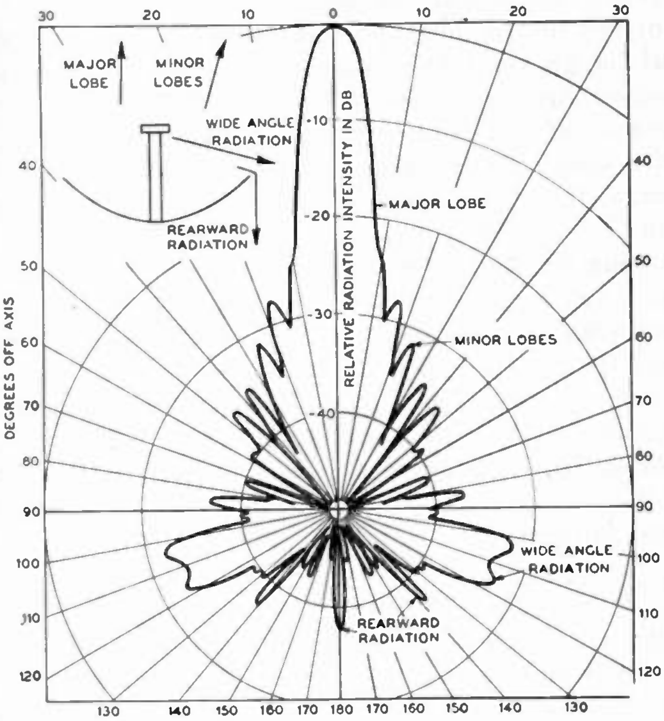

Fig. 4—Representative polar diagram for a microwave antenna.

III. Measurement of Directivity

To determine the directive pattern of an antenna, the radiation in any given direction is compared with the radiation along the axis of the beam. The patterns of antennas at lower frequencies are usually taken by exploring the field of a radiator with a portable detector, but with microwave antennas it is practical to leave the path fixed and measure the pattern by rotating the antenna under test.

A. Directional Properties

The antenna designer usually wants to know the width and shape of the main radiation lobe, the positions and magnitudes of the minor lobes, the wide-angle radiation, and the rearward radiation. All these factors can be shown on a plot of the directional characteristic or pattern of the antenna. With high-gain systems the data is usually plotted on rectangular co-ordinates, in order to spread out the multiplicity of minor lobes. Where a large range of intensities is covered, it is almost essential to use a logarithmic or decibel presentation of intensity. Two ways of presenting the data are shown in Fig. 4 and 5 for a paraboloidal reflector antenna having 28 decibels absolute gain.

The complete analysis of the radiation characteristics ideally would require measurements in all directions, but usually two patterns, one in the plane of the electric vector and one in a plane at right angles (magnetic plane), will suffice. When more data are required, several characteristics may be taken in different planes and either analyzed separately or combined in a contour plot. The contour plot has advantages in studying conditions close to the beam, but with this method it is difficult to cover a wide angular field.

B. Main Radiation Lobe

Often the shape of the main lobe of a directional characteristic is of special interest. When plotted on a logarithmic intensity scale, the curve should be roughly parabolic in shape. The shoulders (vestigial lobes) shown on the major lobe of Fig. 5 are potentially minor lobes, since only a slight change in the antenna will cause them to separate distinctly from the major lobe.

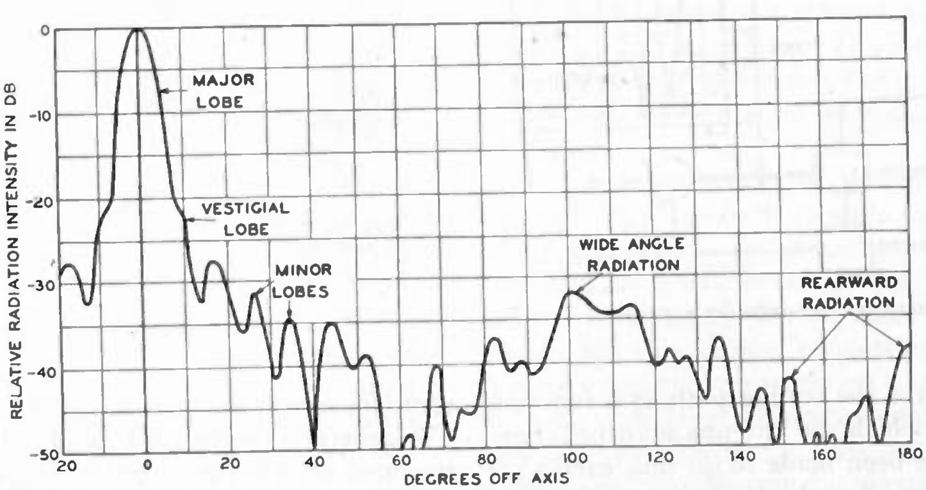

Fig. 5—Directional pattern of Fig. 6, plotted on rectangular co-ordinates.

C. Cross Polarization

Certain types of antennas radiate energy polarized perpendicular to the intended field. This cross polarization may be measured by polarizing the antenna at the distant terminal of the measuring path at 90 degrees to the polarization of the antenna under test. Measurements should be made with especial care in the axial planes at 45 degrees to the intended polarization in which the cross-polarized field is generally maximum. To find the total intensity radiated at any angle, the cross-polarized radiation should be added to the correctly polarized radiation.

When transmitting through the antenna under test, and receiving at the distant terminal, the energy in the two polarizations may be detected simultaneously and added, thus giving a signal which is truly representative of the power radiated per unit solid angle, irrespective of polarization. To do this, a receiving antenna responsive to both polarizations should be used and the components of the field separated in the wave guide according to polarization. The two components should be separately detected in square-law receivers, and the resulting voltages (proportional to the radio-frequency power) added. The total voltage is then proportional to the power intercepted by the antenna.

D. Automatic Pattern Recorder

A tool of extreme usefulness where a number of directional patterns are to be taken is the automatic pattern recorder. This is a device which plots the intensity of

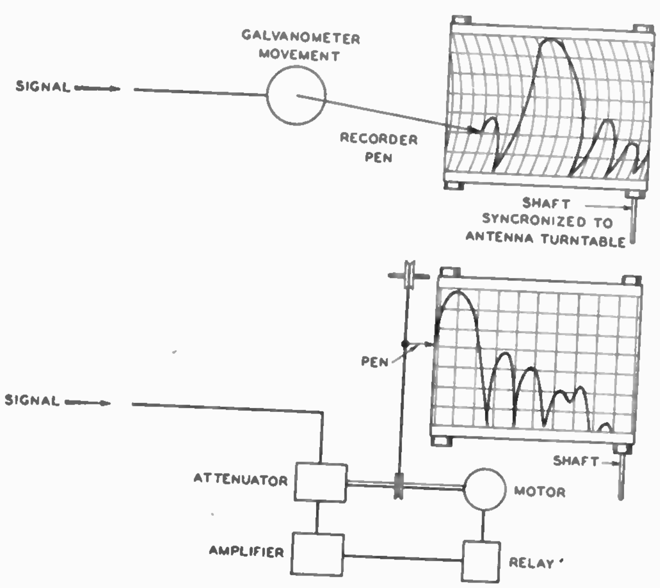

Fig. 6—Automatic pattern-recorder systems.

the signal received over the testing path as a function of the angle through which the antenna is turned. Several mechanisms have been made to do this, each with certain advantages, but all are basically the same. A pen is driven in one direction in accordance with the detected signal intensity, while the paper is moved proportionally to the rotation of the antenna. The pen may be driven from a galvanometer movement, in which case the co-ordinates are slightly curved, or it may be driven on a straight line by a motor which also drives an attenuator and counteracts any change in signal intensity. The two systems are shown in Fig. 6.

IV. Feed Measurements

In appraising a small radiator which is to be used as a feed for a paraboloid or lens, the measurement problems are quite different from those for the complete antenna. Minor lobes are usually of secondary interest provided they do not represent a serious loss of energy, and the gain is seldom measured because there is no obvious direct relationship between the feed gain and the over-all performance of the antenna. The characteristics which are of importance are the distribution of energy, the phase, and the polarization over the main radiation lobe. The energy distribution determines the illumination of the main reflector or lens and the amount of taper to be expected in the final aperture. A knowledge of the phase front of the wave emerging from the feed is important to the design of the feed, and to the correlation of the feed to the reflector or lens. It is also useful in locating the best focal position. The polarization is important, as it can be responsible for undesirable cross-polarization components in the complete antenna.

A. Pattern Measurements

With feeds whose maximum aperture dimension are of the order of a few wavelengths, the path length for the testing site can be short, and it is feasible to make such measurements in the laboratory. A typical laboratory

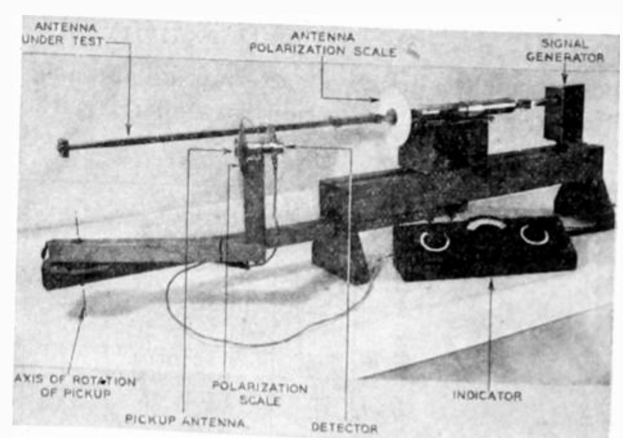

Fig. 7—Primary-pattern-measuring equipment.

setup for measuring primary patterns is shown in Fig. 7. The generator, wave guide, and radiator under test are mounted so that the assembly can be rotated about the axis of the radiator to change the polarization. The receiving antenna is mounted on an arm which is pivoted on an axis through the aperture of the radiator, and it can also be rotated about its own axis. This receiving antenna is made directive so as to discriminate against interference caused by room reflections. The radiation pattern is obtained by measuring the receiver output as a function of its angular position, and may be taken in various planes by rotating the generator and radiator assembly on its axis.

B. Feed Polarization Measurements

The plane of polarization may be ascertained, using the apparatus of Fig. 7, by rotating the detector to obtain a minimum in the received signal. If the field is elliptically polarized, it may be analyzed in components parallel to the E and H planes of the radiator. Otherwise, the data may be presented as shown in Fig. 8,

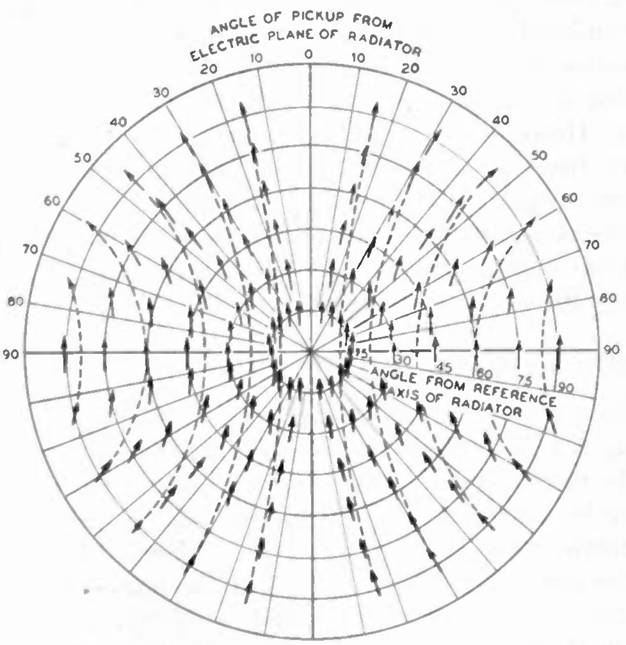

Fig. 8—Polarization characteristic of a typical antenna feed.

where the polarization is indicated by the direction of vectors plotted on a polar diagram representing the surface of a hemisphere, and the completed pattern gives a good picture of the polarization at all points on the surface.

C. Measurement of Phase

The phase of the wave front of the radiated wave relative to an arbitrary reference surface may be measured by mixing the received signal with a sample taken from the generator and adjusting the phase of one of the signals to produce a null at the receiver. As illustrated in the block diagram of Fig. 9, the generator delivers power to the radiator under test and to a branch circuit. The energy in the branch circuit passes through an attenuator and a phase shifter into a mixer, where it is combined with the signal from the pickup antenna and sent into the detector. The attenuator is set to equalize the signal through each path, and the phase shifter is adjusted to give a null in the output. At this position of adjustment the phase difference through the

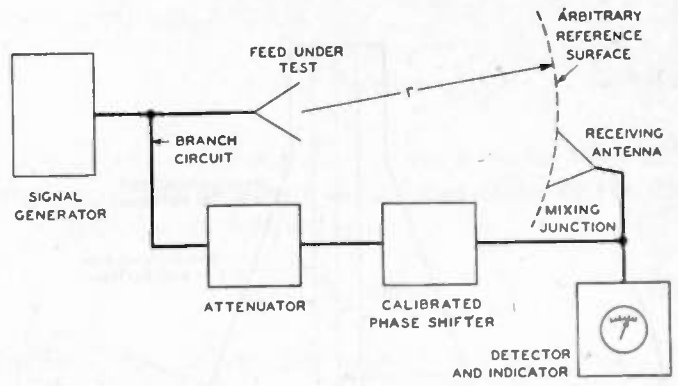

Fig. 9—Basic phase-measuring circuit.

two paths is an odd number of half-wavelengths. As the pickup is moved through the field, the phase, relative to an arbitrary reference point, may be measured. Since variable attenuators used at microwave frequencies also introduce phase shift, the attenuation is usually kept constant, reducing the signal in the branch path to about 6 decibels below the maximum signal to be measured. The phase shifter can be installed in either path, or it may be combined with the mixing junction in a standing-wave detector. In the latter case the branch path is connected to the probe of the standing-wave detector, and the pickup and receiver to either end. In this way the attenuation of the probe is in the path which otherwise has least attenuation, and a phase shift of 360 degrees is spread over a full wavelength of probe motion.

The phase variation may also be obtained by moving the pickup antenna toward and away from the feed antenna under test, in which case the phase is proportional to the distance from the feed. If the phase is measured by the position of the pickup required to produce a null, the successive positions of the pickup describe the phase front of the wave. This method has the advantage of simplicity of apparatus, but has more possibilities of error because the presence of the experimenter near the antenna may disturb the field being measured.

The phase data may be presented by plotting the shape of the phase front to scale and locating its center of curvature. Since the measurement must be taken at a reasonably great distance from the feed, the phase differential is usually a very small fraction of the path length, and this method of plotting is awkward to use. A more suggestive presentation is obtained by subtracting a constant length from the measured phase, thereby reducing the size of the constant-phase curves. This gives a diagram wherein the deviations from the ideal spherical (or circular) phase front are emphasized. An example of this presentation is shown in Fig. 10. The feed may also be sketched in, its size being consistent with the phase scale, to aid in co-ordinating the shape of the phase front with the geometry of the feed.

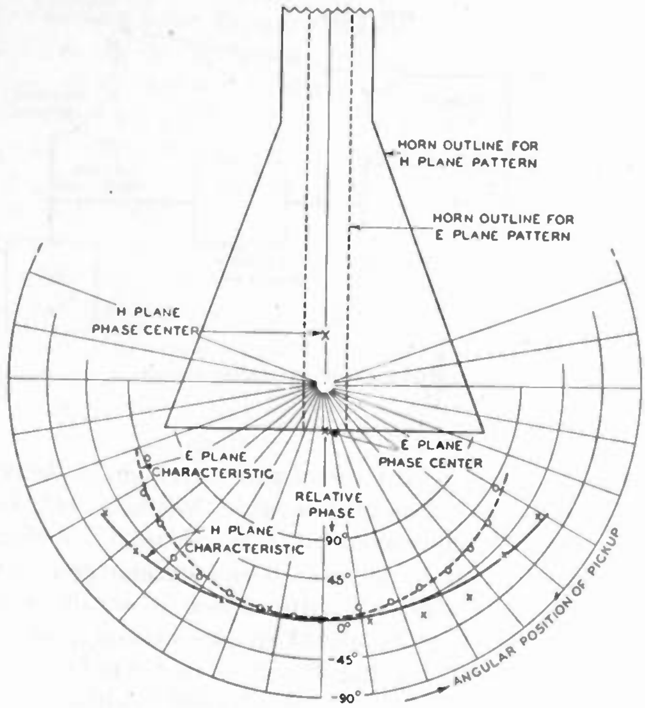

Fig. 10—Typical phase pattern.

It is sometimes desirable to measure the phase and amplitude distribution of a complete antenna. In this case the operation is the same as that described above, except that the pickup is usually moved along a straight line near and parallel to the antenna aperture. The data then indicates how much the phase front of the wave emerging from the antenna deviates from a plane.

V. Mutual Coupling Between Antennas

In some types of high-frequency systems the receiving and transmitting antennas are mounted in close proximity, and it is necessary to reduce the mutual coupling between them to a very small value in order to avoid interference or "cross talk." The measurement of directivity previously discussed is not directly applicable when the antennas are close and either subtends a large angle at the other. However, the average relative attenuation of the directivity patterns in directions which contribute to the mutual coupling may be used to obtain a rough idea of the coupling to be expected.

Cross-talk protection may be defined as the ratio of the power transmitted to the power received, and may be measured with a generator and a detector by comparing the signal transmitted through the mutual coupling of the antennas and through a calibrated attenuator. If the antennas are aligned face-to-face this ratio should be nearly unity, or zero decibels, and as the antennas are rotated away from each other the value should drop and go through wide excursions not unlike the minor lobes of a directional pattern, and finally drop to a very low value when the antennas are back-to-back. With high-gain antennas the coupling in the back-to-back condition may be largely caused by multiple reflections from surrounding objects or the ground, and therefore the site for such measurements should be free from interfering bodies or as like the actual operating site as practical.

VI. Requirements of the Antenna-Measuring Site

It is necessary that the antenna which is under study be placed in a suitable environment; otherwise the effect of the terrain and surrounding objects may introduce errors in measurement. At an ideal location the transmitted wave would arrive at the receiving antenna as a true plane wave, being uniform in amplitude and having a flat wave front over the entire antenna aperture. However, when the departure from a plane wave front is excessive or the field distribution becomes irregular, the measurements will be in error. The degree of variation from the ideal that can be tolerated depends upon how these variations arise and the precision of measurement required.

A. Distance Requirements

Since the wave emerging from the transmitter antenna is spherical, the phase front across the aperture of the receiving antenna will be flat only when the distance between antennas is infinite, and for any finite separation the phase front will be curved. The extent of this curvature or the amount of the phase deviation in terms of the separation r and aperture dimension a can be deduced with the aid of Fig. 11. The path length

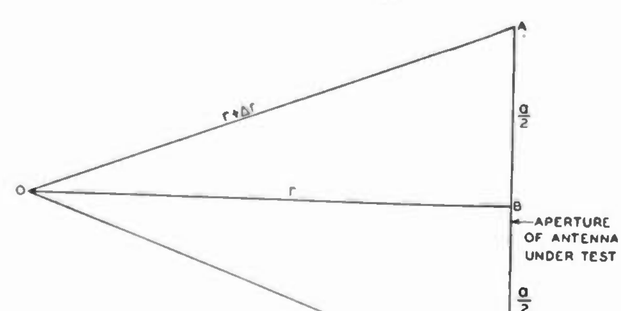

Fig. 11—Calculation of phase deviation due to path length.

of the extreme ray OA is $r+\Delta r$, and solving the right triangle OAB, we get

$$(r + \Delta r)^2 = r^2 + \left(\frac{a}{2}\right)^2 \qquad (8)$$

and, neglecting $(\Delta r)^2$, we have

$$\Delta r = \frac{a^2}{8r}\cdot \qquad (9)$$

Here $\Delta r \times (180/\pi)$ is the maximum phase deviation from a plane, in degrees, for an antenna of aperture a at a distance r from the source.

The effect of such a phase deviation on the measured antenna gain can be ascertained by a vector summation of the contributions of elementary areas of the antenna aperture. For example, for a phase deviation of $\pi/8$, the measured gain of an antenna having a uniformly illuminated aperture and a plane wave front will be in error by only 0.1 decibel, which is sufficiently accurate for most antenna work. Specifying, then, that the phase deviation across the antenna aperture be less than $\pi/8$, i.e.,

$$\Delta r \leqq \frac{\lambda}{16}, \qquad (10)$$

we obtain

$$r \geqq \frac{2a^2}{\lambda} \qquad (11)$$

as the required separation between transmitting and receiving antennas.

This distance requirement may be too lenient under certain conditions. For example, if the phase front of the wave emerging from the antenna under test is curved, as is the case in a horn antenna or a defocused paraboloid, the measured gain may be more seriously in error than that indicated above. Thus, if the test antenna were out of focus so as to produce an additional $\lambda/16$ phase curvature (bringing the total to $\lambda/8$), the measured gain would be in error by 0.3 decibel. Gain measurements on optimum horns are subject to this type of error, since they have a large phase curvature. In Fig. 3, for example, it is seen that at $r=2a^2/\lambda$ the measured gain is approximately 0.4 decibel low. Of course, if the antenna is out of focus in the direction to counteract the effect of the short path, the measurements would give correspondingly optimistic results.

B. Directivity Requirements of Distant Antenna

The directivity of the radiator at the far end of the transmission path should be broad enough to give a substantially uniform field across the aperture of the antenna under test. However, if the antenna directivity is too broad, the presence of surrounding objects and the ground in the field of the beam will produce reradiations which distort the direct wave. These spurious reradiations can be reduced considerably by a choice of testing site relatively free from objects in the main portion of the beam and by increasing the size of the distant radiator. There is, however, a limit to its maximum size, for if the antenna is too large the beam will be so narrow that the test antenna is not uniformly illuminated. Limiting the amplitude variation to less than $\frac{1}{4}$ decibel requires that the directivity of the distant antenna be such that one-eighth of its beam angle between nulls be greater than the angle subtended by the antenna under test, which is

$$\theta = \frac{a}{r} \text{ radians} \qquad (12)$$

where a is the width of the antenna under test. This angle, expressed in terms of the dimensions of the distant antenna, is approximately

$$\theta = \frac{\lambda}{4a_T} \text{ radians} \qquad (13)$$

where a_T is the width of the distant antenna. Thus,

$$\frac{a}{r} \leqq \frac{\lambda}{4a_T} \qquad (14)$$

and

$$a_T \leqq \frac{r\lambda}{4a}\cdot \qquad (15)$$

This expression specifies the maximum size of the distant antenna. In some instances, even with an antenna of the maximum size, reradiation from the ground is still objectionable. Where the ground cannot be removed from the field, some precautions can be taken to reduce its effect upon the measurement of pattern and gain.

C. Elimination of Ground-Reflection Effects

If the ground between the test and the distant antennas is smooth and the reflection path unobstructed, the testing site may be located at a point where the direct and reflected rays add to give a satisfactory phase

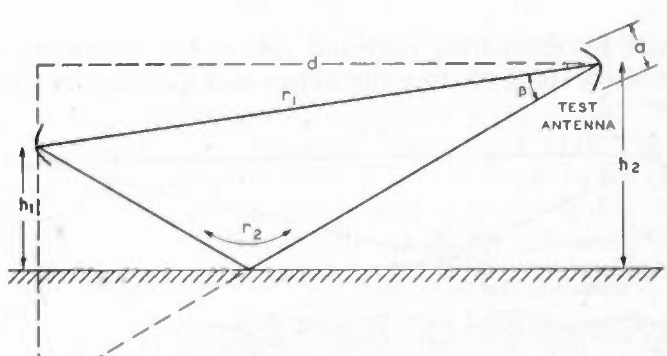

Fig. 12—Path-length determination.

and amplitude distribution of energy. The difference in the path lengths between the direct and reflected rays (Fig. 12) is

$$r_2 - r_1 = \sqrt{d^2 + (h_2 + h_1)^2} - \sqrt{d^2 + (h_2 - h_1)^2} \qquad (16)$$

which, if h_2 and $h_1 \ll d$, becomes

$$r_2 - r_1 = 2\frac{h_1 h_2}{d}\cdot \qquad (17)$$

For perfectly conducting ground the two fields add to give a variation with altitude:

$$E = 2 \sin 2\pi \frac{h_1 h_2}{\lambda d}. \tag{18}$$

This is shown on the dashed curve in Fig. 14. The first maximum occurs when

$$r_2 - r_1 = 2\frac{h_1 h_2}{d} = \frac{\lambda}{2}. \tag{19}$$

If the antenna under test is at the first maximum and

$$a \leqq \frac{h_2}{4}, \tag{20}$$

the intensity will not vary more than 0.2 decibel over the aperture. Also, since the angle between the distant antenna and its image as seen from the testing site at h_2 is approximately

$$\beta = \frac{2h_1}{d}, \tag{21}$$

we have, from (19),

$$\beta = \frac{\lambda}{2h_2}. \tag{22}$$

Now the beam width between nulls of the antenna under test is

$$\alpha \geqq \frac{2\lambda}{a} \tag{23}$$

or, from (20),

$$\alpha \geqq \frac{8\lambda}{h_2}. \tag{24}$$

Thus, by comparing (22) and (24) it can be seen that the angle subtended by the source and its image is 1/16

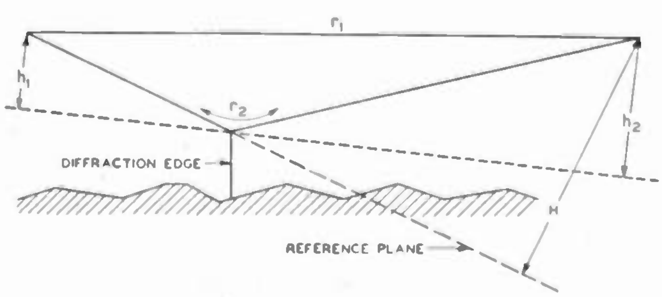

Fig. 13—Use of a diffraction edge.

of the beam angle of the largest antenna allowed by the limitation of the permissible amplitude variation at the first field maximum. To broaden the field maximum at the testing site, the antenna at the distant end of the path should be as near the ground as practical and the testing site elevated to the height of the first field maximum, according to (19).

If the ground is uneven, the field at the testing site may be further distorted. This effect may be avoided by using a straight diffraction edge (which may be a wire fence of fine mesh) perpendicular to the transmission path between the antennas, high enough to shield the antennas from direct ground reflections. In this case, as indicated in Fig. 13, the diffracted field may be obtained from diffraction theory (Cornu's spiral), and is shown in Fig. 14. The first maximum occurs when

$$(r_1 - r_2) = \frac{12}{14}\frac{\lambda}{2} = \frac{3}{7}\lambda,$$

and the requirements are almost the same as before, except that, in order to use the formulas developed,

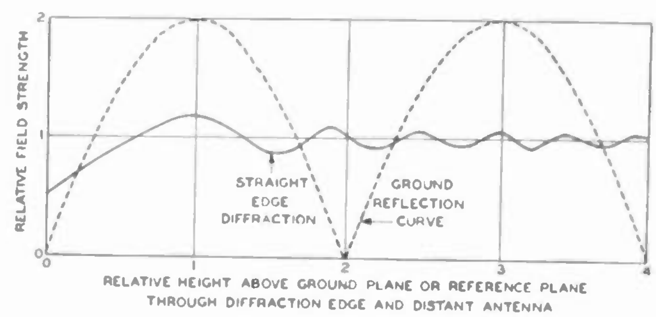

Fig. 14—Function of field strength versus height for antenna-testing site.

h_1 and h_2 must be measured from a plane through the straight edge making equal angles with lines drawn from the straight edge to either antenna. However, in this case the fluctuation in amplitude and phase decreases with height and may be avoided for most practical purposes by raising the antennas so that $r_1 - r_2$ is many wavelengths.

A desirable testing arrangement is to use several low fences[5] to shield the testing site from ground reflections, and then to operate with h_2 and h_1 large enough to be substantially above the irregularities of field strength caused by diffraction at the fences. This arrangement is less critical than operating on the first maximum of the curve, but may involve greater field fluctuations over the antenna aperture.

VII. Components of Measuring System

Most of the components necessary for measuring antenna characteristics have been introduced in the previous discussion; however, for completeness we will tabulate the equipment required. A simple system for gain and pattern measurement is shown diagrammatically in Fig. 15, and a more complex system in Fig. 16. Fig. 17 shows a typical testing site.

The units of the systems are as follows:

1. Microwave signal generator. The emitted power may be c.w. or i.c.w., as required by the receiver to be used.

[5] This procedure was used by A. L. Robinson of the Bell Telephone Laboratories.

2. Adjustable radio-frequency attenuator. This serves as a control of the signal level, and provides some impedance isolation between the antenna and the associated circuit.

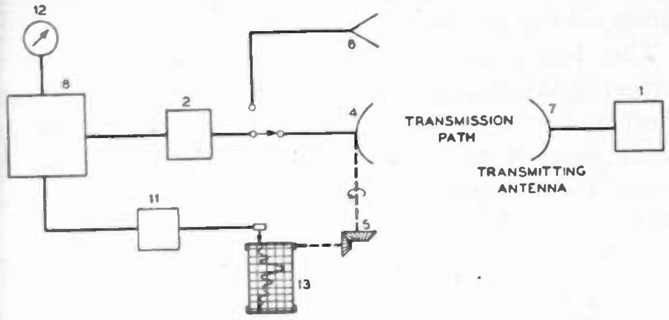

Fig. 15—Simple system for gain and pattern measurement.

3. Standing-wave detector for monitoring the transmission and checking the impedance of the antenna under test (Fig. 16).

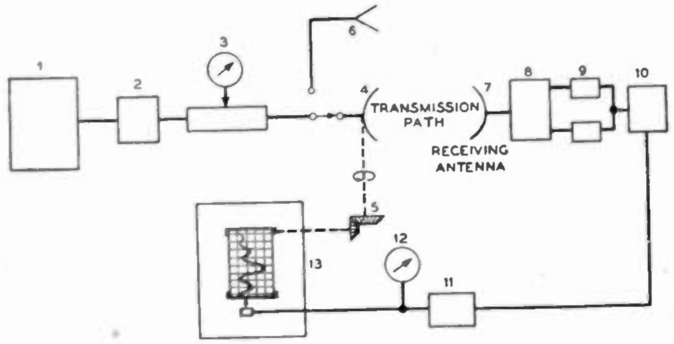

Fig. 16—Complete system for gain-pattern and cross-polarization measurement.

4. Antenna under test.

5. Turntable, coupled to the recorder, for orienting the antenna under test. Preferably the mounting should permit both horizontal and vertical axes of rotation.

6. Gain standard of comparison. (See Section II.)

7. Antenna for transmitting or receiving at distant path terminus. The directivity and location of this unit should be consistent with the requirements of Section VI.

8. Depolarizer: A wave-guide device for separating energy of either polarization. See Section III-C (Fig. 16).

9. Receiver: Maximum sensitivity is obtained by the use of a double-detection receiver, where a beating oscillator, in conjunction with a crystal detector, is used to convert the received signal to an intermediate frequency. Carefully calibrated intermediate-frequency attenuators can then be used to measure, with considerable accuracy, signal ratios as large as 80 to 120 decibels, and, since the beating oscillator maintains a strong signal on the crystal, the weak received signal will be linearly detected by the crystal and no concern need be exercised regarding the crystal characteristics. In such receivers, the sensitivity is limited only by the first-circuit noise impaired by the receiver noise figure. Since the first-circuit noise is KTB,[6] and amounts, at 63 degrees Fahrenheit, to 4×10^{-21} watts per cycle bandwidth, and the receiver noise figure lies between 5 and 15 decibel, it is apparent that large differences in signal level can be measured even with microwave sources having very low output power.

Where great sensitivity is not required, a single-detection receiver is sometimes used, employing a crystal or bolometer as the microwave detector. Such a system relies on the detector maintaining a square-law characteristic, and this is approximately true of both the silicon rectifier and the bolometer, provided the signal levels are low.

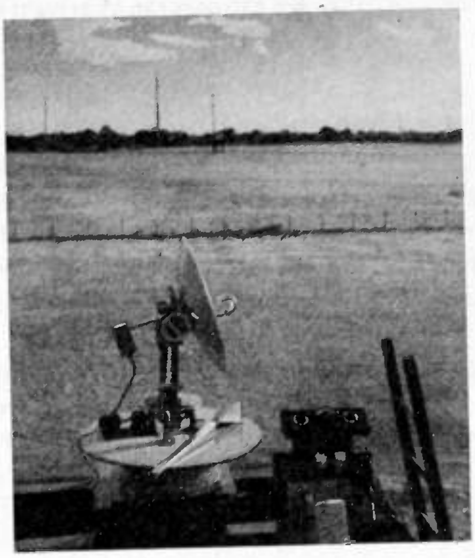

Fig. 17—Antenna-testing site.

With square-law detectors, the output voltage is proportional to the receiver power. Some signal-to-noise advantage may be gained in single-detection receivers by modulating or pulsing the microwave source, and equipping the receiver with a narrow-band audio amplifier. When a depolarizer is used as shown in Fig. 16, two square-law detectors (9) are employed, and their outputs combined and delivered to an audio amplifier (10).

10. Audio amplifier and noise filter (Fig. 16).

11. Adjustable low-frequency attenuator.

12. Indicating level meter.

13. Automatic pattern recorder. (See Section III-D.)

Except for the depolarizing system and the standing-wave detector, either of which may be dispensed with for most antenna work, the roles of the antennas as regards transmitting or receiving can be interchanged. There are many other possible arrangements of equipment, and the two shown are merely given as examples.

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

Slot Antennas*

N. E. LINDENBLAD†, SENIOR MEMBER, I.R.E.

Summary—The development of flush-type radiators of the slot and pocket type is described. Special emphasis is given to types applicable to aircraft. Specific solutions to altimeter and marker beacon pickup antennas are described. Reference to application in other fields is also made.

The general aspects of the phenomena which are involved are examined, and it becomes evident that workable solutions, in the majority of cases, can be obtained only by means of actual experiment, since variations in the surroundings have first-order influence upon such vital characteristics as radiation patterns, slot impedance, and bandwidth.

Progress before and during the war is described in somewhat chronological manner. It is pointed out that, while this progress has been considerable, an appreciable amount of skillful investigation remains to be done before slot antennas can be brought to maximum usefulness.

Introduction

FOR A NUMBER of years, extending throughout the war, the engineers of the RCA Laboratories Division at Rocky Point, L. I., N. Y., have been engaged in the study of such fundamental antenna problems as bandwidth, and the effect of surroundings and location of antennas upon their radiation characteristics.

During this development period the speed of aircraft has been greatly increased. Consequently, streamlining became a necessary consideration. An all-out effort to provide for efficient radiation from flush surfaces was made in order to meet this increasing need. The result of this work is the slot antenna. It comprises slots in the metal surface of an aircraft. These slots are backed by metal cavities inside the surface. Impedance regions exhibiting stability over widest possible frequency bands are chosen or arranged within the cavity for connection to the feed lines.

It is the purpose to briefly review the general aspects of the problems involved in such designs and to describe somewhat chronologically the steps of development. It is hoped this description will serve as a stimulant to further developments.

General Considerations

An early idea that may be considered associated with so-called slot antennas was a scheme devised in 1939 by G. L. Usselman, of the RCA Laboratories Division at Rocky Point, L. I., N. Y., to feed an array of dipoles by a slotted wave guide. The dipoles were distributed along the slot and attached to its edges. By choice of phase-velocity characteristics of the wave guide thus loaded, either broadside or end-fire excitation could be achieved. Usselman also suggested that arrays of closely spaced dipoles may be replaced by continuous sheets of widths corresponding to the length of the dipoles.

This latter method had special merits worthy of further development, which was undertaken in a joint effort by the U. S. Navy, Radio Test, under Lieutenant Commander A. S. Born (now Captain, U.S.N.), and the Rocky Point Section of RCA Laboratories, beginning in 1941.

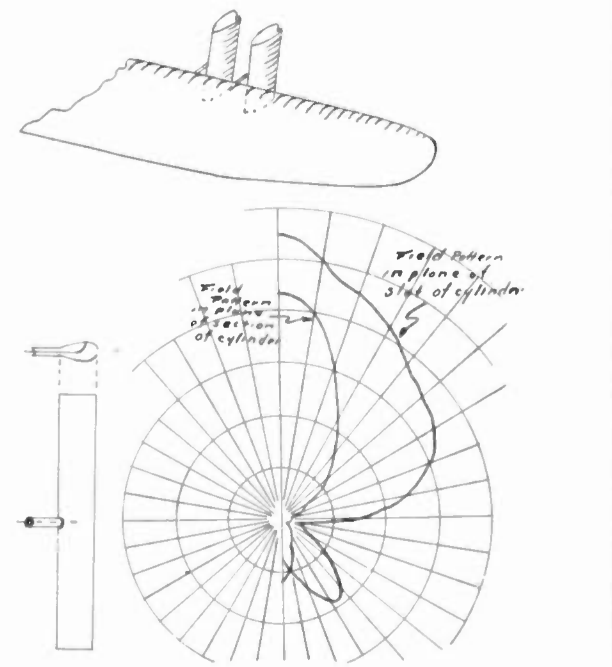

Fig. 1—Open-ended and streamlined, slotted cylinder antennas mounted in front of leading edge of airplane wing. Polarization perpendicular to cylinders. Polar radiation diagrams for one antenna element are included.

The primary purpose was to apply the slot-feed principle to airborne antennas. Slotted wave guides having teardrop or streamlined cross section were used in some of the early experiments (Figs. 1 and 2). No special antenna elements were attached to this streamlined body, but by arranging for co-operative coincidence of internal and external characteristics, its own exterior served as a radiator. While having great usefulness in other fields of application, the limited usefulness of slotted cylinders for airborne purposes became quickly evident in view of the advancing speeds of aircraft.

Antennas for a high-speed aircraft must not add external structure. The designer must consider the possibilities of providing for the emergence of radiation from the surface of the plane. The least radical procedure is, perhaps, to mount a conventional radiation element in an indentation in the plane surface which is then covered with a dielectric window. The primary radiation fields thus originate with a conventional radiation ele-

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† Radio Corporation of America, New York, N. Y.

nent. The cavity is open and nonresonant. The aperture, however, may be made smaller and the cavity itself be made resonant, eliminating the need of a distinct radiation element. The aperture may be in the form of a slot of sufficient dimensions for emergence of radiation from the interior of the cavity. A surface secton

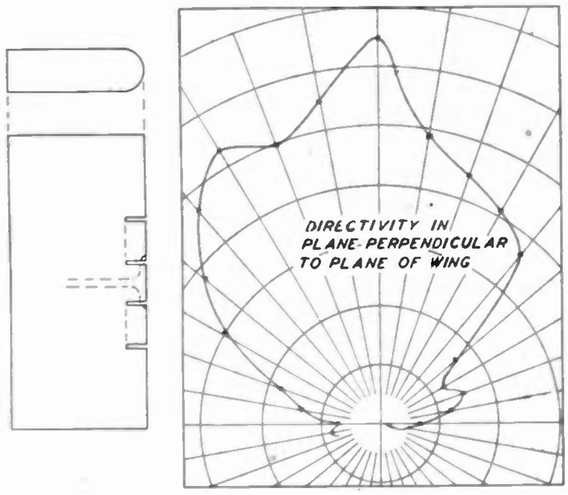

Fig. 2(a)—Combination of four slots across the leading edge of a wing, each pair having a common quarter-wave deep-backing cavity. Polarization in plane of the wing.

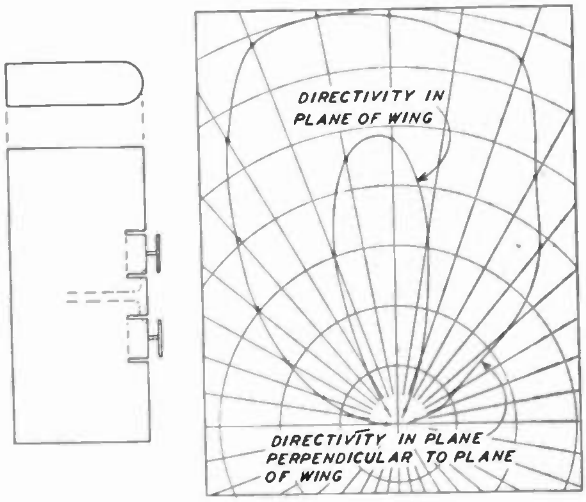

Fig. 2(b)—Same combination as in Fig. 2(a) with parasitic radiators added approximately one-quarter wave in front of each slot pair.

may be electrically uncoupled from the rest of the skin by means of cavity-backed slots. In this case the currents, in the separated area, isolated by the high-impedance slots, may be considered the origin of radiation. Actually, it is difficult to draw any definite lines of distinction between these various methods, since surface currents are always part of the diffraction phenomenon around apertures. Only when the apertures are very large relative to the wavelength, or when the aperture is isolated, by surrounding high-impedance slots, may the adjacent surfaces be considered as possessing a high degree of nonparticipation in the radiation phenomenon. The dimensions of the total metal area thus very often has considerable influence upon the radiation pattern, and may sometimes become the antenna itself.

It is now evident that the most controllable method is the one where an aperture or an area is isolated by high-impedance cavity-backed slots. Of these, the least cumbersome appears to be that of isolating an area. In cases when complete flush mounting is not required, it is possible to mount the isolating cavities like external pockets. They can also be made in the form of a sheet, rolled up like a jelly roll, forming a spiral cavity.

Development

One of the earliest attempts to utilize this idea was to cut pairs of half-wave-spaced vertical slots across the leading edge of an airplane wing. The resulting half-wave ribbon, which then was part of the leading edge, was backed by an approximately quarter-wave-deep cavity. Each side of the ribbon was connected at its maximum voltage point to a transmission line. These lines were then connected together in series or in parallel. This arrangement provided a rather wide, forward-spreading radiation pattern. When spaced coupled parasitic radiators or "directors" were placed a quarter-wave outside and in front of the strip between the paired slots,

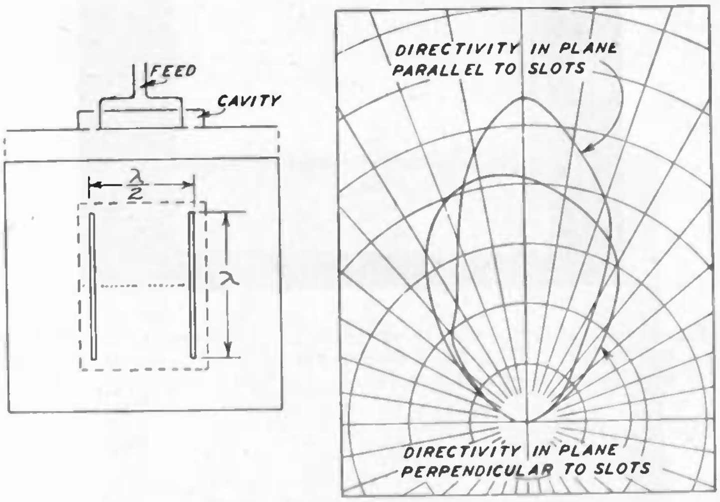

Fig. 3—Simple double-slot antenna.

higher gain was obtained, as shown in Fig. 2(b). This method, however, introduced difficult wing design and aerodynamic conditions, and was not continued.

The practical possibilities of isolating an area had been indicated, and experiments were directed toward flat surfaces. Fig. 3 shows the cross section of one of these early forms of double-slot antennas, affectionately dubbed "bathtubs" by the Navy. In Fig. 4 is shown a photograph of an antenna consisting of a pair of double-slot antennas. As can be seen, the spacing between adjacent slots of different pairs is less than a half-wave. This spacing was determined experimentally with the aim in view of obtaining the cleanest radiation pattern.

Figs. 5, 6, and 7 show typical radiation patterns and the standing-wave-ratio curve as measured by the Navy Radio Test group. Fig. 8 shows a form by means of which it was possible to obtain wider frequency response. As may be noted, the center conductor of the transmission line here expands gradually as a flat wedge before connecting to the slot edge.

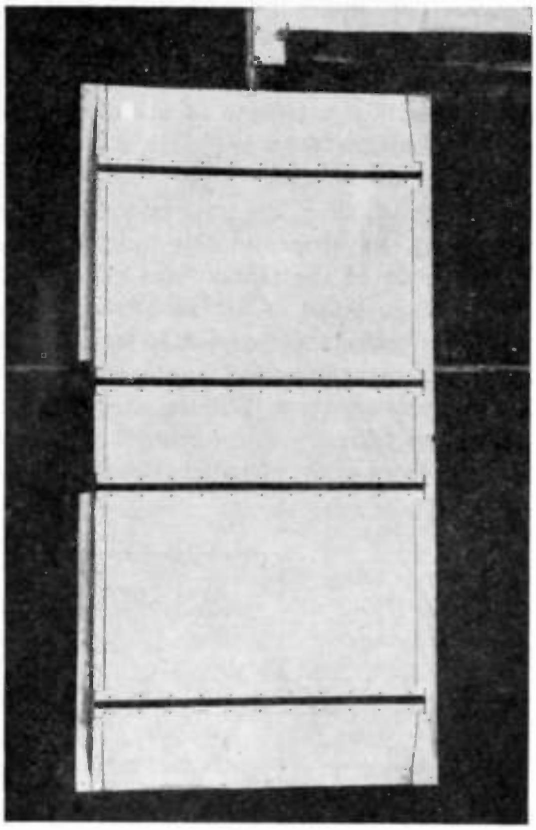

Fig. 4—Pair of double-slot three-quarter by half-wave antennas. This is the Navy "bathtub." Note close spacing between pairs for elimination of secondary lobes. Design by U. S. Navy.

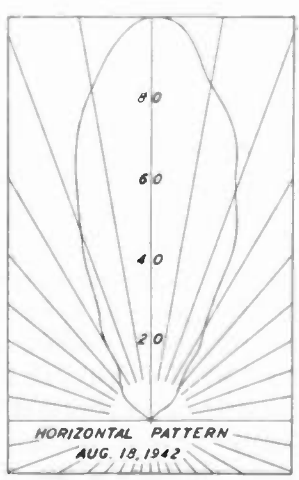

Fig. 5—Slot-crosswise pattern of antenna of Fig. 4. This data taken by U. S. Navy.

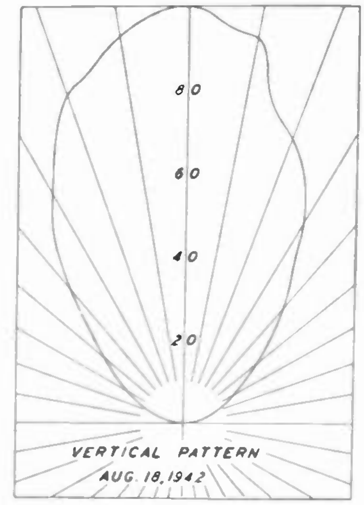

Fig. 6—Slot-lengthwise pattern of antenna of Fig. 4. Data by U. S. Navy.

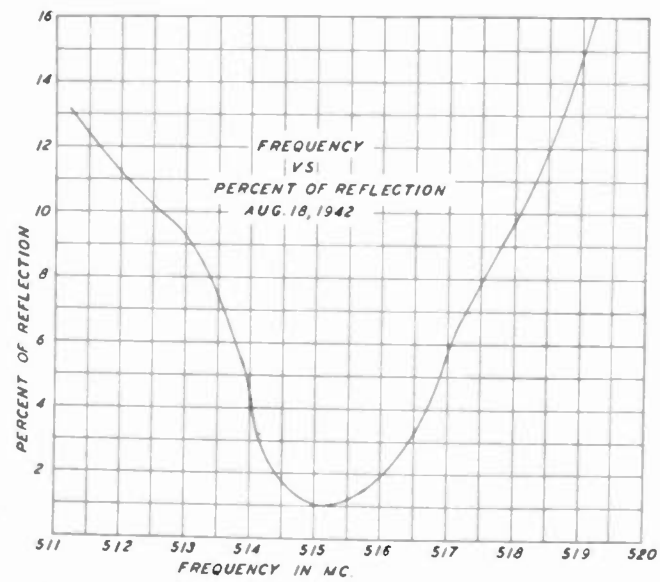

Fig. 7—Standing-wave-ratio curve taken from antenna of Fig. 4. Data by U. S. Navy.

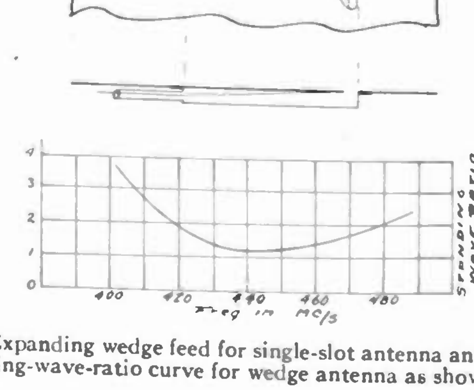

Fig. 8—Expanding wedge feed for single-slot antenna and the standing-wave-ratio curve for wedge antenna as shown.

It should be of interest to notice that the double-slot antennas possess a natural characteristic which is of advantage to lobe switching. If only one of the two slots is fed, the radiation pattern will lean toward the fed slot.

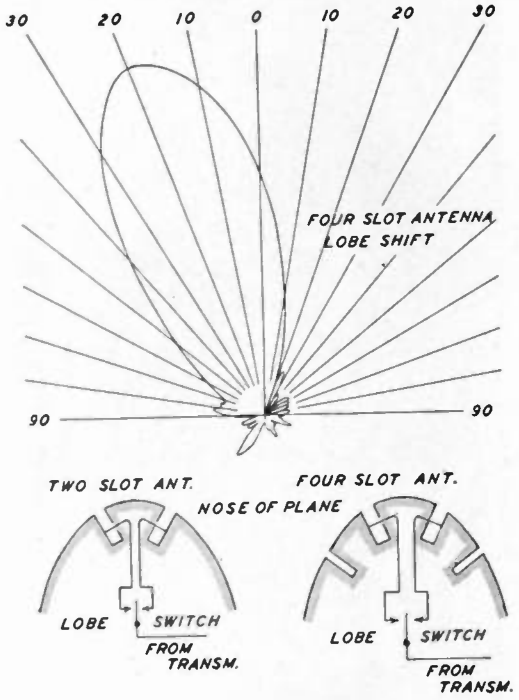

Fig. 9—Shiftable, single-slot feed for lobe switching. Note the application of this slot antenna to the curved contour of the model of the nose of an airplane.

In this way the same antenna can be used for both lobes by simply shifting the feed. This makes it equivalent to one slot per lobe (Fig. 9).

A particular use was made of this phenomenon in a double-slot antenna used as the focal primary in a parabolic-reflector-type antenna at 1250 Mc. It was found, however, that this design could be further simplified to permit the omission of a spark or contact switch. It was only necessary to provide a rotating patch of very small dimensions relative to the wavelength, which would alternately cover a portion of one or the other of the slots (Fig. 10). In order to provide up and down switching, the vertical slot was divided in two sections, parallel fed, but having high coupling impedance. In this way the patch, which was a piece of foil cemented in an eccentric position to an insulating rotatable disk, would cover the upper and the lower half of one slot, and then in sequence the lower and the upper half of the other slot. In this way the same effect as that obtained with the mechanically more difficult type of nutating dipole was obtained. A simplified form of nutating antenna energized by either a single- or by a double-slot primary was, however, also developed. It consisted of a diametrically resonant disk rotated eccentrically at a distance of about one-quarter of a wave or less in front of the double-slot antenna to which it was thus space-coupled (Fig. 11).

The work so far described served the useful purpose of proving the practical possibilities of nonprotruding radiators. It had been shown that the slot principle was sound and workable and that it furnished tools for a new approach to radiator problems. The development

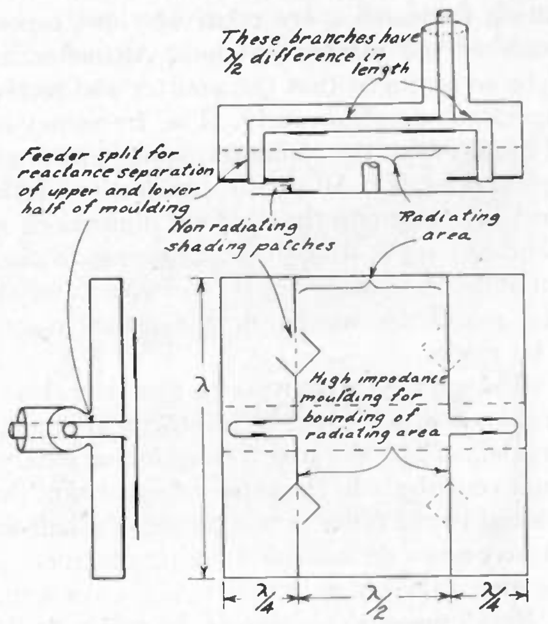

Fig. 10—External pocket-border type, one by one-half-wave surface radiator. As the primary in a parabolic reflector, the lobe switching is performed by 90-degree-displaced rotatable shading patches. Note electrical sectionalizing of pockets and feeders to facilitate "pull" by shading patches, a very practical arrangement. A single patch provides diagonal and a double patch vertical-horizontal lobe pulling.

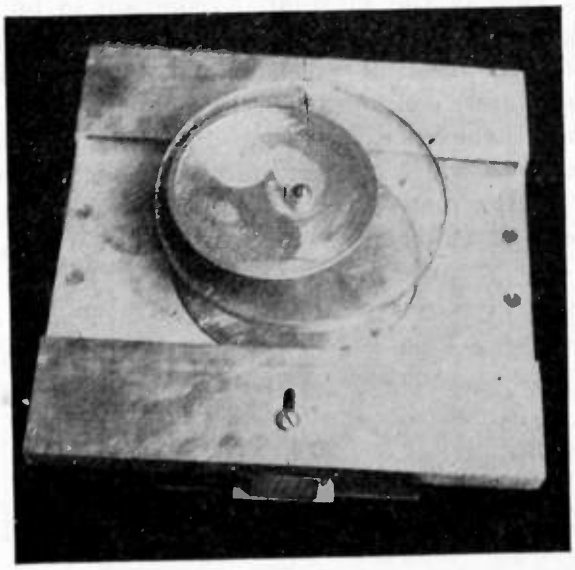

Fig. 11—External pocket-border type, one by one-half-wave surface radiator. The major radiating "area" is located between slots. The diametrically resonant disk in front, eccentrically rotatable, acts as a nutating facility for lobe switching when the combination serves as a focal radiator in a parabolic reflector system.

was, however, insufficient to meet some of the applications for which it was most needed.

Thanks to the interest shown by other researchers who would from now on contribute toward both the general and the special development of the slot principle, it was felt that the work could be directed toward specific applications. Slot antennas for altimeter and marker purposes were chosen as subjects of these efforts. These antennas would have to be applicable to all plane types, including the smallest and fastest. The operating frequencies are relatively low, especially in the case of the marker antenna. Altimeter antennas must be so arranged that transmitter and receiver may be operated simultaneously. The frequency-response band required by the altimeter equipment is also relatively considerable. All such considerations which have a direct bearing upon the antenna dimensions must, in the practical application of slot antennas to aircraft, be accommodated without sacrifice of structural strength. Careful search for minimum dimensions must, therefore, be made.

As a general rule, a cavity with generous cross-section dimensions makes it easier to meet wide-frequency-band requirements. The slot and corresponding cavity length does not contribute in the same way and can, therefore, be reduced to the order of magnitude of a half-wave before it becomes a serious band limiting factor.

The view taken here is that it is always well, in antenna developments, to aim at as much "self"-bandwidth as possible, since it eliminates or reduces either the need or the complexity of impedance-correcting networks. The power of the network method in practical application is, reversely, greatly facilitated by good primary frequency response, especially in cases of exacting requirements of low reflection.

In view of the substantial reduction in bulk that could be obtained by the use of single slots, it was decided that these be given careful consideration. Although future equipment developments may not permit the use of the less exactly shaped radiation patterns obtained by single slots, their other virtues made them appear as the most practical expedients at the present stage of development.

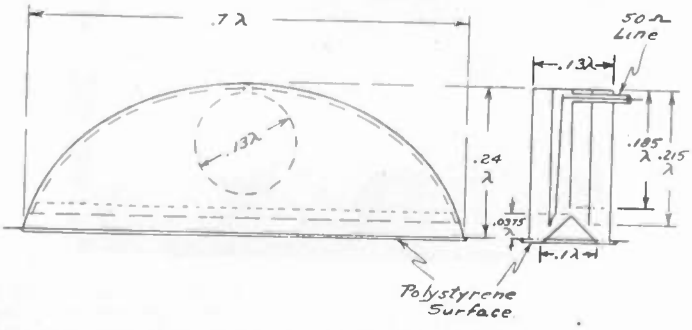

Fig. 12—Elliptic-cavity, wide-band, twin-slot antenna. The slots are spaced one-tenth of a wavelength.

At first, bandwidths greater than needed were aimed at. The opening of a longitudinally elliptic cavity was partitioned by means of a longitudinal strip to form two closely located, parallel slots (Figs. 12 and 13). The cavity was likewise partitioned by a longitudinal wall. Point connections were made between this partition and the narrow strip between the slots. Thus a certain amount of internal coupling was maintained between

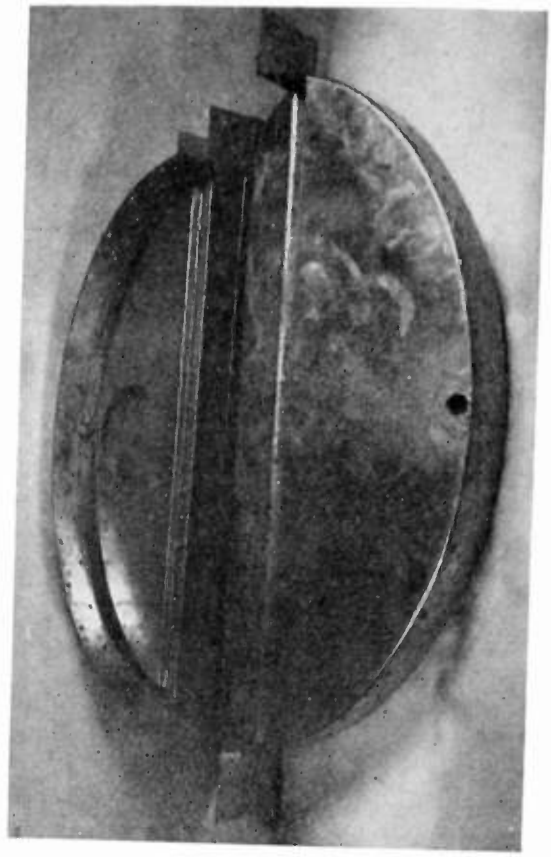

Fig. 13—Showing cavity sections of the antenna of Fig. 12. Note the hole in the feed tongue which divides it into two curved, parallel-connected expanded wedges.

the two half-sections. Each cavity half-section was again partitioned. The transmission line entering the cavity at bottom center connected to the top edge of one of the side partitions by means of an elliptic tongue.

Fig. 14—Standing-wave-ratio curve of the antenna of Fig. 12 as compared with the standing-wave-ratio of the antenna without subdivided cavity.

This multiple partitioning was an expedient by means of which a region of frequency-flat impedance balance could be obtained which was suitable for direct connection to the transmission line. A bandwidth of 30 per cent at a 2:1 reflection tolerance was obtained (Fig. 14). These experiments were done with a slot length of 0.75 wavelength and a maximum cavity depth of 0.2 wavelength. The cavity width was 0.135 wavelength.

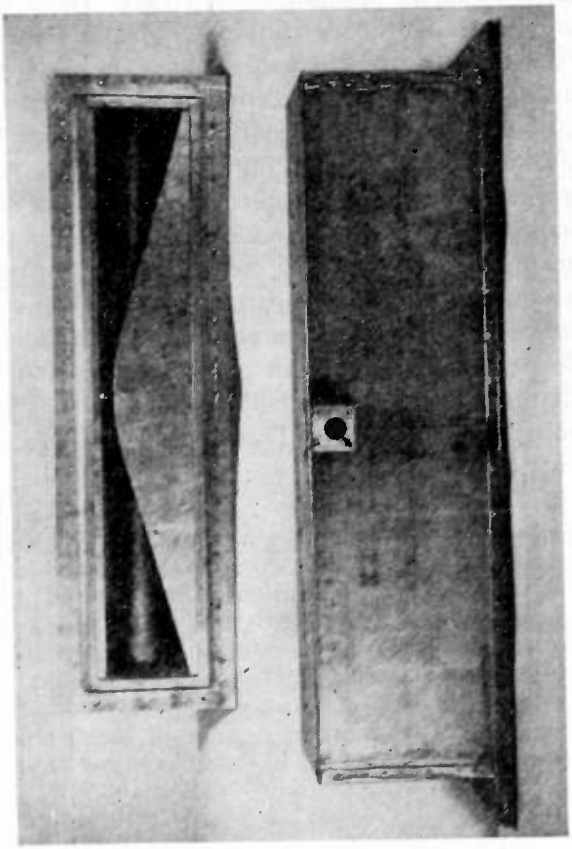

Fig. 15—Rolled-feed tongue, keyhole-slot antenna. Slot length, 0.575 wavelength. The rolled tongue can be seen inside the cavity.

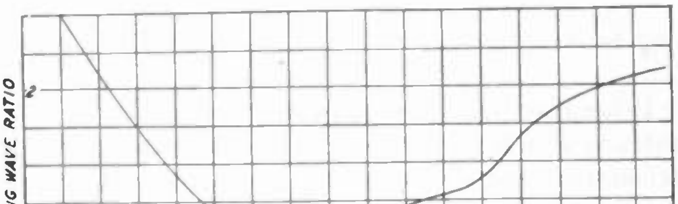

Fig. 16—Standing-wave-ratio curve of the antenna of Fig. 15.

Further reduction of size, however, continued to be very desirable. It appeared that this would then have to be done at the expense of bandwidth. In the case of the altimeter antenna, a search was made for the minimum dimensions required for the maintenance of the necessary 10 per cent bandwidth. A series of tests were carried out in which the influence of varying size and parameters was carefully noted. The partitions were eliminated. The tongue feed was maintained, but the tongue was now bent over itself and shape and position determined empirically for best conformity with region of flat impedance balance (Figs. 15 and 16). In the illustrated example the slot and cavity length was 0.575 wavelength, the cross section was 0.1 by 0.135 wavelength. The bandwidth at 2:1 reflection standard was about 14 per cent.

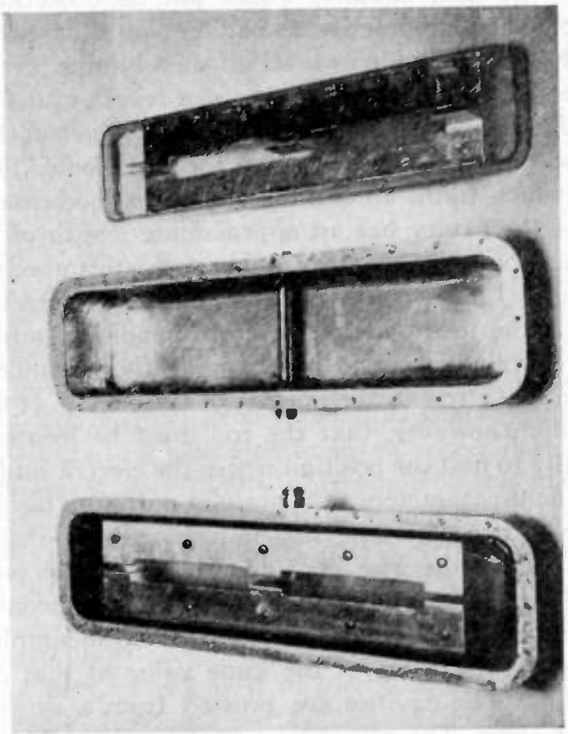

Fig. 17—Three-point capacitance-loaded *H*-slot altimeter antenna. Slot length, 0.4 wavelength.

As may be anticipated, further reduction of size calls for capacitance loading of the cavities at their open end.

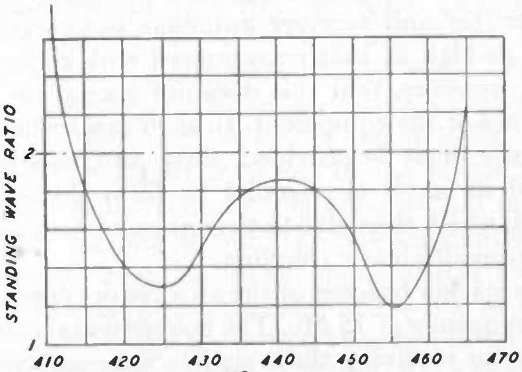

Fig. 18—Standing-wave-ratio curve for the antenna of Fig. 17.

This was first done by using narrower slots and eventually by adding capacitances across the slots. In these continual attempts to reduce the cavity dimensions it became necessary to revise the feed methods from time to time, since their operation is subject to certain parametric conditions.

In the model which was considered sufficiently small to be practical, the attempts to establish a suitably broad impedance zone within the antenna proper into which the feed system may be introduced had not been entirely satisfactory. It was found that a simpler expedient could be had by resorting to external compensation by means of a line stub. This, however, was convenient only because such relatively high standing-wave ratios as 2:1 or 1.5:1 were considered permissible so that the resulting hump in the s.w.r. curve, often unavoidable with such circuits, would not be objectionable. More complicated networks can, of course, be provided which will subdivide or level off such bumps.

The antenna of Fig. 17 has a slot length of 0.4 wavelength. The cross section is 0.1 by 0.1 wavelength. Fig. 18 shows a typical s.w.r. curve for this model. The tail stub which forms a continuation of the feed conductor across the cavity has an approximate length of three-quarters of a wavelength. Its characteristic impedance is 50 ohms. Line stubs of other impedance values can be used if the feed coupling is correspondingly adjusted. This coupling is varied by changing the distance from the coupling rod to the bottom of the cavity. It should be noted, however, that the rod must be located empirically to find the position where the electric and magnetic field parameters co-operate in optimum fashion at at given characteristic line impedance and standing-wave ratio. Otherwise, considerable bandwidth is easily lost. Using a stub having a characteristic impedance of 50 ohms appeared satisfactory and aided in simplifying the system by being of the same value as that of the feed line. The cavities are pressed from a single aluminum sheet. The cover consists of Formica. The total weight of a complete antenna and stub combination to operate at a midfrequency of 440 Mc. is one pound.

In applying such antennas to altimeter equipment, where transmitter and receiver must operate simultaneously, it has been found that the coupling between transmitter and receiver antennas is generally about twice as high as that encountered with dipoles. It appears, however, that this does not exceed the coupling tolerance of the equipment. In such cases where smaller tolerance must be provided, these antennas can be arranged in series or parallel to form double-slot antennas, which then, due to the nature of such a combination, provides lesser coupling.

The marker beacons of the airways operate at present on a frequency of 75 Mc. The conventional external antennas for receiving these signals are cumbersome and inefficient due to this rather low frequency. A slot antenna of such small dimensions as 20×4×5 inches has been developed. The slot length of this antenna is only 0.125 wavelength. The slot is heavily capacitance loaded. As can be understood, an antenna of such dimensions relative the wavelength must of necessity have a very high Q. The s.w.r. curve obtained by the aid of a series stub line is shown in Fig. 19. The signals obtained are better than equal to those obtained with external wire antennas located close to the ship. Greater bandwidth would be desirable, but it is gratifying at this stage of slot-antenna development to be able to report that antennas of such small dimensions relative to the frequency will work at all.

The marker-beacon signals do not call for any large bandwidth. The chief reason for not wishing to apply antennas of too narrow bandwidths is their sensitivity to moisture and ice. A sharply tuned antenna is easily detuned by small reactance variations. Trimming capacitors or inductances can, of course, be used, but they are not very satisfactory. It appears better to rely on means for preventing internal condensation of moisture, as well as both internal and external ice formation. Danger of external icing can be greatly reduced by proper location. The problem of eliminating internal condensation is more formidable. The antenna has either to be kept perfectly sealed to all the pressure variations to which it is subjected as an airplane changes altitude or it must be thoroughly ventilated. The latter is the easiest but does not at times appear entirely adequate, unless heating elements or moisture-absorbing substances be added. This again is, of course, not very attractive.

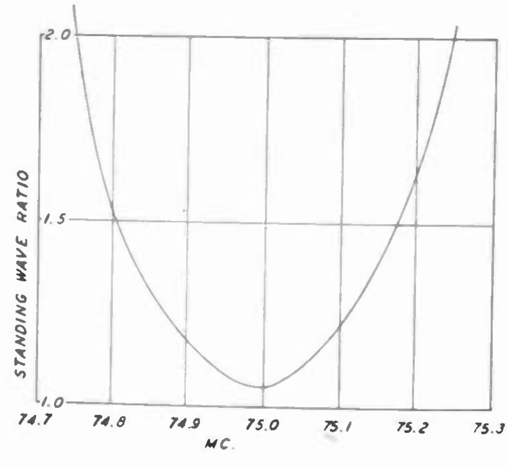

Fig. 19—Standing-wave-ratio curve for marker-beacon antenna.

It has been suggested that the dimensions of a cavity antenna may be decreased by filling it with a dielectric. When analyzing this proposal relative to radiation resistance and bandwidth, there appears to be nothing to gain by such procedure. A magnetic material operable at very-high frequencies would, on the other hand, provide a filler by which the circulating currents in the cavity could be reduced and bandwidth gains could be made. No such material is known at the present.

Dielectric fillers have, however, been considered for the purpose of keeping out the moisture of cavities. Desirable characteristics for such material are: low dielectric constant, low loss, homogeneity, and low weight. Foam is unsuitable, due to surface losses throughout the material. Boundary losses appear to be difficult to avoid even with the use of very solid materials.

Conclusion

An attempt has been made to describe the general aspects of slot antennas. Such antennas are a "must" in high-speed aeronautics and in radio-controlled missiles.

It has been shown that many of the tasks performed by external antennas can be performed by this flush-type radiator. Subjected to careful scientific investigation, as is possible in peacetime, their usefulness should eventually be greatly extended.

Acknowledgment

The author is indebted to the Navy Radio Test personnel and especially to Captain A. S. Born, R. M. Silliman, and Lieutenant J. B. Stout for encouragement during the early stages of development. Similar acknowledgment is due to various members of the radio technical groups of the Army.

To RCA, special acknowledgment is due to H. H. Beverage, C. W. Hansell, P. S. Carter, R. E. Franklin, and W. A. Miller for help and guidance freely given.

Fundamental Limitations of Small Antennas*

HAROLD A. WHEELER†, FELLOW, I.R.E.

Summary—A capacitor or inductor operating as a small antenna is theoretically capable of intercepting a certain amount of power, independent of its size, on the assumption of tuning without circuit loss. The practical efficiency relative to this ideal is limited by the "radiation power factor" of the antenna as compared with the power factor and bandwidth of the antenna tuning. The radiation power factor of either kind of antenna is somewhat greater than

$$\frac{1}{6\pi}\frac{Ab}{l^3}$$

in which Ab is the cylindrical volume occupied by the antenna, and l is the radianlength (defined as $1/2\pi$ wavelength) at the operating frequency. The efficiency is further limited by the closeness of coupling of the antenna with its tuner. Other simple formulas are given for the more fundamental properties of small antennas and their behavior in a simple circuit. Examples for 1-Mc. operation in typical circuits indicate a loss of about 35 db for the I.R.E. standard capacitive antenna, 43 db for a large loop occupying a volume of 1 meter square by 0.5 meter axial length, and 64 db for a loop of 1/5 these dimensions.

I. Introduction

An antenna whose dimensions are much less than the wavelength is subject to limitations which can be expressed by simple formulas. These limitations are fundamentally about the same for a capacitor used as an electric dipole and an inductor (loop) used as a magnetic dipole, if they occupy equal volumes. Either type may have some advantages resulting from variations within this rule or from relative facility in coupling with the associated circuits. This paper is directed to a few of the simplest formulas, and to their significance and application rather than their derivation. The small antenna to be considered is one whose maximum dimension is less than the "radianlength." The radianlength is $1/2\pi$ wavelength; it proves to be a logical unit for this purpose and a convenient one for simplifying the concepts and formulas. The approximations involved within this size depend only on the closeness between an angle and its sine up to ½ radian (4 per cent error). An antenna within this limit of size can be made to behave essentially as lumped capacitance or inductance, so this property is assumed.

It has occasionally been pointed out that a small antenna free of dissipation could take from a radio wave and deliver to a load an amount of power independent of the size of the antenna. This would be true at one frequency if the antenna can be resonated at that frequency without adding dissipation. It results from the fact that a smaller antenna delivers its lesser voltage from a lesser resistance such that the available power remains the same.

The power available from such an antenna is the wave power which would pass through the "effective area" of the antenna. Its effective area is 3/2 the area of a circle whose radius is one radianlength, denoted a "radian circle." The factor 3/2 is the power ratio of the directive gain of a small antenna relative to a theoretical antenna conceived to radiate equally in all directions over the sphere, denoted an "isotropic" antenna. This factor results from the fact that a small dipole (electric or magnetic) radiates in a doughnut pattern which effectively fills only 2/3 of the entire solid angle of a sphere.

Formulas for the efficiency of transmission through space may be stated in terms of the power actually radiated from the transmitting antenna and the power theoretically available from the receiving antenna to a load. In each case, the unavoidable dissipation in the coupling circuit (from generator to antenna or from antenna to load) limits the output to only a fraction of the power input. This fraction is the efficiency of the coupling circuit.

While the radiation pattern and hence the directive gain of a small antenna remain the same for a smaller size, the radiation resistance decreases relative to the other resistance in the coupling circuit. The resulting reduction in coupling efficiency is one of the principal limitations of the smaller antenna.

Another aspect of the same limitation relates to the frequency bandwidth of operation with fixed values of

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the circuit elements. A smaller antenna with the same reactance and radiation resistance must be more sharply tuned to deliver its available power. Therefore, the reduction of size imposes a fundamental limitation on the bandwidth. If the bandwidth so limited is insufficient, further damping must be added at the expense of coupling efficiency.

The limitations verify the experience that larger antennas are generally more efficient, especially for wide-band operation.

By expressing the formulas in fundamental forms, the inherent similarity of the electric and magnetic radiators becomes apparent, as well as the minor differences resulting from the use of available materials and structures.

II. Symbols

a = radius of circular cylindrical volume (meters)
A = area of base of cylindrical volume (meters^2)
b = height of cylindrical volume (meters)
n = number of turns of coil
k_a = shape factor of capacitor = effective area/actual area (A)
k_b = shape factor of inductor = effective length/actual length (b)
C = capacitance of antenna (farads)
L = inductance of antenna (henries)
ω = radian frequency (radians/second)
λ = wavelength (meters)
$l = \lambda/2\pi$ = radianlength (meters)
ϵ = electric permittivity in free space (farads/meter)
μ = magnetic permeability in free space (henries/meter)
k_e = relative permittivity of core in capacitor
k_m = relative permeability of core in inductor
$R = 120\pi = 377$ = wave resistance in free space (ohms)
$G = 1/R$ = wave conductance in free space (mhos)
R_e, R_m = radiation resistance in series with antenna (ohms)
G_e, G_m = radiation conductance in parallel with antenna (mhos)
R_t, G_t = series resistance or shunt conductance in tuner (ohms, mhos)
C_t, L_t = shunt capacitance or series inductance in tuner (farads, henries)
p_e = radiation power factor of capacitor antenna (electric dipole)
p_m = radiation power factor of inductor antenna (magnetic dipole)
k_c = coefficient of coupling between antenna and total capacitance
k_i = coefficient of coupling between antenna and total inductance
k_c^2 = efficiency of coupling of antenna to total capacitance = electric energy in antenna/total electric energy in tuned circuit
k_i^2 = efficiency of coupling of antenna to total inductance = magnetic energy in antenna/total magnetic energy in tuned circuit
e = radiation efficiency of antenna circuit.

III. Formulas

(C) (L)

Capacitance and inductance:

$$C = \epsilon \frac{k_a A}{b}; \qquad L = \mu n^2 \frac{A}{k_b b}. \tag{1}$$

Susceptance and reactance:

$$\omega C = G \frac{k_a A}{bl}; \qquad \omega L = R n^2 \frac{A}{k_b b l}. \tag{2}$$

Radiation shunt conductance and series resistance:

$$G_e = \frac{G}{6\pi}\left(\frac{k_a A}{l^2}\right)^2; \qquad R_m = \frac{R}{6\pi}\left(\frac{nA}{l^2}\right)^2 = 20\left(\frac{nA}{l^2}\right)^2 \tag{3}$$

$$R_e = \frac{R}{6\pi}\left(\frac{b}{l}\right)^2 = 20\left(\frac{b}{l}\right)^2; \quad G_m = \frac{G}{6\pi n^2}\left(\frac{k_b b}{l}\right)^2. \tag{4}$$

Radiation power factor:

$$p_e = \frac{G_e}{\omega C} = \frac{1}{6\pi}\frac{k_a A b}{l^3}; \qquad p_m = \frac{R_m}{\omega L} = \frac{1}{6\pi}\frac{k_b A b}{l^3}. \tag{5}$$

Coupling efficiency, connected as in Fig. 2:

$$k_c^2 = \frac{C}{C + C_t}; \qquad k_i^2 = \frac{L}{L + L_t}. \tag{6}$$

Circuit efficiency, connected as in Fig. 2:

$$e = \frac{G_e}{G_e + G_t}; \qquad e = \frac{R_m}{R_m + R_t}. \tag{7}$$

Circuit efficiency, in general:

$$e = \frac{k_c^2 p_e}{k_c^2 p_e + p_t}; \qquad e = \frac{k_i^2 p_m}{k_i^2 p_m + p_t}. \tag{8}$$

IV. The Antenna

Fig. 1 shows two antennas occupying volumes alike in shape and size, one being a capacitor (C) and the

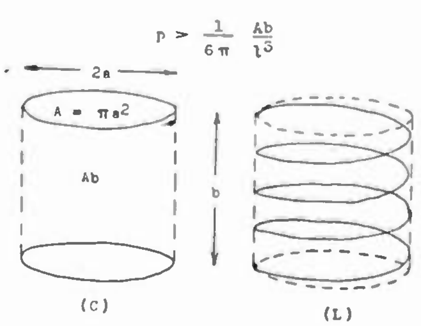

Fig. 1—Capacitor (C) and inductor (L) occupying equal cylindrical volumes.

other an inductor (L). Their maximum dimensions are less than the radianlength of operation. Their shapes are cylindrical because that is the only shape that can alternatively be occupied by either a capacitor or an inductor. The volume may be bounded by a circular cylinder, as shown, or by other cylinders such as square or rectangular.

In both cases the antenna is assumed to operate as a lumped circuit element of the kind indicated (C or L), neglecting distributed properties. The inductor (loop antenna) is assumed to act as a current sheet pervious to alternating magnetic flux, as is customary in the theory of solenoidal coils: this assumption is justified if the coil is wound of several turns of wire or ribbon having a width of about ½ the pitch of winding.

The symbols and principal formulas are tabulated above for convenience. All formulas have the same form for the two kinds of antennas, except for the number of turns n and the correction factors k_a and k_b. These factors are defined to have such values that (1) gives the correct values of C and L.

For the capacitor, the correction factor k_a multiplies the area A to obtain the effective area, as augmented by the electric field outside the cylindrical volume. This factor is greater than unity and (for circular disks, $A=\pi a^2$) greater than

$$\frac{4}{\pi}\frac{b}{a} = 1.27\frac{b}{a}, \tag{9}$$

based on two disks far apart. The value of k_a is asymptotic to unity for $b \ll a$, and is asymptotic to (9) for $b \gg a$.

For the inductor, the correction factor k_b multiplies the axial length to obtain the effective length of the magnetic path as augmented by the external return path. This factor also is greater than unity. If $b>a$ for a circular coil ($A=\pi a^2$), this factor is closely approximated by the asymptotic value:

$$k_b = 1 + \frac{8}{3\pi}\frac{a}{b} \quad \text{or better,} \quad k_b = 1 + 0.9\frac{a}{b}. \tag{10}$$

The effective volume becomes

$$k_b Ab = Ab + 0.9aA = Ab + 2.8a^3. \tag{11}$$

If $b<a$, the factor is somewhat less than this value.

The electric-dipole radiation from the capacitor is represented by shunt conductance G_e or series resistance R_e. The magnetic-dipole radiation from the inductor is represented by series resistance R_m or shunt conductance G_m. In both kinds of antenna the resistance formula is free of the correction factor, because the radiation is caused by the current which is confined to certain definite dimensions of the structure. Therefore, the radiation resistance is the concept ordinarily used. Its value is given not only in the general form but also in the simplified form valid in free space.

The fundamental limitation on the bandwidth and the practical efficiency of a small antenna is the radiation power factor, p_e or p_m, given by (5). It is always much less than unity because of the small size. It has the same value, whether computed from radiation resistance or conductance. It has the same form for both kinds of antennas. Its value, except for the correction factor, is the same for both kinds, and depends only on the ratio of the antenna volume Ab to the radian cube l_r^3.

In (5) the coefficient $1/6\pi$ is the product of the two factors $1/4\pi$ and $2/3$. The former is the reciprocal of the solid angle of a sphere, which appears in rationalized formulas involving spherical waves. The latter is the fraction of the sphere which is filled with the doughnut pattern of radiation characteristic of a small dipole.

As a special case of the radiation power factor, consider an antenna occupying a cubic space Ab equal to a radian cube l^3. The resulting power factor is $1/6\pi = 0.053$, multiplied by the correction factor. In this case, approximately, $k_a=2.7$ and $k_b=1.5$, so the power factors are $p_e=0.14$ and $p_m=0.08$. Therefore, this size of antenna has sufficient radiation damping to operate over a bandwidth of the order of 1/10 the mean frequency, even if there is no other damping.

A cubic antenna of this size (and one turn on the inductor) has a reactance comparable with the wave resistance of the medium ($R=1/G=120\pi=377$ ohms in free space). The reactance ($1/\omega C$ or ωL) of each kind is reduced by the correction factor, so it has a value of 140 ohms for the capacitor, or 250 ohms for the inductor. Reducing the size or the frequency increases the reactance of the capacitor and reduces that of the inductor. The latter has greater flexibility in that its reactance can be increased with the number of turns.

In the cubic shape, the correction factor is slightly greater for the capacitor than for the inductor. This advantage is real, though it is small and may be overbalanced, in some cases, by circuit disadvantages.

If the axis of the cylinder is vertical, either antenna radiates in a pattern like a horizontal doughnut. Since the polarization is expressed with reference to the electric field, the capacitor radiates with vertical polarization and the inductor with horizontal. The required polarization is likely to be the determining factor in choosing which kind to use, if the horizontal doughnut is the desired pattern of radiation.

A plane reflector doubles the radiation power factor if it is located close enough to either kind of antenna and in such relation as to re-enforce the radiation. The plane reflector acts by virtue of its great conductivity or relative permittivity. A surface of water or ground may approximate a plane reflector. The size of the antenna and its proximity to the reflector must be such that the antenna and its image fall within a maximum dimension less than the radianlength, if the radiation power factor of (5) is to be doubled. Also, the reflector must have a radius greater than 1/4 wavelength. To re-enforce the radiation, the plane must be perpendicular to the axis of the capacitor or parallel to the axis of the inductor, so the polarization is perpendicular to the plane.

The cylindrical volume may be filled with a dielectric core in the capacitor or a magnetic core in the inductor. In either case, the radiation shunt conductance (not the series resistance) remains the same, because it is determined by the energy in the field outside of the antenna, regardless of that inside. A dielectric core of relative permittivity k_e increases the capacitance to

$$C = \epsilon \frac{A}{b}(k_a + k_e - 1) \tag{12}$$

approximately if $b<2a$. This reduces the radiation power factor in the ratio

$$\frac{k_a}{k_a + k_e - 1} = \frac{1}{1 + \dfrac{k_e - 1}{k_a}}. \tag{13}$$

A magnetic core of relative permeability k_m increases the inductance to

$$L = \mu n^2 \frac{A}{b(k_b + 1/k_m - 1)} \tag{14}$$

approximately if $b>2a$. This increases the radiation power factor in the ratio

$$\frac{1}{1 - \dfrac{k_m - 1}{k_m k_b}} = \frac{1 + 0.9\dfrac{a}{b}}{\dfrac{1}{k_m} + 0.9\dfrac{a}{b}}. \tag{15}$$

The efficiency may be further increased by reduction in the effective coil resistance.

The structure of the antenna is a subject by itself, outside the scope of this monograph.

The same principles may be applied to the design of a reactor in which radiation is undesired and low power factor ("high Q") is desired. If the reactor is unshielded, the optimum size is a compromise between larger size to reduce internal series resistance and smaller size to reduce internal shunt conductance and radiation. The optimum size for a single-layer coil with negligible dielectric power factor is that for which the radiation power factor is a minor fraction of the total, say between 1/6 and 1/2, depending on the nature of the factors which determine the internal resistance. In ordinary cases, the volume of the coil should not exceed about 1/100 of a radian cube, which means the diameter and length, if equal, should not exceed about 1/5 radian-length, or 1/30 wavelength. If this size is too small, a larger coil with shielding may be required.

V. The Circuits

Efficient operation of a small antenna requires tuning to the operating frequency with a circuit which offers little additional dissipation. How much the circuit may detract from the efficiency depends on the nature of the generator or load coupled therewith, and on other requirements such as bandwidth. The simplest case will be described as an example.

Fig. 2 shows a generator or load coupled with an antenna of either kind (C or L) through its tuner. In the case of a generator, it is assumed that it is so coupled with the tuner as to deliver all of its available power to the tuner and antenna. In the case of a load, it is assumed that it is so coupled with the tuner as to receive the maximum power therefrom, which is called the "available power."

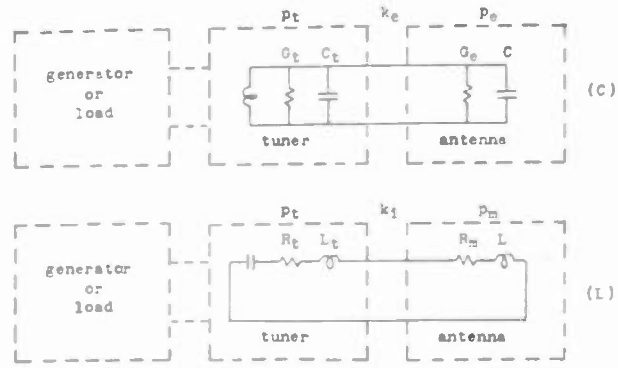

Fig. 2—Tuned coupling of antenna with generator or load.

In general, the efficiency of the coupling circuit is increased by increasing the coefficient of coupling between the tuner and the antenna, and by decreasing the power factor of the tuner.

The coefficient of coupling (k_e or k_i) between the tuner and the antenna is defined in the usual way. Its square is called the "coupling efficiency" because it denotes the fraction of the total electric or magnetic energy of the tuned circuit which is in the antenna. It is expressed in (6) for the simple connection of Fig. 2, but has more general significance.

The power factor p_t of the tuner is taken to include all dissipation in the tuner and antenna, except the desired radiation. In Fig. 2, this is lumped in the effective shunt conductance G_t or the effective series resistance R_t. It is connected directly in parallel or in series with the radiation effective conductance G_e or resistance R_m. In this connection the circuit efficiency is merely the ratio of the radiation power to the total power in the circuit, as expressed in (7).

The more general expression of circuit efficiency is given by (8), in terms of power factor and coupling efficiency. This gives an indication of the relative importance of all factors.

After the coupling circuit has been designed for the maximum efficiency at the frequency of resonance, consistent with available space, materials, and precision, the total power factor of the circuit will exceed $k_e^2 p_e$ or $k_i^2 p_m$ by the amount of the tuner power factor p_t and the added power factor contributed by the generator or load. If the antenna comprises all the reactance of one kind in the circuit, and the tuner losses are small, the total power factor of the circuit may be that of the antenna plus an equal value coupled from the generator or load. Therefore, a very efficient design may have a loaded power factor $2p_e$ or $2p_m$, and a corresponding bandwidth of the tuned circuit.

If the bandwidth desired in the coupling circuit is either less or greater than that obtained by designing for maximum efficiency at the frequency of resonance, the redesign for different bandwidth will be at the expense of efficiency. Lesser bandwidth may be obtained by decreasing the coupling with generator or load, decreasing the coupling between tuner and antenna, multiple tuning, or decreasing the antenna size. Greater bandwidth may be obtained by increasing the coupling with generator or load, increasing the power factor of the tuner, or developing the tuner into a wide-band circuit.

Some types of generator or load do not double the power factor of the tuned circuit when coupled for normal operation. An efficiency generator, for example, operates best into an impedance much different from its internal impedance. A current generator of high resistance, such as a high-μ screen-grid tube, contributes little damping to a tuned output circuit. On the other hand, a voltage generator of low resistance, such as a low-μ triode tube or cathode-output circuit, more than doubles the damping in a tuned output circuit. Likewise, there are load circuits which are essentially voltage-operated, such as a voltmeter or the grid circuit of an amplifier; or current-operated, such as an ammeter. Either type of load may not be designed to utilize the available power, in which case it may add little to the damping. In view of the various effects of the associated circuits, the radiation power factor of the antenna is not the ultimate limitation on the bandwidth of efficient operation, but does indicate the order of magnitude and the trends with changes of antenna design.

VI. Examples

First example: A loop antenna is intended for operation with horizontal axis in a radio receiver cabinet in a small frame building. Its size is 1 meter square by 0.5 meter axial length.

Wavelength:	$\lambda=$	300	m. (at 1 Mc.)
Radianlength:	$l=$	48	m.
Radian cube:	$l^3=$	110,000	m.^{3}
Antenna volume:	$Ab=$	0.5	m.^{3}
Shape factor:	$k_b=$	2	

The radiation power factor is computed by doubling (5) to include approximately the effect of the ground plane.

$$\text{Radiation power factor: } p_m=0.96\times 10^{-6}.$$

The loop is assumed to be one-half the entire inductance of the tuned circuit (6).

$$\text{Coupling efficiency: } k_c^2=0.5.$$

The power factor of the entire tuned circuit is assumed to be 0.01 and the efficiency is computed from (8).

$$\text{Efficiency: } e=0.48\times 10^{-6}/0.01=0.048\times 10^{-3}.$$

This represents a loss of 43 db. It is noted that the essential performance is obtained without reference to incidental factors, such as the number of turns, which are supplied by ordinary design procedure.

A capacitive antenna of comparable volume would give comparable performance, with some practical advantages and disadvantages. Its disuse indicates that the disadvantages usually predominate.

A loop antenna as small as 1/5 the dimensions of this example, namely, $0.2\times 0.2\times 0.1$ meter, is used in small receivers. The efficiency is approximately 0.4×10^{-6}, representing a loss of 64 db at 1 Mc.

Second example: A capacitive antenna over ground is connected with a radio receiver. The antenna is a wire so its area is undefined. Including lead-in, its effective height is 4 meters and its capacitance is 200 micromicrofarads, the I.R.E. standard. Therefore, its effective area is determined by (1).

Antenna capacitance:	$C=$	200 $\mu\mu$fd.
Effective height:	$b=$	4 m.
Effective area:	$k_aA=bC/\epsilon=$	90 m.^{2}
Effective volume:	$k_aAb=$	360 m.^{3}
Wavelength:	$\lambda=$	300 m. (at 1 Mc.)
Radianlength:	$l=$	48 m.
Radian cube:	$l^3=$	110,000 m.^{3}

The radiation power factor over the ground plane is computed by doubling (5).

$$\text{Radiation power factor: } p_e=0.35\times 10^{-3}.$$

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This is a loss of 35 db, chargeable 15 db to circuit dissipation and 20 db to decoupling for reducing the reaction of antenna changes on the tuning. Part of the latter (20 db) can be recovered by greater coupling and providing for retuning on each antenna. Otherwise, it is noted that this antenna is only 8 db better than the loop antenna of the first example.

Third example: A loop antenna is intended for operation with vertical axis in a television receiver cabinet. Its size is 0.5 meter cube. It is tuned to the desired frequency channel.

Wavelength:	$\lambda=5$	m. (at 60 Mc.)
Radianlength:	$l=0.8$	m.
Radian cube:	$l^3=0.51$	m.^{3}
Antenna volume:	$Ab=0.12$	m.^{3}
Shape factor:	$k_b=1.5$	
Radiation power factor:	$p_m=0.019.$	

of the capacitor or parallel to the axis of the inductor, so the polarization is perpendicular to the plane.

The cylindrical volume may be filled with a dielectric core in the capacitor or a magnetic core in the inductor. In either case, the radiation shunt conductance (not the series resistance) remains the same, because it is determined by the energy in the field outside of the antenna, regardless of that inside. A dielectric core of relative permittivity k_e increases the capacitance to

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$$\frac{1}{1 - \dfrac{k_m - 1}{k_m k_b}} = \frac{1 + 0.9\dfrac{a}{b}}{\dfrac{1}{k_m} + 0.9\dfrac{a}{b}}. \tag{15}$$

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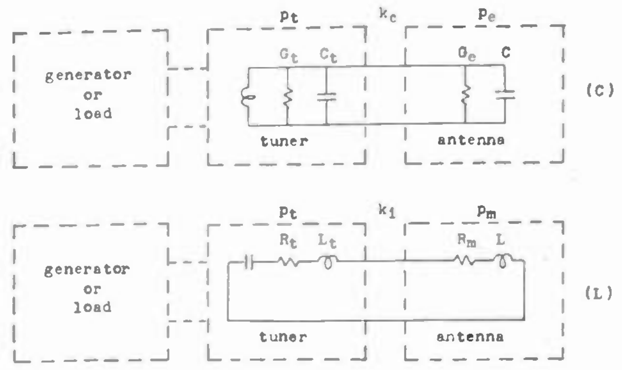

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Antenna volume:	$Ab =$ 0.5	m.^{3}
Shape factor:	$k_b =$ 2	

The radiation power factor is computed by doubling (5) to include approximately the effect of the ground plane.

Radiation power factor: $p_m = 0.96 \times 10^{-6}$.

The loop is assumed to be one-half the entire inductance of the tuned circuit (6).

Coupling efficiency: $k_i^2 = 0.5$.

The power factor of the entire tuned circuit is assumed to be 0.01 and the efficiency is computed from (8).

Efficiency: $e = 0.48 \times 10^{-6}/0.01 = 0.048 \times 10^{-3}$.

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Antenna capacitance:	$C =$	200 $\mu\mu$fd.
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This is a loss of 35 db, chargeable 15 db to circuit dissipation and 20 db to decoupling for reducing the reaction of antenna changes on the tuning. Part of the latter (20 db) can be recovered by greater coupling and providing for retuning on each antenna. Otherwise, it is noted that this antenna is only 8 db better than the loop antenna of the first example.

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Wavelength:	$\lambda = 5$	m. (at 60 Mc.)
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Radian cube:	$l^3 = 0.51$	m.^{3}
Antenna volume:	$Ab = 0.12$	m.^{3}
Shape factor:	$k_b = 1.5$	
Radiation power factor:	$p_m = 0.019$.	

Since the required bandwidth is about 0.1 of the center frequency, or about $5p_m$, there is a loss of only 4 to 7 db, depending on the nature of the circuits connected with the antenna for increasing the bandwidth.

Fourth example: A loop antenna is intended for operation with horizontal axis in a portable f.m. receiver. Its size is 0.2 meter cube. It is tuned to the desired frequency. All losses except radiation and load are assumed to yield a tuner power factor of 0.01.

Wavelength:	$\lambda = 3$	m. (at 100 Mc.)
Radianlength:	$l = 0.48$	m.
Radian cube:	$l^3 = 0.11$	m.^{3}
Antenna volume:	$A_b = 0.008$	m.^{3}
Shape factor:	$k_b = 1.5$	
Radiation power factor:	$p_m = 0.0058$	
Tuner power factor:	$p_t = 0.01$	
Efficiency:	$e = 0.37.$	

This is a circuit loss of 4 db. The bandwidth is 2 or 3 Mc., more than enough for a single channel 0.2 Mc. wide. However, if the same antenna were required to cover the entire band of 88 to 108 Mc. without retuning, a width of 0.2 times the mean frequency, the loss would be 12 to 15 db caused by the wide-band circuit.

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A Helical Antenna for Circular Polarization*

HAROLD A. WHEELER†, FELLOW, I.R.E.

Summary—A helical coil radiates a wave of circular polarization in a doughnut pattern if the area and pitch of the turns are properly related to the radianlength of the wave. For a coil whose dimensions are less than the radianlength, circular polarization requires that the area A of each turn and the pitch p be related to the radianlength l as follows:

$$A = pl.$$

The simplest form of helical antenna is a self-resonant coil of several turns. To obtain greater radiation power factor and efficiency, a multifilar winding is preferred, having a fractional turn for each of several helical wires connected in parallel with symmetry around the axis. This type of antenna offers television the advantages of circular polarization in suppressing echoes from reflecting surfaces.

I. INTRODUCTION

A HELICAL COIL can be designed to radiate waves of circular polarization by properly proportioning the area and pitch of the turns with relation to the wavelength or radianlength. The screw direction of the helix determines the direction of rotation of the wave polarization. The simplest case of this helical antenna is a small one whose dimensions are less than the radianlength.

A small antenna whose dimensions are less than the radianlength behaves essentially as a dipole with a coaxial doughnut pattern of radiation. If it is an electric dipole or current element, the polarization of the electric vector is in the plane of the dipole. If it is a magnetic dipole or current loop, the polarization of the electric vector is normal to the plane of the dipole (the plane through the axis of the loop).

The helical antenna is a superposition of electric and magnetic dipoles to radiate a wave with circular polarization. Reference (9) of the Bibliography shows that a capacitor and an inductor occupying equal cylindrical volumes have approximately the same power factor of radiation. If these are made of equal reactance and connected together to form a circuit resonant at the operating frequency, it follows that the radiated power is about equally divided between the electric-dipole radiation from the capacitor and the magnetic-dipole radiation from the inductor.

A coaxial superposition of the capacitor and inductor is possible if the structure of each is designed to give the required freedom to both fields; there is no inconsistency in their coexistence in the same space.

Circular polarization requires two relations between the crossed fields in a wave. They must have equal intensity and phase quadrature in time. Then the direction of rotation of the polarization depends on the phase sequence of the crossed components of either field.

The helical antenna inherently obtains the phase quadrature. The equality of intensity of the crossed components is obtained by making the area of each turn equal to the product of the pitch of the turn times the radianlength ($1/2\pi$ wavelength). The rotation is determined by the screw direction of the helix. Ideally, there should be no other radiating or reflecting conductors in the vicinity. Capacitive loading at the ends of the coil is permissible, as well as circuit connections

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in the center. These conditions require that the helix operate as a balanced antenna, or unbalanced and connected with a nonradiating, nonreflecting shield.

Efficient operation of a capacitor or inductor as an antenna, as treated in reference (9) of the Bibliography, requires that the radiation power factor exceed the circuit power factor of the antenna and tuner. High efficiency requires that the dimensions of the antenna be comparable with the radianlength. To secure this result in a helical antenna with a length of wire consistent with half-wave resonance, the helix must be made of very few turns, or even a fraction of one turn. For this purpose a multifilar helical winding is employed to maintain the required symmetry about the axis.

The helical antenna opens up some interesting possibilities in diversity reception with small antennas located close together.

The use of transmitting and receiving antennas of the same screw direction offers television some striking advantages in suppressing echoes. It happens that the rotation of polarization is reversed on reflection from a metallic plane surface, so the receiver becomes insensitive to the echo. The degree of suppression which can be obtained in practice remains to be proved by tests. Such a transmitter would also permit the use of either horizontal or vertical receiving antennas if the benefit of echo suppression were not required.

II. Symbols

a = radius of circular cylindrical volume (meters)
A = area of base of cylindrical volume (meters^2)
b = height of cylindrical volume (meters)
$p = b/n$ = pitch per turn of coil (meters)
n = number of turns of coil
k_b = shape factor of inductor = effective length/actual length (b)
ω = radian frequency (radians/second)
$l = 1/2\pi$ wavelength radianlength (meters)
ϵ = electric permittivity in free space (farads/meter)
μ = magnetic permeability in free space (henries/meter)
$R = 120\pi = 377$ = wave resistance in free space (ohms)
E = electric field intensity (volts/meter)
H = magnetic field intensity (amperes/meter)
p_m = radiation power factor of inductor antenna (magnetic dipole)
p_{em} = radiation power factor of helical antenna
I = alternating current (vector) (amperes)
t = time (seconds).

III. The Basic Relations in the Helical Antenna

Fig. 1(a) shows a helical coil of several turns. Fig. 1(b) shows how each turn of the helix may be resolved into two radiating components, one an axial line of length equal to the pitch p, and the other a flat turn of area A normal to the axis. The former radiates as an electric dipole, and the latter as a magnetic dipole. A certain relation between the area and pitch of each turn is re-

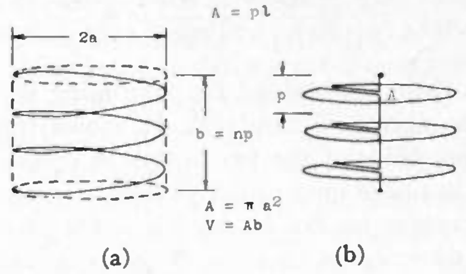

Fig. 1—The helical antenna (a) and the components of a single turn (b).

quired to give the equality of the crossed fields which characterizes circular polarization. However, the relative phase of the crossed fields is the first question.

The radiation of each type is proportional to the dipole moment. The moment of the electric dipole is equal to the product of length times charge at the poles. For an alternating current of vector amplitude I at radian frequency ω flowing in a short conductor of length p, the electric moment is

$$p \int I dt = \frac{Ip}{j\omega} \text{ ampere-second-meters} \tag{1}$$

where dt is the differential of the time t, and $j\omega$ is the differential operator. The moment of a magnetic dipole is similarly defined. The corresponding magnetic moment of a loop is equal to the product of current times area times magnetic permeability. For the same current as above, flowing around a loop of area A in a medium of magnetic permeability μ, the magnetic moment is

$$\mu I A \text{ volt-second-meters.} \tag{2}$$

The resulting radiation fields at the same distance from both dipoles have a ratio determined by the ratio of dipole moments.

The ratio of the radiation flux components parallel to the axis of the dipoles is

$$\frac{j\omega\mu H_m}{j\omega\epsilon E_e} = \frac{\mu I A}{Ip/j\omega} \frac{\text{volts}}{\text{amperes}} \tag{3}$$

where E_e is the axial component of electric field intensity from the electric dipole, H_m is the axial component of magnetic field intensity from the magnetic dipole, and ϵ is the electric permittivity of the medium. These ratios are expressed in such a way as to show the identity of units. In each wave the electric and magnetic intensities are inherently related by the wave impedance R of the medium, as follows:

$$\frac{E_e}{H_e} = \frac{E_m}{H_m} = R = \sqrt{\frac{\mu}{\epsilon}} \frac{\text{volts}}{\text{amperes}} \tag{4}$$

where H_e and E_m are, respectively, the magnetic intensity from the electric dipole and the electric intensity from the magnetic dipole. The ratio of intensities of corresponding fields may be expressed:

$$\frac{E_m}{E_e} = \frac{H_m}{H_e} = R\frac{H_m}{E_e} = \frac{j\omega\mu A}{Rp}. \qquad (5)$$

The last ratio is obtained by combining with the preceding equations (3) and (4). It shows that the corresponding fields of the two waves of crossed polarization are in phase quadrature as indicated by the differential operator $j\omega$. No attempt is made here to formulate the direction of rotation of the polarization.

The condition for equality of crossed components is a structural relation best expressed in terms of the radianlength l rather than the angular frequency:

$$\omega = \frac{1}{l\sqrt{\epsilon\mu}}. \qquad (6)$$

Substituting in (5) for unity ratio of corresponding components,

$$A = p\frac{R}{\omega\mu} = pl. \qquad (7)$$

This is the basic relationship for a small helix equivalent to superposed coaxial electric and magnetic dipoles. The area of each turn is equal to the product of its pitch times the radianlength.

The relation (7) can be derived more simply from the condition for equal field intensities from a current element I_p and a current loop I_A, which gives

$$\frac{A}{l^2} = \frac{p}{l}; \qquad A = pl. \qquad (8)$$

Also, the phase difference of the crossed fields can be deduced from this simple relation because the ratio of the two fields involves the first power of the radianlength or frequency, inevitably associated with phase quadrature.

The above relation is based on the assumption that all the radiation comes from the helix, and none from other conductors. This places a severe restriction on the associated connections. To avoid unbalanced current in the connecting leads, these leads are preferably balanced lines connected or coupled with a balanced helical antenna at its center. While a self-resonant helix is the simplest example of this type of antenna, it is permissible to add capacitive loading at each end in the form of radial spokes connected at a central hub. Such spokes contribute a negligible amount of radiation.

While circular polarization can be obtained from independent dipole and loop structures, connected together as in Fig. 1(b), for example, the helical antenna seems to make the best use of a limited space. The entire conductor contributes to both of the crossed components of radiation.

The diameter of the conductor can be enlarged to a substantial fraction of the space between conductors, so the energy in the space close to the wires and also the wire resistance are reduced in the interest of increasing the radiation power factor and the efficiency. Some economy of space may be obtained by curtailing the ends of the helix, which contribute little to the radiation, and replacing them by capacitive loading in the form of radial spokes.

Two helical antennas of like screw direction operate together as transmitter and receiver by circular polarization. Other helical antennas of the same screw direction may be used as reflectors and directors, either driven or parasitic, because they both receive and transmit the circular polarization with the same screw direction. In view of the circular polarization, the transmission between the two antennas depends only on their patterns of directivity, and not on their orientation about their common line of centers. Interference from reflected signals is greatly reduced because their polarization is reversed by the mirror image of the helical antenna. There is no advantage, however, against diffraction loss around the earth, or against double reflection which restores the same polarization.

A helical antenna of opposite screw direction is theoretically uncoupled, although there is a residual coupling caused by imperfections. The discrimination between adjacent channels in the frequency spectrum would be improved by assigning reverse circular polarization on adjacent channels.

If it is desired to discriminate in favor of reflected signals, as in radar, the use of reverse circular polarization in the transmitter relative to the receiver would accomplish this result.

IV. The Helical Antenna of Several Turns

The assumption of uniform current around each turn of the helix requires that each turn be much less than the resonant length of wire (approximately $\frac{1}{2}$ wavelength). In other words, a self-resonant helix must be small enough so that the resonant length of wire is wound in a substantial number of turns. This means a coil radius which is a small fraction of the radianlength.

As a rough approximation to a self-resonant helix, a half-wave wire is wound in accordance with the critical relation of (8) to radiate circular polarization. The length of wire is

$$\pi l = 2\pi an = 2\pi ab/p \qquad (9)$$

where a is the coil radius and n is the number of turns. The critical relation for circular polarization is, from (8) and (9),

$$pl = \pi a^2 = 2ab; \qquad n = b/p = l/2a. \qquad (10)$$

The dimensions of the coil cylinder become

$$a = \frac{2}{\pi}b = \frac{l}{2n}; \qquad b = \frac{\pi}{2}a = \frac{\pi l}{4n}. \qquad (11)$$

Therefore, a self-resonant helix of this shape inherently gives approximately circular polarization.

The radiation power factor of a self-resonant helix may be deduced roughly from (5) in reference (9) of the Bibliography for lumped circuits. It is defined as the ratio of radiation resistance over reactance. The self-resonant helix has an average current about $2/\pi$ times the maximum current in the center. Therefore, its length of wire is about $2/\pi$ effective, so the active length of the coil is about equal to the radius. The shape factor, k_b in reference (9), is about 2, so the effective volume is about $\pi a^2 \times 2a = 2\pi a^3$. The total radiation power factor caused by both electric and magnetic radiation is obtained approximately from formula (5) in reference (9):

$$p_{em} = 2p_m = \frac{2}{6\pi}\,\frac{2\pi a^3}{l^3} = \frac{2}{3}\left(\frac{a}{l}\right)^3 = \frac{1}{12n^3}. \qquad (12)$$

Therefore, a self-resonant helix of several turns has a very small power factor of radiation and resulting low efficiency. The tabulation below gives the power factor for various numbers of turns:

n	p_{em}
2	0.010
3	0.0031
4	0.0013
5	0.00067
10	0.000083
20	0.000010.

(13)

In the preceding formula, it is assumed that the wire diameter is a substantial fraction of the pitch of winding, as is customary in helical coils. A multifilar winding may be preferable in some cases, as will be described further in the next section. Usually the wire diameter should not be much over half the pitch, especially in the case of very few turns.

V. The Helical Antenna of Fractional Turns

To obtain a fairly large radiation power factor and resulting high efficiency in a helical antenna, it is necessary to use effectively only one turn or less. The maintenance of axial symmetry of current distribution then requires a multifilar winding of several wires connected in parallel at the poles of the coil.

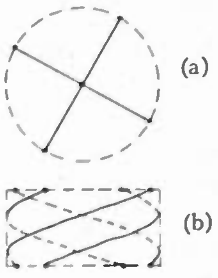

Fig. 2—The multifilar helical antenna of fractional turns.

Fig. 2 shows an example of such a multifilar winding having effectively $\frac{1}{2}$ turn; (a) shows the spokes and hub which joins all the wires at each pole of the coil, and (b) shows the four helical wires of $\frac{1}{2}$ turn each. While the wires are shown as lines for clarity, they should have a diameter which is a substantial fraction of their separation but not so great as to obstruct the magnetic field or to cause excessive capacitance.

The multifilar construction requires the radial spokes at each end to connect the wires in parallel. These spokes do not radiate in appreciable amount, but do contribute capacitive loading at the ends of the coil.

A self-resonant helix along the lines of Fig. 2 would have approximately one-half wavelength of wire in each of the parallel branches from pole to pole. There is an optimum fractional number of turns and a corresponding optimum shape for the self-resonant case to obtain maximum "radiation power factor." The optimum number of turns is probably in the range between $\frac{1}{4}$ and 1 turn, and the optimum shape probably has an axial length about equal to the radius, so Fig. 2 is drawn near the optimum proportions.

The number of wires in parallel, for n turns effective, is preferably at least $2/n$. For example, $\frac{1}{2}$ turn should have 4 wires as in Fig. 2, or more if practical. In general, the more the better.

The same basic relation (8) applies to the multifilar helix of fractional turns, but the pitch becomes greater than the axial length of the coil. The simplified formulas (9) to (12) fail because a substantial part of the length of wire is used in the end spokes and an added length over the circumference of the cylinder. Also, the diameter may slightly exceed the radianlength, so the assumed equivalence of small dipoles is inadequate. For the self-resonant half-turn helical antenna of Fig. 2, the radius and the axial length for circular polarization are roughly 0.6 the radianlength.

VI. Circuit Connections

The connections between a helical antenna and a transmitter or receiver are chosen to preserve the radiation characteristics of the helix free of any other contributions to the radiation. This requires maintaining the balance of the two ends, or some expedient for avoiding radiation from unbalanced currents in associated circuits.

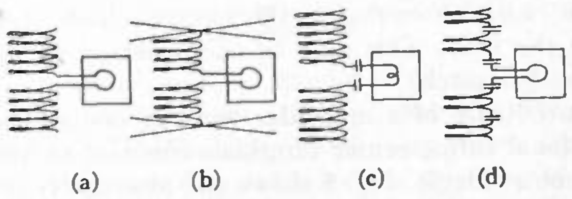

Fig. 3—Connections with a near-by set.

Fig. 3 shows several balanced connections between a helical antenna and a near-by set. They all involve short leads from two terminals in the center of the helix. A self-resonant helix (a) is connected to a small coupling coil for coupling a tuned circuit in the set. The helix (b) is similar, except that it is shorter and has capacitive loading at the ends. The helix (c) is somewhat longer than the resonant length, and the excess reactance is tuned out by series capacitors in the set. The helix (d)is still longer and its reactance is tuned out at intervals by series capacitors; these are separated by a length of wire of about $\frac{1}{4}$ wavelength or less, so the current in between is nearly uniform.

Fig. 4 shows several arrangements for connecting a helical antenna with a set through a transmission line. The first example (a) is an unbalanced connection which

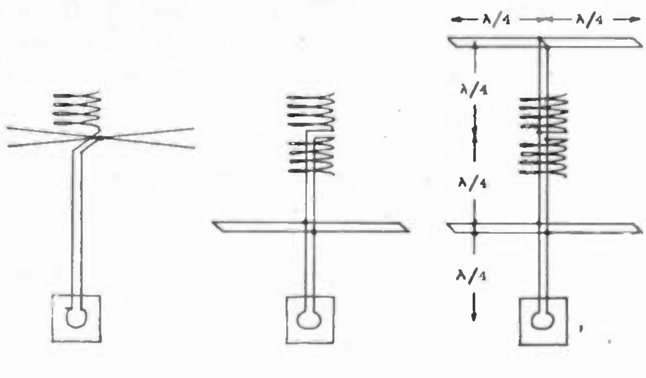

Fig. 4—Connections through a transmission line.

is well adapted for a low-impedance line connected with a coil self-resonant over a shield made of several quarter-wave radial spokes. It is especially useful for a helix of fractional turns, which has a size which is a substantial fraction of the size of the radial spokes.

Figs. 4(b) and (c) show examples of a helix located at a distance from the set and connected therewith through a half-wave line. The stubs in the center of the line do not affect the desired balanced currents in the line, but do prevent resonance toward unbalanced currents. The simpler form (a) disturbs the symmetry in the vicinity of the helix, so the symmetrical form (b) may be preferred. In either case, the capacitive currents between the coil and the line are unbalanced on the line and cause some radiation opposing the electric-dipole radiation from the helix. This may be compensated by a slight increase in pitch.

In the case of a multifilar winding of few turns or fractional turns, center terminals common to all wires are not available. Fig. 5 shows two alternative connections, the helix being shown unwound in the diagrams. In one arrangement (Fig. 5(a)) the leads are connected in only one of the parallel wires, and therefore carry only a fraction of the current. The radiation resistance at these terminals is much greater than that presented to the entire current; 16 times as great in the case of 4 parallel wires. In the other arrangement (Fig. 5(b)) the leads are tapped on one of the parallel wires at points far enough from the center to present the desired impedance. Series capacitors are inserted in the leads if needed to cancel the reactance component of the impedance between the tapping points. Either of these two arrangements may be adapted to match a long transmission line.

Fig. 5—Connections with a multifilar helical antenna.

One form of diversity reception may be obtained by two antennas responsive to opposite circular polarization. Two helical antennas of opposite screw directions may be used for this purpose, as shown in Fig. 6. The

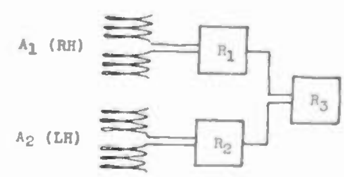

Fig. 6—Diversity receiver using helical antennas of opposite circular polarization.

antennas A_1 and A_2 have, respectively, right-hand and left-hand winding. They are connected to separate receivers R_1 and R_2, then to a combining receiver R_3 containing the mixing and switching circuits which are customary in diversity reception.

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An Adjustable Wave-Guide Phase Changer*

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Summary—A very interesting and useful component of the wave-guide art is the differential phase-shift section, wherein dominant waves of one polarization are caused to travel through a section of wave guide at a different velocity than waves polarized at right angles to the first. Particularly useful are the Δ90-degree and Δ180-degree differential phase-shift sections which produce differential delays between the two polarizations of 90 degrees and 180 degrees, respectively. The properties of these sections are discussed, and it is shown how they may be combined to form a phase changer which will transmit substantially 100 per cent of the incident power with a phase which is readily adjustable. Several different methods of building these sections are finally described.

Introduction

A CONTINUOUSLY adjustable phase changer, by means of which the phase of an output wave may be shifted with respect to the input, is a very convenient and often necessary component in the radio-frequency art. A number of types have been used in the past, and these have usually taken one of two forms: a network of lumped-circuit elements in which the phase change is obtained by varying the magnitude of certain of the elements, or a rotary capacitor in which a pickup plate is caused to rotate over a set of stationary plates which are driven by quadrature voltages in such a way that a rotating electric field is set up thereby. In the first case, the lumped circuits do not allow a continuous adjustment of phase from zero through 360 degrees. The rotary capacitor usually does. However, both types have the severe disadvantage that they are primarily voltage devices. That is, they are composed of high-impedance circuits, and the output power taken by the load must be kept very small so as not to disturb the phase relations in the circuit, or the field in the capacitor.

The following will describe a new form of phase changer which has proven very useful and which is primarily adapted for wave-guide applications. It provides continuous and cumulative shift of phase by means of a rotary adjustment. This is a rather important feature, since a reciprocating adjustment would practically rule out any high-speed applications. But perhaps the most outstanding and unique property of this device is that it is capable of transmitting with arbitrarily variable phase substantially 100 per cent of the power available from the source. This is a very useful property in the microwave range where power is precious and cannot easily be regained by amplification. Furthermore, it is capable of handling power of the order of several hundred kilowatts.

This device is made possible by the fact that in a wave guide of circular or square cross section it is possible to have two traveling waves of the dominant type which have their electric axes at right angles to one another and are therefore independent. It is well known that the dominant transverse electric wave in a circular wave guide has an electric field pattern in any particular cross section of the wave guide as shown in Fig. 1. As the wave travels past this cross section, the orientation and shape of the field contours remain fixed, although the magnitude of the field will vary. Consequently, such a wave may be characterized by a certain direction of polarization indicated by the vector E. We may then say that the wave shown in Fig. 1 is vertically polarized because the electric field component passing through the center of the cross section of the wave guide is oriented vertically. We might have

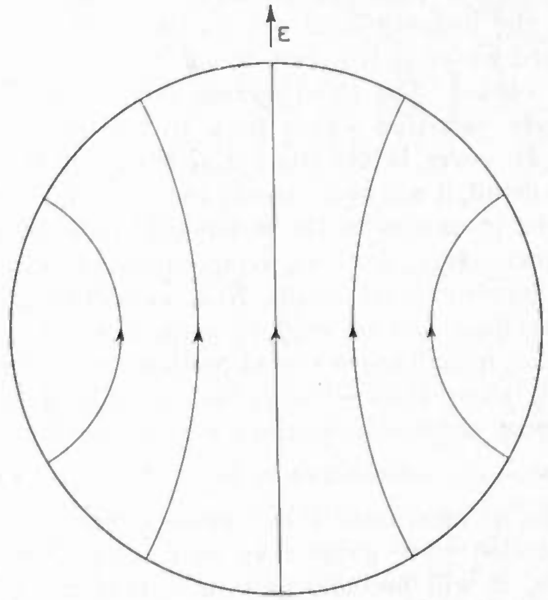

Fig. 1—Field pattern and polarization of dominant wave in circular guide.

an infinite number of other dominant waves of the same type which are polarized at all possible angles. However, it can be shown that any of these waves may be resolved into two dominant waves, one of which is polarized vertically, and the other of which is polarized horizontally. If a linearly polarized receiver is so arranged as to absorb vertically polarized dominant waves, the presence of horizontally polarized waves will not be detected, and we may therefore conclude that the wave guide can be used as if it constituted two independent transmission lines, one for vertically polarized waves, and the other for horizontally polarized waves. Since a wave at any other angle has components in the vertical and horizontal directions, the number of inde-

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pendent transmission lines available is limited to two.

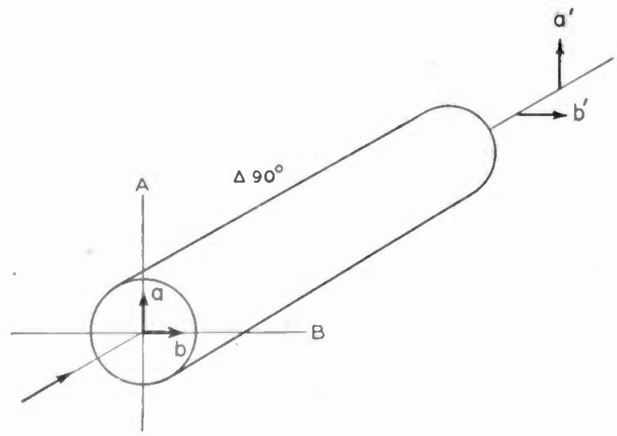
It has also been shown that the phase velocity of waves in a metal-tube guide is greater than the velocity of light in free space, and this phase velocity is dependent upon the physical dimensions of the tube. It is, therefore, possible to control the phase velocity by suitably designing the wave guide section. Furthermore, it is possible to build a section of wave guide which will produce two different phase velocities depending upon whether waves are polarized parallel with or perpendicular to a certain axis.

A section of wave guide having this property, namely, the ability to transmit two sets of waves polarized at right angles to one another with different speeds, will, of course, produce two different phase delays for the two polarizations, and accordingly will be called a "differential phase-shift section." Such sections have a number of interesting and very useful properties, both independently and in combination.

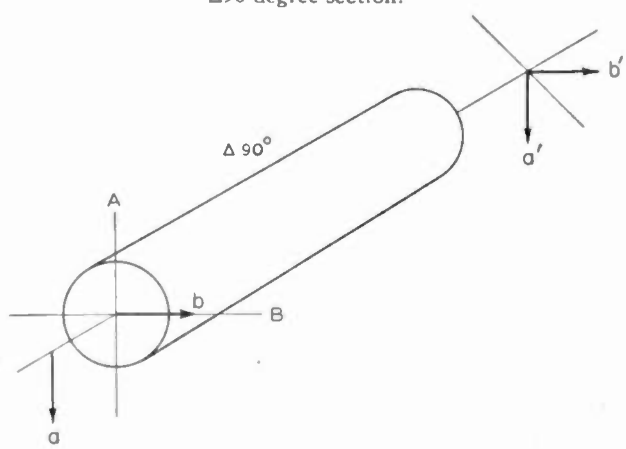
One such combination is the phase changer mentioned above which comprises three differential phase-shift sections of wave guide assembled in tandem. The first of these converts incident linearly polarized waves into circularly polarized waves. The second serves to rotate the instantaneous orientation of the circularly polarized waves as required, thereby shifting the phase of the output. The third section then reconverts the circularly polarized waves back to linearly polarized waves. However, before attempting to explain the operation in detail, it will be necessary to familiarize ourselves with the properties of the differential phase-shift sections themselves, deferring temporarily any discussion of the constructional details. Next we will explain how these sections can be used to perform several useful functions, including the one of producing an arbitrarily variable phase shift. Finally, the methods of building differential phase-shift sections will be described.

Δ90-Degree Section

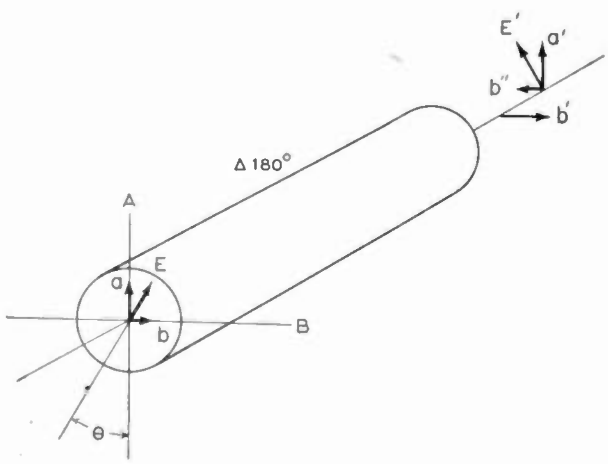
While, as mentioned above, either circular or square cross-section wave guide may be employed for these sections, it will become clear later that the circular cross section is in general preferred, and we will restrict our attention principally to this form. Suppose then that we choose a length of circular wave guide of suitable dimensions to transmit the dominant transverse electric wave for the frequency in which we are interested. And suppose that we equip the section with elements, to be described later, so that waves polarized parallel to axis A (Fig. 2(a)) travel faster[1] than those polarized parallel to axis B, which is at right angles to A. This is indicated schematically by showing diametral electric vectors a and b corresponding to adjacent voltage maxima for two waves polarized parallel to the axes A and B, respectively, and entering the section from the left at the same instant. These vectors are convenient tags which we have hung on the two waves at significant points, and by following them through the phase-shift sections we may observe the effects upon the waves as a whole. At the right these two vectors are shown emerging displaced from one another, a' having traveled a greater distance than b' by virtue of its greater phase velocity. For convenience the section is shown alone in space, but it should be understood that the waves are conducted into and out of the section by means of adjoining wave guides hav-



(a)
Fig. 2(a)—Conversion from linear to circular polarization by Δ90-degree section.



(b)
Fig. 2(b)—Conversion from circular to linear polarization by Δ90-degree section.



(c)
Fig. 2(c)—Rotation of polarization by means of Δ180-degree section.

[1] All mention of speed refers to *phase velocity*.

ng the same cross section. If, now, the properties of he section are adjusted so that a' precedes b' by one-quarter wavelength, the differential phase shift will be 90 degrees and the section will be denoted by the symbol Δ90 degrees. It should be noted that this phase differential bears no direct relation to the absolute phase delay, which does not concern us here, but is the difference between the two absolute phase delays.

Let us now examine the properties of the emerging wave as seen at some particular cross section to the right. First the wave will appear to have an instantaneous electric vector a' which points upward. Ninety degrees later in time the field pattern will have moved forward by one-quarter wavelength, and the electric vector b' will point to the right. One hundred and eighty degrees later the vector will point downward. Two hundred and seventy degrees later the vector will point to the left. Thus we may say that these two emerging waves form a circularly polarized[2] wave which rotates clockwise looking in the direction of propagation. Similarly, the two in-phase waves entering at the left, when added together vectorially, may be considered to form a linearly polarized wave at an angle of 45 degrees to axes A and B. Or, conversely, we might say that the two waves a and b are components of a linearly polarized wave oriented at 45 degrees between the axes. Thus, we may conclude that a Δ90-degree section has the property of converting a linearly polarized wave into a circularly polarized wave, provided that the input is oriented at 45 degrees to the principal axes A and B.

Of course, there are two orientations for the input polarization which will be at 45 degrees to the principal axes. For example, if the B axis wave had been phased so that in Fig. 2(a) the vector b pointed toward the left, the input polarization would have again been at 45 degrees to the principal axes, but this time it would be perpendicular to the original orientation. In this case the emerging circularly polarized wave would rotate counterclockwise instead of clockwise. Consequently, we may generalize by saying that, if the input linear polarization is at an angle of 45 degrees clockwise with respect to axis A (the higher speed axis), the circular polarization will rotate clockwise. Conversely, if the input polarization is 45 degrees counterclockwise from axis A, the circular polarization will rotate counterclockwise.

[2] There may be some confusion as to the meaning of the term "circularly polarized wave" when applied to guided waves. The usage adopted here refers not to the shape of the lines of electric or magnetic force commonly denoted by the subscript numbers associated with the wave type, such as $TE_{0,1}$; but to the way the field pattern, changes with time. Thus, in order to be consistent with optical terminology, a "linearly polarized wave" is one whose pattern does not change direction with progression of time but merely varies in amplitude. A "circularly polarized" wave is one whose cross-sectional field pattern rotates in the plane of the cross section as time progresses, and does not change in amplitude. Waves of the circular electric ($TE_{0,n}$) and circular magnetic type cannot be said to have a direction of polarization, and hence the terms "linear polarization" and "circular polarization" are meaningless when applied to such waves.

Next, let us consider what happens if a circularly polarized wave is sent into a Δ90-degree section. Figure 2(b) shows the same section used in Fig. 2(a). Now, however, we are sending a clockwise-rotating circularly polarized wave in from the left. The first two voltage maxima are indicated by the vectors at the left as b and a. Again, the a component travels more rapidly than the b component and catches up with it. a' and b' when added together now form a linearly polarized wave at an angle of 45 degrees counterclockwise from axis A. Similarly, if a counterclockwise-rotating wave is sent into the section from the left, the emerging wave will be linearly polarized at an angle of 45 degrees clockwise from axis A.

Finally, let us take note of the following extremely important fact. The instantaneous phase of the emerging linearly polarized wave is going to depend upon two things. First, it will depend upon the time of transmission through the differential phase-shift section. Second, it will depend upon the instantaneous phase of the input wave. But the instantaneous phase of the input circularly polarized wave depends upon, and is synonymous with, its instantaneous polarization or orientation in the input plane. Thus the *time phase of the output* depends upon the *spatial orientation of the input*. Consequently, if we can devise some means for controlling the instantaneous orientation of the input wave, we will have the means for adjusting the time phase of the output. As we shall see, a Δ180-degree section will give us this control.

Δ180-Degree Section

Let us now assume that we can build a section of circular wave guide which will produce a differential phase shift of 180 degrees. Fig. 2(c) shows such a section. Linearly polarized waves represented by vector E are being introduced from the left, and these are polarized at an angle θ clockwise from axis A. Vector E may be resolved into components a and b along axes A and B, as shown. Again the A-axis component travels at higher speed than the B-axis component, with the result that upon emerging from the other end of the section, b' lags behind a' by 180 degrees or one-half wavelength. Hence, at the position of a' the B-axis component will be pointing in the opposite direction from b', as indicated by b''. Now, when a' and b'' are added together vectorially, the resultant will be a linearly polarized wave represented by E' polarized at an angle θ counterclockwise from the A axis. We may conclude, then, that the effect of a Δ180-degree section upon linearly polarized waves is to cause a rotation of the angle of polarization in the direction of the A axis by 2θ, or twice the angle between the A axis and the input polarization. (The B axis could equally well have been chosen as the reference axis, and the same result would have been obtained.) If the input polarization remains fixed, rotation of the Δ180-degree section by angle θ will cause a rotation of the plane of the output polarization by twice θ. One-half turn of the section will cause the output vector to swing through a complete circle and return to its original position.

If, instead of a linearly polarized input, we should apply a clockwise-rotating circularly polarized wave, we may deduce the results in exactly the same way as before. Or we may think of the circularly polarized wave as a linearly polarized vector which, however, is rotating in the clockwise direction. Since the angle between this input vector and axis A is constantly increasing in the clockwise direction, we may simply use the conclusions of the preceding paragraph to show that the angle of the output vector is constantly increasing in the counterclockwise direction. It is, therefore, an interesting property of the $\Delta 180$-degree section that it converts clockwise circularly polarized waves into counterclockwise circularly polarized waves. The significant point is, however, that even for circularly polarized waves, if we examine the field patterns existing at a particular instant in time, the instantaneous angle of the output vector will depend upon the instantaneous angle of the input vector *with respect to the principal axes of the section*. Therefore, by rotating the section the instantaneous output polarization can be rotated. This is just the property we need to make up a complete phase changer.

Adjustable Phase Changer

Fig. 3 shows the essential parts of a complete wave-guide phase changer. It is usual microwave practice to work with linearly polarized waves in rectangular wave-guide. Consequently, some suitable transition such as a taper section should be employed to pass from rectangular to circular wave guide. The waves are still linearly polarized, however, and are denoted by the vector E at the left of the drawing. Our first job is to convert these linearly polarized waves into circularly polarized waves, and accordingly they are passed through a

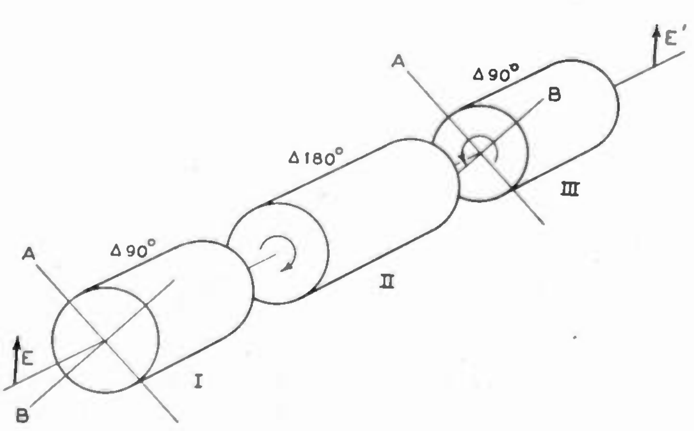


Fig. 3—Adjustable phase changer.

$\Delta 90$-degree section (I) whose principal axes are oriented at 45 degrees to the input polarization, as shown. The emerging waves are now circularly polarized and rotate in the clockwise direction. Next they pass through a $\Delta 180$-degree section (II) which is mounted in bearings so that it is free to rotate. The resulting counterclockwise-rotating waves finally encounter a second $\Delta 90$-degree (III) section which performs the task of converting them back into linearly polarized waves, which may then be handled as required. The output polarization may be oriented at any angle required merely by setting the final $\Delta 90$-degree section at the proper angle. If the output polarization is desired in the same plane as the input, then the axes of the second $\Delta 90$-degree section will be parallel with the corresponding axes of the first, as shown in Fig. 3.

As was pointed out before, the instantaneous time phase of the output waves will depend upon the instantaneous orientation of the circularly polarized wave at the input of section III. And this instantaneous orientation is under our control by means of section II. Consequently, rotation of the $\Delta 180$-degree section will cause a change in the time phase of the output waves. Because one-half revolution of section II produces 360 degrees rotation of its output vector, it follows that one-half revolution of this section produces 360 degrees change of time phase. The sense in which the time phase is changed may be determined as follows. The input to section III is a vector rotating counterclockwise. At the present instant this vector has a particular position. At some future instant the vector will lie in a new position counterclockwise from the present position. Consequently, if by rotating section II counterclockwise we cause the present vector to assume a new position which would normally have been represented by this future epoch, we have advanced the phase of the wave in time. Conversely, rotating section II clockwise would retard the phase of the emerging wave. In general, then, rotating the 180-degree section in the same direction as the rotation of the wave entering section III will cause an advance in phase. All of these conclusions may be verified mathematically, as demonstrated in the Appendix to this paper.

We see, therefore, that the assembly shown in Fig. 3 constitutes a complete adjustable phase changer. Because we have assumed that the individual sections of which it is composed do not cause any appreciable attenuation, substantially 100 per cent of the incident power will be transmitted with altered phase. There is no limit on the range of phase control, and continuous rotation of the $\Delta 180$-degree section will cause continuous retardation or advancement of the phase. This also means that continuous rotation of the $\Delta 180$-degree section at constant speed will cause a fixed increase or decrease in the frequency of the transmitted waves. Furthermore, waves passing through the assembly will suffer the same phase shift regardless of the direction of transmission, and rotation of the $\Delta 180$-degree section will produce the same change in phase, both in amount and in sense, for either direction. Thus, the phase-changer assembly is the equivalent of an elastic piece of transmission line which is capable of being arbitrarily stretched or compressed to any desired length. This means that if we transmit waves through the phase changer toward a mismatched termination, the reflected waves which are seen returning toward the source will have passed through the phase changer twice and will accordingly suffer twice the phase shift of a single traversal. Rotation of the Δ180-degree section through θ degrees will cause 4θ degrees change in phase of the the reflected wave, and consequently the input impedance seen looking into the phase changer in the direction of the load will vary as the Δ180-degree section is rotated.

Such a phase changer has actually been built, and performs as predicted. With properly built and adjusted components, the linearity of phase change versus rotation of the Δ180-degree section is excellent, and is limited only by the accuracy with which the components are built. One particular application in which it has proven particularly useful is in the construction of a MUSA[3] type of antenna for a main-battery fire-control radar for the Navy. Such an antenna may consist of a broadside array of radiators each of which is separately fed as in Fig. 4. If all of the radiators are in phase, a very directional lobe is radiated at right angles to the line of the array. If, however, the phases of the radiators are progressively varied from one end of the antenna to the other, the radiated lobe will assume some

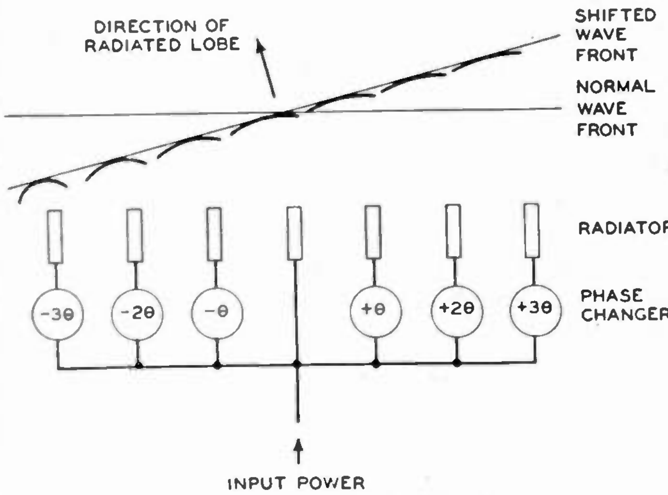


Fig. 4—Parallel-fed MUSA.

other angle in accordance with angle of the new composite wave front. By continuously varying the phases of the individual elements, the radiated beam can be caused to scan a sector of the horizon. An important point is that in older MUSA schemes the fact that each phase changer entailed a considerable loss of power, which had to be made up again with amplification, dictated that the individual radiators be fed through separate phase changers so that none of the radiated or received power had to traverse more than a single phase changer. This is indicated schematically in Fig. 4. Since it is also true that the change in phase for each radiator must be proportional to its distance from the center of the array, the phase changers feeding the end elements of such an array must be rotated at much higher speeds than those on either side of center. With the wave-guide type of phase changer this is no longer necessary. Because their transmission loss is negligible, these phase changers may be arrayed in a series rather than a parallel arrangement, as shown schematically in Fig. 5. Between the phase changers are junctions from which appropriate amounts of power are bled off and fed to the corresponding radiators. Power delivered to the nth radiator away from center will have passed through n phase changers, and so will have accumulated a total phase shift equal to the sum of the individual phase shifts. Consequently, all of the phase changers may be ganged together and rotated at the same speed—equal to that of the slowest one of Fig. 4—and the result will be exactly as desired, with the radiated phase displacements being proportional to the distance of the element from the center of the array.

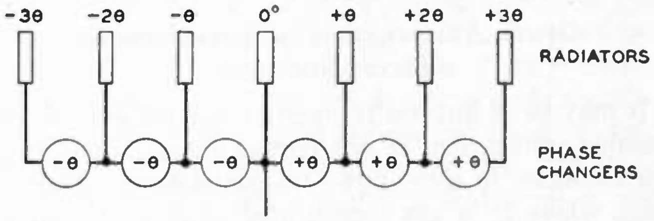


Fig. 5—Series-fed MUSA.

As a corollary to the above it follows that, if we want to produce a more rapid change in phase for the same rotary speed of the Δ180-degree section, we may do so simply by cascading a number of complete phase changers and ganging their rotors together. However, for a single load where it is unnecessary to tap off power at several points as in the MUSA antenna, the cascaded structure may be greatly simplified. This is because the conversion from linear to circular polarization is being unnecessarily duplicated. The rear Δ90-degree section of one changer is converting from circular to linear polarization, while the front section of the following changer is converting back to circular polarization again. Consequently, it is possible to drop out all intermediate Δ90-degree sections, leaving us with a final assembly comprising one Δ90-degree section at either end with a number of Δ180-degree sections between them. Alternate Δ180-degree sections must now be rotated in opposite directions in order to obtain the cumulative addition of phase.

The frequency band over which one of these phase changers will operate satisfactorily is, of course, dependent upon the operating band of the differential phase-shift sections. In general they work perfectly only at one frequency. Departure from midband frequency will cause some reflection of power and/or failure of the individual sections to maintain the required differential phase shift. For the phase changer as a whole, either or both of these effects will result in (1) slight loss in linearity of phase shift versus rotor position, (2) slight variation in the input impedance as the

[3] H. T. Friis and C. B. Feldman, "A multiple unit steerable antenna for short wave reception" Proc. I.R.E. vol. 25, pp. 841–917; July, 1937. The systems described herein are broadside rather than end-fire arrays, and differ in a number of details from those of the above paper.

rotor is turned, and (3) the development of a small component of field at right angles to the output polarization. This last error is the most troublesome of the three. Since the output is generally delivered to rectangular wave guide which cannot possibly transmit any cross-polarization, this spurious cross-component will be reflected back into the phase changer, and will be subsequently re-reflected from the input end for the same reason. The net result is that for certain settings of the rotor this multiply reflected cross-component will resonate and cause very sharp dips in transmitted power and also sharp anomalous phase changes. These spurious effects may be largely eliminated simply by the insertion of a suitable polarized absorber between the ends of the phase changer and the transitions to rectangular wave guide. These are arranged so that they absorb most of the spurious cross-component without affecting waves of the desired output polarization. By this means the phase changer can be made reasonably uncritical of the exact operating frequency.

Other Applications of Differential Phase Sections

It may be of interest to mention briefly several other possible applications of differential phase-shift sections. For example, in some radio transmission or radar systems, where it is not known in advance just what the angle of polarization of the distant receiver or reflector will be, it may be convenient to radiate a circularly polarized wave, because then half of the power would be received regardless of the angle. This circularly polarized wave may be more conveniently obtained simply by passing linearly polarized waves through a Δ90-degree section rather than by splitting the power between two separate transmission lines, delaying one component 90 degrees with respect to the other, and radiating them separately by means of two radiators having mutually perpendicular polarization.

Another problem which frequently arises is the wave-guide rotating joint. A transmitter may deliver microwave power via a wave guide to an antenna which must be free to rotate. It is obvious, however, that the polarization received on the antenna side of the rotating joint will turn as the antenna is rotated, and this must be avoided. One solution is to employ a Δ90-degree section on either side of the rotating joint, as shown in Fig. 6. The one on the transmitter side is oriented so as to convert the linearly polarized wave delivered from the transmitter into a circularly polarized wave. This circularly polarized wave is then transmitted across the joint and is reconverted into a linearly polarized wave at the required angle by the upper section. Since the angle of this final polarization is determined only by the orientation of the upper phase-shift section, and this section turns with the antenna as a unit, it follows that, relative to the antenna, the output polarization is independent of the orientation of the antenna. It might be mentioned in passing that this also is a phase changer, since the phase of the wave delivered to the antenna will depend upon the orientation of the antenna. For this reason the input impedance on the transmitter side of the joint will vary as the antenna is rotated unless the antenna provides a good match for the transmission line.

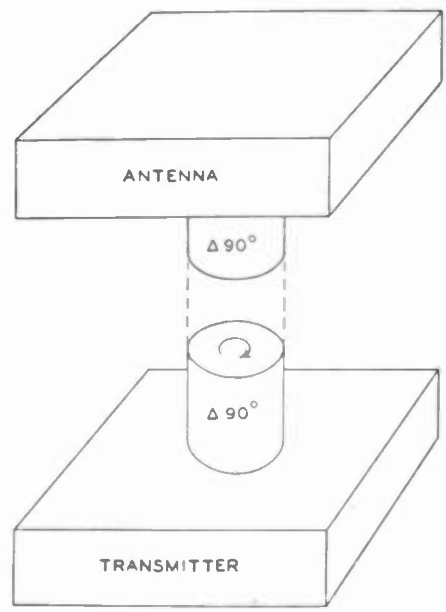


Fig. 6—Rotary joint using circular polarization.

The Δ180-degree section can advantageously be used in making a wave-guide power divider. Suppose, for example, that we have need for some way of dividing power in varying proportions between two separate loads. Incident linearly polarized waves may be first passed through a Δ180-degree section and subsequently into a 120-degree wave-guide Y-junction. Such a junction (Fig. 7) having its three arms symmetrically disposed at equal angles of 120 degrees can be designed with additional elements so that only vertically polarized waves will be transmitted down branch A and only

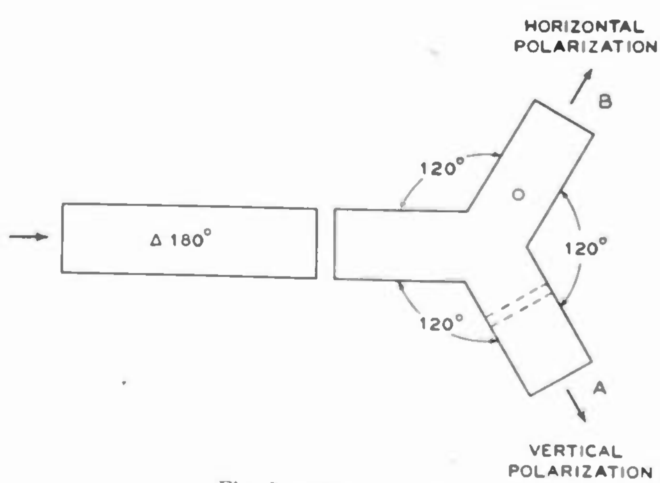


Fig. 7—Power divider.

horizontally polarized waves down branch B. As the Δ180-degree section is rotated, the waves leaving the section can be made to have any desired angle with respect to plane of the junction. They may then be resolved into vertically and horizontally polarized components which will travel down their respective branches o the two loads. All of the power may be sent into ither load, or it may be divided between them in varying proportions. The power delivered to the loads will ary as $\sin^2 2\theta$ or $\cos^2 2\theta$ where θ is the angle between ither of the axes of the differential phase-shift section nd the plane of the junction.

Construction of Differential Phase-Shift Sections

The differential phase-shift sections may take any one of several different forms which will be mentioned here. n general, all forms may be divided into two types; namely, the distributed-parameter type, and the lumped-element type.

Distributed-Parameter Sections

It has probably already occurred to the reader that one way of making a differential phase-shift section is simply to use a section of rectangular or elliptical wave guide. Since the cutoff wavelength will be different for waves polarized parallel to the major and minor axes, their phase velocities will also be different. However, such a section will not fit properly against an adjacent section of circular wave guide.

One very simple way[4] of solving this problem consists of taking a section of circular wave guide and deforming it by judiciously squeezing it in a vise at the middle of the span. In this way, it may be made elliptical in cross section at the middle, and yet be circular at the ends where it must fit other circular sections. When properly done, the transition between the circular ends and the elliptical center is gradual enough that it constitutes a taper transformer which will give a good impedance match between the different cross sections. The number of degrees of differential phase shift will be determined by the amount of flattening produced and the length of guide which has been flattened. To obtain the desired results is a fairly simple procedure experimentally. However, it has the important drawback that it is hard to specify the distortion in such a way that it can be easily reproduced on a manufacturing basis.

Another method,[5] which overcomes the above objection, is to equip an undistorted section of circular wave guide with two diametrally opposed metal fins attached to the walls of the wave guide and extending along the guide axially. This is shown in Fig. 8. Since these fins are fairly thin, they have little effect on waves whose electric field is perpendicular to them. But for waves polarized parallel to the fins, they load the guide with shunt capacitance, thereby not only reducing the characteristic impedance of the section but also decreasing the phase velocity of the waves. In this sense the fins produce very much the same effect as continuously loading the wave guide with a high-dielectric-constant material. Obviously, the phase differential will again depend upon the length of the loaded section and upon the amount of loading, which is determined primarily by the diametral extent of the fins. The notches cut in the ends of the fins are for the purpose of matching the

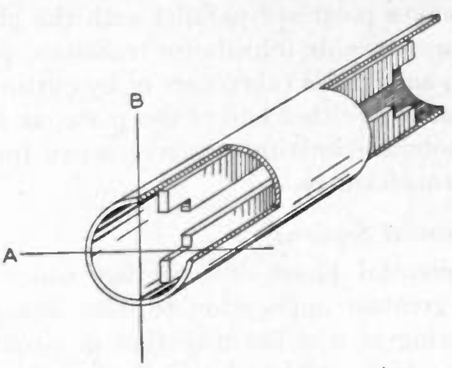


Fig. 8—Fin-type phase-shift section.

loaded-line impedance to the unloaded-line impedance. The section of wave guide containing the step constitutes a loaded section which has effectively the geometric mean impedance between the impedances of the fully loaded and the unloaded guides, and is effectively one-quarter of a guide wavelength long. Since the fins are very simply specified mechanically, it is consequently easy to reproduce such sections. They probably will not stand quite as high power as the other sections because of the rather intense concentration of field around the edges of the fins. However, they are much better in this regard than might at first appear, and tests on suitably designed models at a 3.2-centimeter wavelength indicate a power-handling capability in excess of one hundred kilowatts. In order to obtain such performance it is necessary to round the opposing edges of the fins and make the fins of sufficient thickness so that the radius of curvature of the edges is quite appreciable.

Still another way of producing differential phase shift is to insert a plate of dielectric material in a section of circular wave guide so that it extends across the wave guide diametrally, as shown in Fig. 9. Waves polarized perpendicular to the plate will be slowed up to some extent, but waves polarized parallel to the plate will be slowed up even more. And it is the difference between these two velocities which gives us the differential phase shift which we desire. In general, the use of high-dielectric-constant materials is to be preferred, as this will permit the plate to be made thin enough so that it will affect the waves of transverse polarization very little. This is important, since if they are affected to a slight-

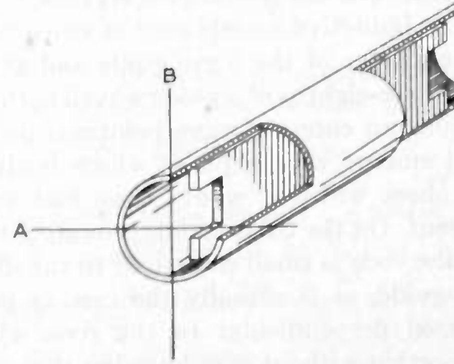


Fig. 9—Dielectric-plate phase-shift section.

[4] This method was developed by W. A. Tyrrell, Radio Research Department, Bell Telephone Laboratories.

[5] This method also was invented by W. A. Tyrrell.

enough extent there is no problem of impedance match into and out of the section for this particular polarization. For waves polarized parallel with the plate there will be an appreciable impedance transition going into the section, and this is taken care of by cutting a quarter-wave notch in either end of the plate, as shown, so that the notches constitute quarter-wave impedance-matching transformers.

Lumped-Element Sections

The differential phase-shift section which has received the greatest application to date is a polarized filter consisting of a uniform section of circular wave guide across which are placed diametral conducting rods at appropriate intervals. For a Δ90-degree section this takes the form shown in Fig. 10(a), the equivalent circuit of which is shown in Fig. 10(b). As indicated, for waves whose electric field is parallel with the rods, the rods behave like inductances shunted across an equivalent transmission line. The susceptance of the rods is

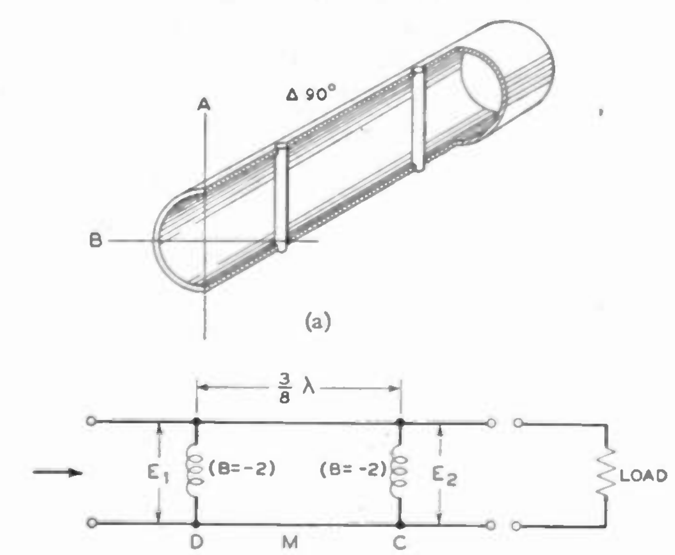


Fig. 10—Rod-type Δ90-degree section. (a) Cut-away view. (b) Equivalent circuit.

roughly proportional to their diameter, and may, therefore, be adjusted to any desired value by choosing the correct diameter. For the Δ90-degree section, both rods should have an inductive susceptance of twice the characteristic admittance of the wave guide and should be separated by three-eighths of a guide wavelength. Under these conditions an entering wave polarized parallel to the rods will emerge with a phase which leads by 90 degrees the phase which it would have had were the rods not present. On the other hand, providing that the diameter of the rods is small compared to the diameter of the wave guide, as is actually the case in practice, waves polarized perpendicular to the rods will pass through the section without ever knowing that the rods are there. Consequently, waves polarized parallel with the rods will receive a differential phase advance of 90 degrees with respect to waves polarized at right angles to the rods, and the A axis will be parallel with the rods.

Because such a section is a type of band-pass filter, complete transmission of power will take place at only one frequency. As the operating frequency departs from the nominal midband frequency, the transmission will begin to fall off and the phase differential will depart from 90 degrees in a manner very similar to the behavior of a parallel-resonant circuit. However, the effective Q of this circuit is quite low, being, in fact,

$$Q = \frac{3}{4}\pi\left(\frac{\lambda_g}{\lambda_a}\right)^2 \tag{1}$$

where λ_g is the guide wavelength, and λ_a is the air wavelength at midband. Consequently, this section is reasonably broadband in its performance. If the length of the section is not an important requirement, the frequency performance may be still further broadened by extending the section and using three or more rods.

When we come to build a Δ180-degree section, it is evident that this may be done by taking two Δ90-degree sections of the type just described and connecting them in tandem with all of the rods parallel. Since each section individually transmits all of the power for either polarization, the two in series must do likewise; and the phase differential must be twice as great as for a single section. Some simplification and greater compactness are obtainable by pushing the two sections together until the adjacent rods occupy the same position. These may then be replaced by a single rod whose susceptance is just twice that of the original rods, and we are left with a three-rod section in which the two end rods have a susceptance of -2 and the middle rod has a susceptance of -4, all of them being separated by $\frac{3}{8}\lambda_g$ spacing (Fig. 11).

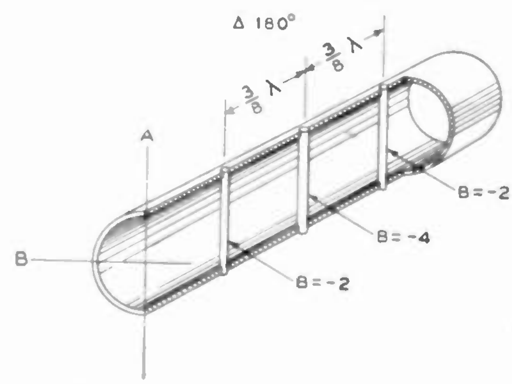


Fig. 11—Rod-type Δ180-degree section.

In the construction of such lumped-element sections we are not restricted to the use of only inductive elements. There is no theoretical reason why capacitive elements might not be used instead. In fact, there might appear to be considerable advantage in the use of such elements, inasmuch as a shorter section should then be possible. Thus, if capacitive elements are used in the form of thick diametral conducting rods cut away at the center so as to leave two opposed cylindrical plugs attached to opposite walls of the wave guide, these may be adjusted both as to diameter and length of gap to give a capacitive susceptance equal to twice the characteristic admittance of the wave guide, and should then theoretically be placed $\frac{1}{8}\lambda_g$ apart in the wave guide, as contrasted with $\frac{3}{8}\lambda_g$ for the inductive elements. Waves polarized parallel with the capacitive elements should now receive a differential *phase delay* of 90 degrees with respect to waves polarized perpendicular to the elements. Actually, however, experience has shown that, when such elements are placed as close together as $\frac{1}{8}\lambda_g$, there is so much mutual coupling between the elements that they do not behave at all as expected, and consequently it is necessary to space the capacitive elements at least $\frac{5}{8}\lambda_g$ apart in order to obtain the expected operation.

In concluding our remarks about the several types of differential phase-shift sections, it may be of interest to make some general comparisons. The lumped-element sections are capable of being made physically shorter than any of the distributed-parameter type, and are therefore to be preferred when compactness is a requirement. On the other hand, their frequency characteristics are appreciably narrower than are those of the distributed-parameter sections. This disadvantage is not necessarily an inherent one. As mentioned earlier, the operating band of the lumped-element sections may be broadened as much as desired by making the sections longer and increasing the number of reactive elements used. In fact, it seems likely that, for the same length of section, the lumped-element type is capable of better performance. However, if extended sections are used it is probably easier to build the distributed type than the lumped-element type, which calls for a plurality of accurately dimensioned and spaced elements.

Conclusions

The construction of several types of wave-guide differential phase-shift sections has been described and their properties analyzed. Δ90-degree sections are capable of converting linearly polarized dominant waves into circularly polarized dominant waves, and vice versa. Δ180-degree sections are capable of converting linearly polarized waves into *rotated* linearly polarized waves, and of converting *clockwise*-rotating circularly polarized waves into *counterclockwise*-rotating circularly polarized waves. Such sections may be combined to perform a number of useful functions. Among these, one of particular interest is that of producing a continuously adjustable phase change by means of pure rotation of one of the sections. A phase changer of this type has the rather unique property that it is capable of transmitting substantially 100 per cent of the incident power at high power levels, and this allows it to be used in a number of applications where high-impedance phase changers cannot be used.

Appendix

The following analysis will demonstrate the change in phase produced by rotation of the Δ180-degree section in the phase-changer assembly. Fig. 12 shows the assembly, with the several sets of principal axes appropriately

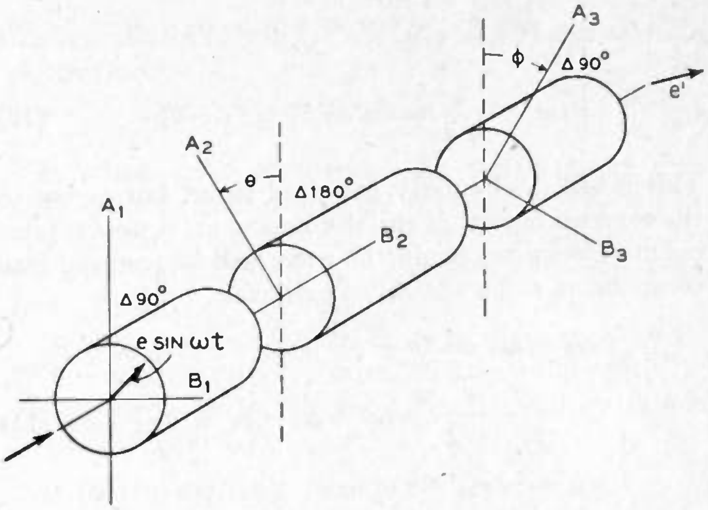


Fig. 12—Generalized phase changer.

labeled and the input wave $e \sin \omega t$ polarized at 45 degrees to the principal axes of the first section. Resolving this into components, we obtain

$$e_{A1} = \frac{1}{\sqrt{2}} e \sin \omega t; \qquad e_{B1} = \frac{1}{\sqrt{2}} e \sin \omega t. \tag{2}$$

After passing through the first section, the output components will be

$$e_{A1}' = \frac{1}{\sqrt{2}} e \sin\left(\omega t - \beta_1 + \frac{\pi}{2}\right); \tag{3}$$

$$e_{B1}' = \frac{1}{\sqrt{2}} e \sin (\omega t - \beta_1) \tag{4}$$

where β_1 is the absolute phase delay through the section for B-polarized waves, and A-polarized waves have received a relative advance of $\pi/2$ over B-polarized waves. Thus,

$$\begin{cases} e_{A1}' = \dfrac{e}{\sqrt{2}} \cos (\omega t - \beta_1) & (5) \\[2ex] e_{B1}' = \dfrac{e}{\sqrt{2}} \sin (\omega t - \beta_1). & (6) \end{cases}$$

This represents a clockwise-rotating circularly polarized wave which we must now resolve into components along the A_2 and B_2 axes.

$$\begin{cases} e_{A2} = e_{A1}' \cos \theta - e_{B1}' \sin \theta = \dfrac{e}{\sqrt{2}} \cos (\omega t - \beta_1 + \theta) & (7) \\[2ex] e_{B2} = e_{A1}' \sin \theta + e_{B1}' \cos \theta = \dfrac{e}{\sqrt{2}} \sin (\omega t - \beta_1 + \theta). & (8) \end{cases}$$

It may be noted that the space angle θ has now entered as part of the phase angle.

After passing through the Δ180-degree section, the emerging components will be

$$e_{A2}' = \frac{-e}{\sqrt{2}} \cos(\omega t - \beta_1 - \beta_2 + \theta + \pi)$$

$$= -\frac{e}{\sqrt{2}} \cos(\omega t - \beta_1 - \beta_2 + \theta) \tag{9}$$

$$e_{B2}' = \frac{e}{\sqrt{2}} \sin(\omega t - \beta_1 - \beta_2 + \theta). \tag{10}$$

This is still a circularly polarized wave, but owing to the reversal in sign of the A component, it now rotates counterclockwise. Again the wave will be resolved into components along the A_3 and B_3 axes:

$$e_{A3} = e_{A2}' \cos(\theta + \phi) + e_{B2}' \sin(\theta + \phi)$$

$$= -\frac{e}{\sqrt{2}} \cos(\omega t - \beta_1 - \beta_2 + 2\theta + \phi) \tag{11}$$

$$e_{B3} = -e_{A2}' \sin(\theta + \phi) + e_{B2}' \cos(\theta + \phi)$$

$$= +\frac{e}{\sqrt{2}} \sin(\omega t - \beta_1 - \beta_2 + 2\theta + \phi). \tag{12}$$

And finally, after passing through the third section the components are:

$$e_{A3}' = \frac{-e}{\sqrt{2}} \cos\left(\omega t - \beta_1 - \beta_2 - \beta_3 + 2\theta + \phi + \frac{\pi}{2}\right)$$

$$e_{A3}' = \frac{e}{\sqrt{2}} \sin(\omega t - \beta_1 - \beta_2 - \beta_3 + 2\theta + \phi) \tag{13}$$

$$e_{B3}' = \frac{e}{\sqrt{2}} \sin(\omega t - \beta_1 - \beta_2 - \beta_3 + 2\theta + \phi). \tag{14}$$

These add up to make a single wave of the same magnitude as the original and oriented at 45 degrees to the A_3 and B_3 axes, as shown. The total phase shift undergone by the wave in passing through the whole assembly is the sum of the individual B-axis delays plus the space angles $2\theta+\phi$. Thus, by adjusting either or both of these space angles the phase of the output wave may be arbitrarily altered. It is clear that a rotation of section II through θ degrees produces 2θ change in electrical phase angle.

Plane Discontinuities in Coaxial Lines*

JOHN W. MILES†

Summary—**The present paper establishes the equivalent circuit of a plane discontinuity in a coaxial line as a simple shunt capacitance. This capacitance is calculated for concentric changes of cross-section and concentric disks. In order to utilize the results of the analogous discontinuities in parallel-plate guides, "equivalent radii" are asymptotically calculated; and the results are sufficiently accurate for most practical applications.**

Introduction

THE PAPER by Whinnery, Jamieson, and Robbins[1] has treated the most important cases of coaxial-line discontinuities and gives results which are sufficiently accurate for the majority of engineering applications. The following is intended to supplement their work by: (a) obtaining approximate results which, the author believes, are more accurate near the cutoff frequency of the TM_{01} mode (although the differences are primarily of academic interest); (b) illustrating a somewhat different approach to the solution of the boundary-value problem; (c) solving separately the problem of a disk or window; and (d) obtaining "equivalent radii" for use with the plane-parallel-plate results from asymptotic comparisons of the two sets of results.

Relative to item (a), footnote reference 1 utilizes a frequency factor F which is obtained by a systematic solution of a finite number of the simultaneous equations for the amplitudes of the field modes excited by the discontinuity. This factor becomes infinite at the cutoff frequency of the TM_{01} mode, while the following treatment offers a factor which is finite at this point. While the author believes that the latter factor is more accurate (at least near the cutoff point in question) than that of reference 1, neither the analyses of footnote references 1 nor 2 have been shown to be mathematically rigorous. While a rigorous investigation of this point would be quite involved, it would be of considerable interest.

Relative to item (b), the following analysis and results follow from the fundamental theory presented in footnote reference (2). Equations from that paper[2] will be denoted by [ ], in contrast to equations of the present paper which will be denoted by ( ). The notation is that of reference 2.

The Characteristic Impedances

From [39] to [115], it is seen that [35, 36] are valid for plane discontinuities in coaxial lines where only the principal mode is allowed to propagate, a condition universally met in practice. The "transformer ratio" N is then given by

$$N = [\log(a_1/b_1)/\log(a_2/b_2)]^{1/2} \tag{1}$$

where $a_{1,2}$ and $b_{1,2}$ are the outer and inner radii for lines 1 and 2, respectively, and where the characteristic

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[1] J. R. Whinnery, H. W. Jamieson, and T. E. Robbins, "Coaxial-line discontinuities," Proc. I.R.E., vol. 32, pp. 695–709; November, 1944.

[2] John W. Miles, "The equivalent circuit for a plane discontinuity in a cylindrical wave guide," Proc. I.R.E., vol. 34, pp. 728–742; October, 1946.

impedance in each of the lines is, according to (8), $Z_0^{1,2}=\eta_{1,2}$. The equivalent circuit is then given by Fig. 3 of footnote reference 2, and is a simple shunt element plus a transformer.

In order to eliminate the transformer from the equivalent circuit, it is expedient to redefine the characteristic impedances as

$$Z_0 = \eta \log (a/b). \tag{2}$$

Inasmuch as coaxial lines frequently utilize a dielectric other than air, it must be observed that $\eta_{1,2}$ are not necessarily identical; however, it may be shown that the correct specifications of η and β (i.e., λ) in all equations are the only changes necessary to make the analysis of reference (2) valid for non-air dielectrics. In most cases it will suffice to use the approximations in section G[1] for changes of dielectric.

When the characteristic impedances of the principal modes are taken as in (2), the equivalent circuit of a plane discontinuity is rigorously established as a shunt element, and the definition (2) will be implicit in the following analysis.

The Eigenfunctions

The eigenfunctions for the present problem are given by [107–115]. For the problem to be treated, circular symmetry exists; hence, $m=0$, and only TM_{0n} modes need be considered for the case where only the principal mode is freely propagated. Accordingly, the superscript TM and the subscript m may be dropped, and the eigenfunctions become:

$$\Phi_0(r) = \bar{r}_1[\log (a/b)]^{-1/2}r^{-1} \tag{3}$$

$$\Phi_n(r) = \bar{r}_1 M_n[J_1(\mu_n r)N_0(\mu_n b) - N_1(\mu_n r)J_0(\mu_n b)] \tag{4}$$

$$M_n = \frac{\pi\mu_n}{2^{1/2}}\left\{\left[\frac{J_0(\mu_n b)}{J_0(\mu_n a)}\right]^2 - 1\right\}^{-1/2} \tag{5}$$

while the eigenvalues μ_n are given by

$$J_0(\mu_n a)N_0(\mu_n b) = J_0(\mu_n b)N_0(\mu_n a)$$
$$\mu_0 = 0. \tag{6}$$

In writing (5), the Wronskian [116] has been used to eliminate the Bessel functions N_0. Since the eigenfunctions (3, 4) are all radial, the vector notation may be dropped.

The Asymptotic Expansions

In obtaining approximate solutions to the problems to be treated, it is expedient to introduce the asymptotic expansion of the eigenfunctions. The leading terms in the asymptotic expansions of Bessel's functions are[3]

$$J_p(x) = \left(\frac{2}{\pi x}\right)^{1/2} \cos\left[x - \left(\frac{2p+1}{4}\right)\pi\right] \tag{7}$$

$$N_p(x) = \left(\frac{2}{\pi x}\right)^{1/2} \sin\left[x - \left(\frac{2p+1}{4}\right)\pi\right]. \tag{8}$$

[3] E. Jahnke and F. Emde, "Tables of Functions," Dover Press, New York, N. Y., 1943.

Substituting (7, 8) in (4, 5) yields

$$\phi_n(r) = (a-b)^{-1/2}r^{-1/2}\cos[\mu_n(r-b)],\quad n \geq 1, \tag{9}$$

while substitution of (7, 8) in (6) yields the asymptotic eigenvalues

$$\mu_n = n\pi(a-b)^{-1}. \tag{10}$$

It is important to observe that (9, 10) form a complete orthonormal set over the range $r=b$ to $r=a$ if (9) is allowed for $n=0$, but ϕ_0 given by (9) is different than the true ϕ_0 given by (3), and the asymptotic eigenfunctions given by (9) are not orthogonal to (3).

In order to illustrate the order of approximation involved in using (10) in place of (6), the values given by (6) are compared with those given for (10) by Jahnke and Emde[3] (Table I).

Table I

	$\mu_n b$ from (6)			$\mu_n b$ from (10)		
(a/b)	1.2	2	3	1.2	2	3
n						
1	15.70	3.12	1.55	15.71	3.14	1.57
2	31.41	6.27		31.42	6.28	
3	47.12	9.42		47.12	9.42	

The agreement is clearly excellent for small values of (a/b) and is still within $1\frac{1}{2}$ per cent for $a/b=3$, the largest value normally encountered in practice. As indicated, the agreement becomes increasingly better for large n.

Changes in Cross Section

Since the present paper is relatively more of academic than practical interest, and since the results of footnote reference (1) are sufficient for most engineering applications, only the changes of cross section involving either a change of the inner- or outer-conductor diameter (but not both) will be solved for useful results. These two problems can be treated together by setting up the problem where the inner-conductor diameter in-

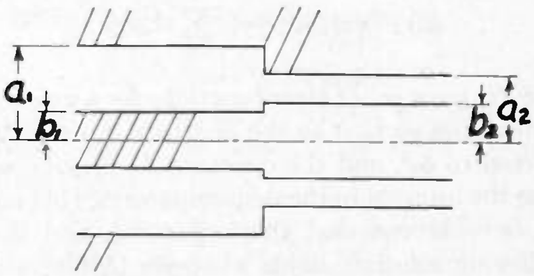

Fig. 1—Coaxial line discontinuity.

creases and the outer diameter decreases at the discontinuity, as shown in Fig. 1. From the results obtained for the changes only of outer or inner diameter, a reasonable approximation to the general case of Fig. 1 will be inferred, since this case is not treated in reference (1). Many other changes of cross section of practical interest are shown in reference (1), and approximations yielding their equivalent circuits from the simpler cases are presented in the literature.[1,4]

The Variational Solution

From [38, 34, 20, 9] the shunt susceptance B for the discontinuity of Fig. 1 is given by

$$\frac{B}{Y_0^1} = \sum_{n=1}^{\infty} \sum_{p=1}^{2} \overline{B}_n{}^p \left[\frac{\int_\sigma \phi_n{}^p(r)E(r)rdr}{\int_\sigma \phi_0{}^1(r)E(r)rdr} \right]^2 \tag{11}$$

$$\overline{B}_n{}^p = \frac{B_n{}^p}{Y_0^1} = \left(\frac{\zeta^p}{\zeta^1}\right)\left[\left(\frac{\mu_n{}^p}{\beta^p}\right)^2 - 1\right]^{-1/2}. \tag{12}$$

$\overline{B}_n{}^p$ is the field susceptance of the nth mode in line $p(p=1$ or $2)$ expressed relative to Y_0^1, and β^p is the phase constant $(2\pi/\lambda^p)$ in line $p(\lambda^1$ differs from λ^2 if the dielectrics in lines 1 and 2 differ); for the cases to be treated, where only the principal modes propagate in the two lines, $\overline{B}_n{}^p$ is positive real for n greater than zero. In setting up (11) the integrations with respect to the angular co-ordinate have been carried out (since all fields are functions of r alone), and the range of integration σ is over the aperture, in the case of Fig. 1 from $r=b_2$ to $r=a_2$.

It is appropriate to emphasize again that the fact that Z_0 was defined differently in (1) than in [9] does not affect the validity of (11), since all impedances are calculated relative to Y_0^1.

It has already been pointed out that the asymptotic eigenfunctions are not orthogonal to the principal wave, and in solving (11) the exact eigenfunctions will be used for purposes of integration. As to the form of the solution, it is expedient to take advantage of the fact that the form of the principal wave is independent of the dimensions of the guide (i.e., for the coaxial line it varies as r^{-1}, independent of a and b). Hence, if the field is expanded in the form

$$E(r) = A_0\phi_0{}^1(r) + \sum_1^\infty A_s\psi_s(r) \tag{13}$$

where $\psi_s(r)$ are a set of eigenfunctions for a guide having the same cross section as the aperture σ, the ψ_s will be orthogonal to $\phi_0{}^1$, and the constant A_0 may be selected to make the integral in the denominator of (11) unity. It should be observed that this argument, and therefore the following solution, holds wherever [39] is valid.

In solving (11) through the substitution of (13), the coefficients A_s are found by appealing to Schwinger's variational principle[2]; for, since only one set of modes (TM) is excited by the discontinuity under consideration, (11) is an absolute minimum with respect to variations of $E(r)$ about its true form. Thus, if the terms B^0, C_s, and D_{ss} are defined as

$$\overline{B}^0 = (A_0)^2 \int_\sigma \int_\sigma \phi_0{}^1(r)G(r,r')\phi_0{}^1(r')rr'drdr' \tag{14}$$

$$C_s = A_0 \int_\sigma \int_\sigma \phi_0{}^1(r)G(r,r')\psi_s(r')rr'drdr' \tag{15}$$

$$D_{ss'} = \int_\sigma \int_\sigma \psi_s(r)G(r,r')\psi_{s'}(r')rr'drdr' \tag{16}$$

$$G(r,r') = \sum_{n=1}^{\infty}\sum_{p=1}^{2} \overline{B}_n{}^p\phi_n{}^p(r)\phi_n{}^p(r'), \tag{17}$$

the substitution of (13) in (11) yields

$$\frac{B}{Y_0^1} = \overline{B}^0 + 2\sum_s C_sA_s + \sum_s\sum_{s'} D_{ss'}A_sA_{s'}. \tag{18}$$

Minimizing (18) with respect to each of the expansion coefficients A_s yields the determining equations

$$\sum_{s'} D_{ss'}A_{s'} = -C_s, \tag{19}$$

and substituting (19) in (18) yields

$$\frac{B}{Y_0^1} = \overline{B}^0 + \sum C_sA_s. \tag{20}$$

To complete the solution, the definition of A_0 yields

$$1/A_0 = \int_\sigma [\phi_0{}^1(r)]^2 rdr. \tag{21}$$

By virtue of the variational principle, B^0, the first approximation to (B/Y_0^1), is larger than the true value of (B/Y_0^2), and as more terms C_sA_s (each of which is implicitly negative) are successively included in (20), the value of B obtained approaches the true value uniformly but is always larger.

For the change of cross section in a coaxial line shown in Fig. 1, the functions $\psi_s(r)$ may be chosen[2] as $\phi_n{}^2(r)$, so that $D_{ss'}$ vanishes for $p=2$, unless $s=s'=n$. The substitution of $\phi_n(r)$ and $\phi_n{}^2(r)$ in (19–21) and integration from $r=b_2$ to $r=a_2$ yields

$$B^0 = (A_0)^2\sum_1^\infty \overline{B}_n{}^1(I_{n0})^2 \tag{22}$$

$$C_s = A_0\sum_1^\infty \overline{B}_n{}^1 I_{n0}I_{ns} \tag{23}$$

$$D_{ss'} = \sum_1^\infty \overline{B}_n{}' I_{ns}I_{ns'} + \delta_{s's}\overline{B}_s{}^2 \tag{24}$$

$$A_0 = \left(\log\frac{a_1}{b_1}\right)\left(\log\frac{a_2}{b_2}\right)^{-1} \tag{25}$$

$$\begin{aligned} I_{n0} = {} & \left(\log\frac{a_1}{b_1}\right)^{-1/2}\left(\frac{M_n{}^1}{\mu_n{}^1}\right)[J_0(\mu_n{}^1b_2)N_0(\mu_n{}^1b_1) \\ & - N_0(\mu_n{}^1b_2)J_0(\mu_n{}^1b_1) - J_0(\mu_n{}^1a_2)N_0(\mu_n{}^1b_1) \\ & + J_0(\mu_n{}^1b_1)N_0(\mu_n{}^1a_2)] \end{aligned} \tag{26}$$

$$I_{ns} = \frac{2}{\pi}M_n{}^1M_s{}^2\left(\frac{\mu_n{}^1}{\mu_s{}^2}\right)[(\mu_n{}^1)^2 - (\mu_s{}^2)^2]^{-1}$$

[4] 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.

$$\cdot\left\{\left[J_0(\mu_n{}^1b_2)N_0(\mu_n{}^1b_1) - J_0(\mu_n{}^1b_1)N_0(\mu_n{}^1b_2)\right] - \left[\frac{J_0(\mu_s{}^2b_2)}{J_0(\mu_s{}^2a_2)}\right]^2\left[J_0(\mu_n{}^1a_2)N_0(\mu_n{}^1b_1) - J_0(\mu_n{}^1b_1)N_0(\mu_n{}^1a_2)\right]\right. \tag{27}$$

where the integrals have been evaluated from the standard forms.[3]

At this point it is convenient to make use of the asymptotic results (7, 8) which yield, after some algebraic manipulation,

$$B^0 = 2\Gamma\kappa\sum_1^\infty [n^2-\kappa^2]^{-1/2} \cdot\left[\frac{\sin n\pi(\alpha+\gamma) - \left(\frac{b_2}{a_2}\right)^{1/2}\sin n\pi\gamma}{n\pi\alpha}\right]^2 \tag{28}$$

$$C_s = \frac{4k\Gamma^{3/2}\kappa}{\pi^2}\sum_1^\infty [n^2-\kappa^2]^{-1/2} \cdot\left[\sin n\pi(\alpha+\gamma) - \left(\frac{b_2}{a_2}\right)^{1/2}\sin n\pi\gamma\right] \cdot\left[\frac{(-)^s\sin n\pi(\alpha+\gamma) - \sin n\pi\gamma}{(n\alpha)^2 - s^2}\right] \tag{29}$$

$$D_{ss'} = \frac{8k^2\Gamma\kappa}{\pi^2}\sum_1^\infty [n^2-\kappa^2]^{-1/2}(n\alpha)^2 \cdot\left[\frac{(-)^s\sin n\pi(\alpha+\gamma) - \sin n\pi\gamma}{(n\alpha)^2 - s^2}\right] \cdot\left[\frac{(-)^{s'}\sin n\pi(\alpha+\gamma) - \sin n\pi\gamma}{(n\alpha)^2 - s'^2}\right] + \delta_{s'}^{s}\cdot 2\alpha\kappa[s^2-(\alpha\kappa)^2]^{-1/2}\left(\frac{\zeta^2}{\zeta^1}\right) \tag{30}$$

$$\alpha = \left(\frac{a_2-b_2}{a_1-b_1}\right),\quad \gamma = \left(\frac{a_1-a_2}{a_1-b_1}\right),$$

$$\kappa = \frac{\beta^1(a_1-b_1)}{\pi} \tag{31}$$

$$\Gamma = \alpha^2\left(\frac{a_1-b_1}{b_2}\right)\left(\log\frac{a_1}{b_1}\right)\left(\log\frac{a_2}{b_2}\right)^{-2},$$

$$k = 2^{-1/2}\left(\frac{a_1-b_1}{b_2}\right)^{-1/4}. \tag{32}$$

Now, if a_1, b_1, a_2, and b_2 are allowed to approach infinity, the differences (a_1-b_1) and (a_2-b_2) being kept constant, it is evident that (28) through (32) go directly over to the solution for the parallel-plate problem having the cross section seen between the inner and outer conductors of the coaxial line (Γ approaches unity under these conditions, and the constants C_sA_s are independent of K). Moreover, independent of the values of a_1, b_1, a_2, and b_2, (28) through (32) differ from the parallel-plate solution only in the factor Γ and in the factor $(b_2/a_2)^{1/2}$, which occurs in the numerator of B^0 and C_s.

The factor $(b_2/a_2)^{1/2}$ is unfortunate, but for the case of a step only in the inner conductor ($a_1=a_2$, $\gamma=0$) so that if $\overline{B}_c$ is the susceptance relative to the characteristic admittance $Y_0{}^1$, and $\overline{B}_{pp}$ is the equivalent susceptance for a parallel-plate guide relative to its own characteristic admittance, the result may be expressed

$$\overline{B}_c = \left(\frac{a_1-b_2}{a_1-b_1}\right)\left(\frac{a_1-b_2}{b_2}\right)\left(\log\frac{a_1}{b_1}\right)\left(\log\frac{a_1}{b_2}\right)^{-2}\overline{B}_{pp}. \tag{33}$$

For the case of a step only in the outer conductor ($b_1=b_2$, $\gamma=1-\alpha$), it is found that

$$\overline{B}_c = \left(\frac{a_2-b_1}{a_1-b_1}\right)\left(\frac{a_2-b_1}{a_2}\right)\left(\log\frac{a_1}{b_1}\right)\left(\log\frac{a_2}{b_1}\right)^{-2}\overline{B}_{pp}. \tag{34}$$

As already noted, the general case ($a_2\neq a_1$, $b_2\neq b_1$) of Fig. 1 does not yield a susceptance which is directly proportional to that of the analogous parallel-plate problem; although this general case is not of as great practical importance as are the two special cases covered by (33) and (34), an interpolation between the last two results yields

$$\overline{B}_c = \Gamma\left[\frac{1+\left(\dfrac{a_1-a_2}{b_1-b_2}\right)\left(\dfrac{b_2}{a_2}\right)}{1+\left(\dfrac{a_1-a_2}{b_1-b_2}\right)}\right]\overline{B}_{pp}, \tag{35}$$

which should furnish a reasonably accurate approximation.

The factors introduced in (33) and (34) could be converted to "equivalent radii" of the coaxial lines. Whinnery and Jamieson, for instance, have specified equivalent radii for coaxial lines in order to use their parallel-plate results,[4] and if their results are expressed relative to the characteristic impedances of the two problems, they will approximate those of (33) and (35), although the present analysis gives results which are considerably more accurate.

More complex discontinuities can be approximately treated by the use of the above results.[1]

Parallel-Plate Results

The plane-parallel-plate results for a change of cross section valid up to the cutoff point of the TM_{01} mode is given by [95] as

$$\overline{B}_{pp} = 2\kappa\log\left[\frac{1}{4}\left(\frac{1}{\alpha}-\alpha\right)\left(\frac{1+\alpha}{1-\alpha}\right)^{1/2(\alpha+1/\alpha)}\right] \tag{36}$$

$$+ 4\kappa\left[\frac{\Delta_1\cos^4\left(\dfrac{\pi\alpha}{2}\right)}{1+2\Delta_1\sin^2\left(\dfrac{\pi\alpha}{2}\right)}\right] \tag{37}$$

$$\Delta_1 = (1-\kappa^2)^{-1/2} - 1, \tag{37}$$

which is valid for a step in either the inner or the outer conductor.

For the case of a window, [87] gives

$$\overline{B}_{pp} = 4\kappa \log\left[\csc\left(\frac{\pi\alpha}{2}\right)\right] + 8\kappa\left[\frac{\Delta_1 \cos^4\left(\frac{\pi\alpha}{2}\right)}{1 + 2\Delta_1 \sin^2\left(\frac{\pi\alpha}{2}\right)}\right]. \tag{38}$$

It should be noted that neither (36) nor (38) becomes infinite at the cutoff frequency of the TM_{01} mode, and the resonance predicted by the use of the frequency factor[1] does not actually occur. (It is easily shown that this result is independent of the use of the asymptotic approximations.)

Accuracy of Asymptotic Expansions

Inasmuch as the variational principle demands that the first variation of B vanish with respect to first-order variations of the trial field about the true form of the field, and variation of the eigenfunctions is tantamount to variation of the trial field, it may be inferred that the accuracy of the results obtained in using the asymptotic expansions is limited by the accuracy of the eigenvalues; in most practical cases this amounts to an error of less than 1 per cent, depending only on the coaxial ratio (b/a). It was demonstrated[2] that the error in (38) up to the cutoff frequency of the next mode was a small fraction of 1 per cent, and it was heuristically argued that a similar, although somewhat greater, error could be expected from the use of (36). The author, therefore, believes that the results given by (33) (34), (36), and (39) will be more accurate than those in the literature[1] near the cutoff frequency of the TM_{01} mode, although this conclusion is evidently open to question.

The error introduced by the use of the asymptotic eigenvalues may be decreased by adding to the results (33) and (35) a perturbation calculated by substituting for $\overline{B}_n{}^p$ in (11) the differences between $\overline{B}_n{}^p$ calculated for the exact and asymptotic eigenvalues, respectively, and using $E(r) = r^{-1}$, but the improvement to be expected (about 1 per cent) scarcely justifies the additional computation.

Numerical Calculations

Comparison of the results of (33) and (34) with Figs. 8 and 9, respectively, of footnote reference 1 shows differences of at most $2\frac{1}{2}$ per cent, so that no advantage results from plotting them separately.[5] However, there is considerable difference between the results when dimensions are comparable to the wavelength. From (36), the frequency factor to be used for the susceptance is

$$F = 1 + \left\{\frac{2\Delta_1 \cos^4\left(\frac{\pi\alpha}{2}\right)}{\left[1 + 2\Delta_1 \sin^2\left(\frac{\pi\alpha}{2}\right)\right] L(\alpha)}\right\}, \tag{39}$$

where $L(\alpha)$ is the logarithm in (37) for a change of cross section or the logarithm in (38) for a disk. At $\lambda = \frac{1}{2}(a_1 - b_1)(= b/2$ in footnote reference 1), (39) simplifies to

$$F_{\max} = 1 + \left[\frac{\cos^4\left(\frac{\pi\alpha}{2}\right)}{\sin^2\left(\frac{\pi\alpha}{2}\right) L(\alpha)}\right]. \tag{40}$$

F is plotted in Fig. 2 for a change of cross section for $\alpha = \frac{1}{2}, \frac{1}{4}$ and compared with the results of footnote reference 1, Fig. 13.

In Fig. 3, $F_{\max}$ versus α is plotted for the change of cross section. It is evident that there is considerable difference between the two methods of frequency correction, and since the present results give the exact correction for the next higher mode,[2] while that of Whinnery[1] contains the approximate treatment of several higher modes, the author believes that the correction of (39) is more accurate, at least near cutoff where the first-mode correction predominates. As stated earlier, this conclusion is still open to question.

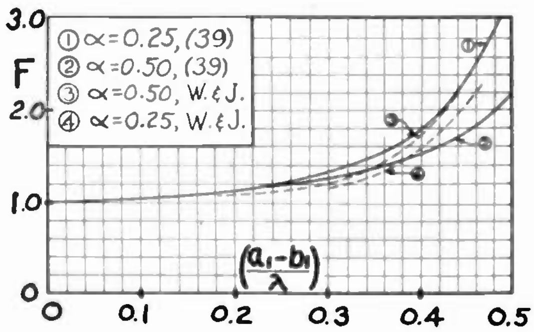

Fig. 2—Frequency-correction factor from equation (39).

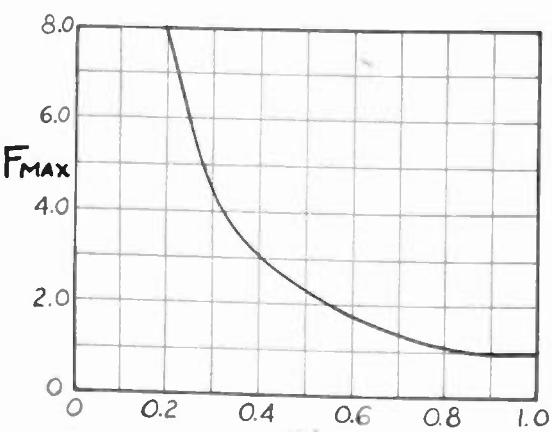

Fig. 3—Maximum frequency-correction factor from equation (40).

[5] Results (in the form of formulas and curves) calculated by Julian Schwinger are given for changes of diameter of both inner and outer conductors and for disks on both conductors in the "Wave Guide Handbook," M.I.T. Radiation Laboratory Report 41-1/23/45 (available through Dept. of Commerce, Washington, D. C., and soon through McGraw-Hill Book Co., New York, N. Y.) These results are apparently about the same as calculated above.

The Inverse Nyquist Plane in Servomechanism Theory*

GEORGE B. CRISS†, ASSOCIATE, I.R.E.

Summary—**The application of an inverse Nyquist diagram to the study of servomechanism performance is discussed. The use of the inverse plane has some advantages over the customary procedure. It is shown that system stability may be studied in a manner analogous to the conventional Nyquist method. The system frequency response and the effect of system parameters upon performance may be determined by simple graphical methods.**

THE RESULTS which will be obtained in this paper are equally applicable to servomechanism theory and feedback-amplifier theory. For the sake of clarity, the conventional terminology of servomechanisms will be employed throughout, with the understanding that the proper interpretation of the symbols will give the corresponding result for feedback amplifiers.

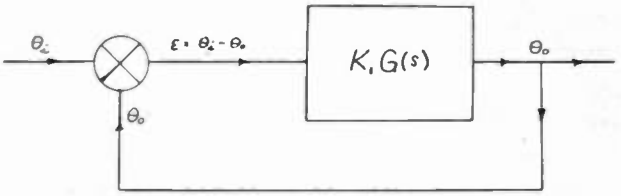

Fig. 1—Block diagram of single-loop servomechanism.

The familiar feedback loop of a single-loop servomechanism is shown in Fig. 1. As is well known,[1] the response of such a system is given by

$$\frac{\theta_0(s)}{\theta_i(s)} = \frac{k_1G(s)}{1 + k_1G(s)} \tag{1}$$

where the loop gain G, and the input and output angles θ_i and θ_0, are functions of the complex frequency, $s=\sigma+j\omega$, and the gain factor k_1 is invariant with s. It is customary to obtain information about the stability of the system and the response in terms of the frequency spectrum by examining $k_1G(j\omega)$. A method which has some advantages under many conditions is described below.

Let

$$k = \frac{1}{k_1}$$

$$L(s) = \frac{1}{G(s)}.$$

Then

$$\frac{\theta_0(s)}{\theta_i(s)} = \frac{1}{1 + kL(s)} \tag{2}$$

which is simpler in form than (1). We shall call kL the loss function, and k the loss coefficient.

We wish, first, to obtain information about the stability of the system by examination of kL; second, to obtain the system frequency response by as simple a method as possible; and third, to obtain a means of determining the system parameters for proper operation.

It is a well-known theorem in the theory of functions of a complex variable[2] that, if $f(z)$ is a function of a complex variable z, and if certain prescribed conditions are fulfilled, then, as z traces a given contour C_1 in the z plane in the counterclockwise direction, the function $f(z)$ will traverse a contour C_2 in the $f(z)$ plane, such that the net number of times C_2 encircles the origin of the $f(z)$ plane in a counterclockwise direction will be equal to the number of zeroes of $f(z)$ within C_1, minus the number of poles of $f(z)$ within C_1.

Following this scheme, let s trace the contour shown in Fig. 2, where we deliberately by-pass any poles of $(1+kL)$ lying on the real frequency axis, and the semicircle is to be taken at infinite radius in the limit.[3] It is evident that the existence of any zeroes of $(1+kL)$ within the right half-plane thus encircled will be an

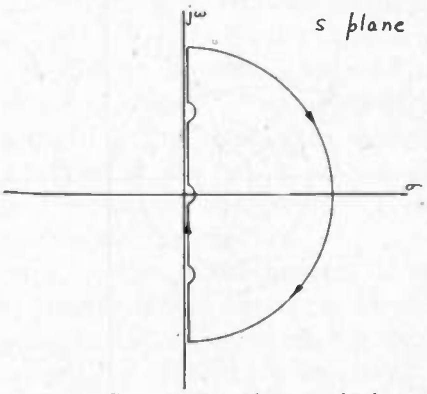

Fig. 2—Contour in s plane equivalent to kL contour.

indication of system instability. The resulting contour traversed by $(1+kL)$ in the $(1+kL)$ plane will then encircle the origin a number of times equal to the excess of zeros over poles of $(1+kL)$ lying within the right half-plane of s. We may instead consider the kL plane: then the corresponding contour traced by kL in the kL plane will encircle the point $(-1+j0)$ a number of times equal to the excess of zeros over poles of $(1+kL)$ in the right half-plane of s.

There are two cases to be considered: first, if $(1+kL)$ has no poles in the indicated area, the kL contour will

* Decimal classification: 621.375.104×R363.23. Original manuscript received by the Institute, November 27, 1946.

† Frankford Arsenal, Philadelphia, Pa.

[1] L. A. MacColl, "Fundamental Theory of Servomechanisms," D. Van Nostrand Co., Inc., New York, N. Y., 1945; pp. 10–57.

[2] R. Rothe, F. Ollendorff, and K. Pohlhausen, "Theory of Functions," Technology Press, M.I.T., Cambridge, Mass., 1942; p. 51.

[3] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., Inc., New York, N. Y., 1945; pp. 103–169.

encircle the critical point $(-1+j0)$ in the counterclockwise sense a number of times equal to the number of roots of $(1+kL)$ with positive real parts, the existence of which is an indication of system instability. Our result in this case is formally similar to the conventional Nyquist criterion.

In the second case, consider the existence of poles of $(1+kL)$ in the indicated area. Poles of $(1+kL)$ will occur only at zeros of k_1G, so that the number of poles of $(1+kL)$ is equal to the number of times k_1G encircles the origin of the conventional Nyquist plane. We have our choice of two possible courses in this case. We may take account of the poles of $(1+kL)$ as indicated, or we may eliminate the poles of $(1+kL)$ by not permitting the contour of s or k_1G to encircle these poles by the customary expedient of cutting the contour and encircling the poles by vanishingly small circles.

In Fig. 3 there is shown a typical kL contour, only the portion corresponding to positive real frequencies being shown. Examination of (1) reveals that at any frequency ω, θ_0/θ_i is represented in magnitude by the reciprocal of the length of the vector from $(-1+j0)$ to the point on the contour corresponding to ω, and in angle by the negative of the angle of this vector. Loci of constant $|\theta_0/\theta_i|$ are then concentric circles about the point $(-1+j0)$, with radius equal to $1/|\theta_0/\theta_i|$. Such circles, with values of $|\theta_0/\theta_i|$ marked, are shown in Fig.

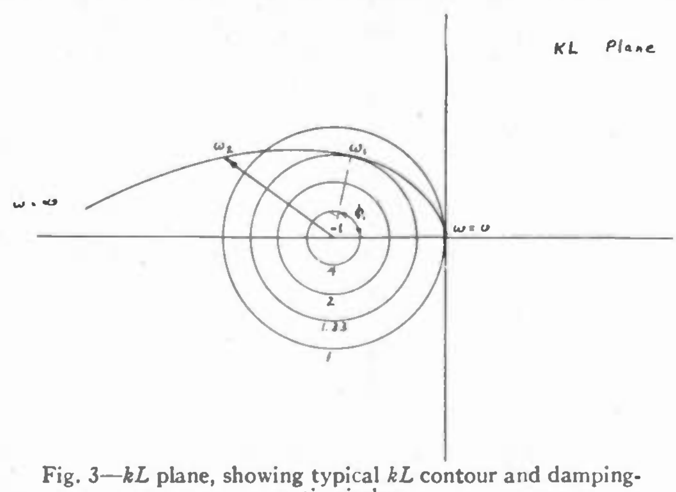

Fig. 3—kL plane, showing typical kL contour and damping-ratio circles.

3. By making use of such a diagram, the determination of the system frequency response from a known characteristic is considerably simpler than by other methods. For example, in Fig. 3 it may be seen by inspection that a resonance peak of magnitude 1.33 and phase angle $-\phi_1$ occurs at frequency ω_1, so that the so-called "damping ratio" is 1.33. The complete characteristics of magnitude and phase of $|\theta_0/\theta_i|$ versus frequency may be plotted with ease.

The third, and perhaps most important, function which we wish to perform is to determine the system parameters required for proper operation. One frequently recurring problem is the determination, in a given system, of the gain coefficient k_1 which will result in a given damping ratio. Using the inverse Nyquist plane, the determination becomes quite simple. We note that one possible method would be to plot kL for different values of k, and choose the contour which is tangent to the circle representing the desired damping ratio. The gain coefficient is then the reciprocal of the loss coefficient corresponding to the chosen contour.

However, it will be seen that the same effect may be produced by changing the scale of the kL plane, maintaining the contour constant. Thus, a decrease in gain corresponds to a shrinkage of the co-ordinate system by the same ratio. In order to make best use of this scheme, we may plot kL initially with k set equal to unity, as is shown in Fig. 4.

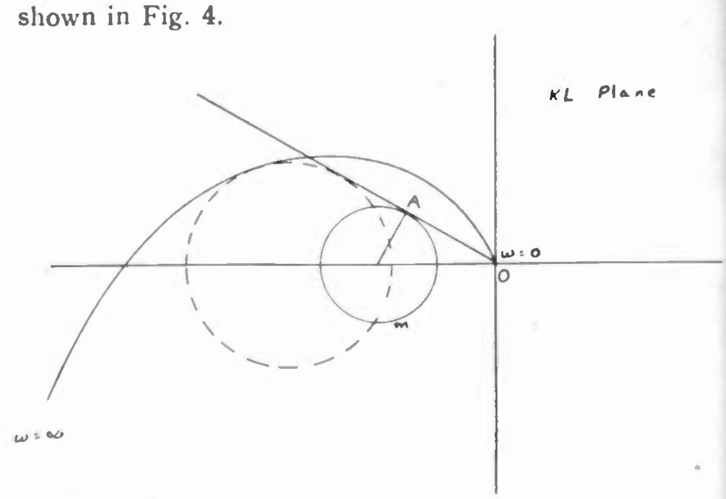

Fig. 4—Graphical construction to determine gain coefficient for desired damping ratio.

Let the desired damping ratio be m. The radius of the corresponding circle will be $1/m$. Now we may increase the scale of the kL plane until the damping-ratio circle is tangent to the contour. The ratio by which the scale was increased will then be k_1, the gain coefficient for the desired damping ratio.

The desired result may be obtained easily by a graphical method. In Fig. 4, the line OA, of slope $-1/\sqrt{m^2-1}$, is the locus of circles of damping ratio m as the scale is increased. We must, therefore, find the circle with center on the negative real axis, tangent to both the line OA and the contour. It is evident that this operation may be performed easily with a compass or dividers. In many cases the intersection of the line OA with the contour may be used as a first approximation for the point of tangency, since the result often approximates the correct result within the desired precision.

The general behavior of the system under change of parameters may be studied using the kL plane. For example, it is seen that the effect of the gain upon the resonant frequency of the system may be obtained. The effect of the addition of tandem lead, lag, and attenuating networks may be studied in a manner analogous to that employed with the conventional k_1G diagram, by virtue of the fact that the loss function for a tandem combination of individual networks is equal to the product of the individual loss functions.

Discussion on

"Factors Affecting the Accuracy of Radio Noise Meters"*

HAROLD E. DINGER AND HAROLD G. PAINE

Alan Watton, Jr.:[1] There was published in the January, 1947, issue of the PROCEEDINGS an excellent paper by H. E. Dinger and H. G. Paine discussing the factors that contribute to the difficulty of accurately measuring radio-frequency noise voltages and field strengths. The paper contains experimental data demonstrating the limitations of the conventional noise meter, particularly in the measurement of impulse noise. This type of noise meter uses a D'Arsonval meter as the indicating instrument and, as a means of obtaining a logarithmic scale, an automatic-volume-control circuit acting upon the gain tubes. The paper also reviews a number of factors which are pertinent in attempting to devise improved noise meters. The comments which follow are intended to set out from a somewhat different point of view some other possibilities for improvement.

The basis for these remarks is the observation that the most fundamental defect in present-day noise meters lies in the wide difference in transient behavior between these meters and conventional radio receivers. In consequence, there is the possibility that significant progress can be made along the line of reducing this divergence. Of course, one difficulty arises from the fact that receivers vary widely in their transient characteristics depending upon the application (a.m. and f.m. communication, loran, radar, etc.). But in any given classification there apparently is sufficient resemblance to make possible an attack upon the problem from this standpoint.

The aircraft a.m. communication (superheterodyne) receiver may be analyzed in detail as an example. For this purpose this type of receiver may be separated into four components: (a) the tuned stages, (b) the second detector, (c) the audio amplifier, and (d) the headset including the ear cavity of the operator.

With regard to the transient behavior of the tuned stages, the two important factors are: (a) the band-pass characteristics of these stages, and (b) the automatic-volume-control (a. v. c.) action.

The band-pass characteristics of the tuned stages of the noise meter must match those of the tuned stages of the receiver in order that the transient action of the two may be the same. Reasoning on general terms, Fourier integral theory shows that two networks will have the same transient behavior if they have identical steady-state response characteristics in both transmission and phase shift. Furthermore, the tuned stages are ladder networks, and Bode[2] has shown that for two ladder networks to have identical phase-shift characteristics, the necessary and sufficient condition is that the two networks have identical transmission characteristics over the entire frequency range extending to zero and infinite frequencies, respectively. In more particular terms, the requirement is that the pass band of the tuned stages of the noise meter and of the receiver shall be the same both in bandwidth and in the bandshape, at least over the range of frequencies at which appreciable gain is obtained. Detailed treatments of the transient behavior of the tuned stages of receivers are available.[3,4]

The action of the delayed-a.v.c. circuit, normally used in the type of receiver we are considering, is such that in the critical case of barely detectable interference, the output of the tuned stages is too small to overcome the delay bias of the a.v.c. circuit. In consequence, the tuned stages are in the region of linear operation. To maintain similarity to this condition in the design of the noise meter, the a.v.c. circuit used in present-day noise meters would have to be abandoned and the tubes in the tuned stages of the meter operated with fixed bias. However, a multiple-tap attenuator would have to be provided, possibly at the input, to handle the wide range of input voltages (1 microvolt to 1 volt) of interest in radio noise measurements.

Coming now to the second detector, it seems that a match in transient characteristics can be obtained most easily by using in the noise meter the same tube and other components as used in the receiver.

Providing in the noise meter a counterpart to the audio amplifier of the receiver is a problem similar to that encountered in connection with tuned stages. Again, Fourier integral theory shows that two networks will have the same transient behavior if they have identical steady-state response in both transmission and phase shift.

By the same reasoning as that used above, it can be seen that it is necessary to provide, in the noise meter, an amplifier that is similar in steady-state transmission in both the useful and the drop-off frequency bands to the audio amplifier of the receiver. Such an amplifier can

* PROC. I.R.E., vol. 35, pp. 75–81; January, 1947.

[1] Propeller Laboratory, Engineering Division; Headquarters, Air Matériel Command, Wright Field, Dayton, Ohio. The opinions expressed herein are those of the author and do not necessarily reflect the official viewpoints of the Air Force.

[2] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., Inc., New York, N. Y., 1945.

[3] Samuel Sabaroff, "Impulse excitation of a cascade of series tuned stages," PROC. I.R.E., vol. 32, pp. 758–760; December, 1944.

[4] David B. Smith and William E. Bradley, "Theory of impulse noise in ideal frequency-modulation receivers," PROC. I.R.E., vol. 34, pp. 743–751; October, 1946.

be easily designed by appropriate methods. It should be noted that the receiver audio-amplifier response used shall be that obtained with the head set as the output load.

In attempting to construct in the noise meter an arrangement that will serve as the analogue of the head set and associated ear cavity of the radio receiver, we certainly face a complex design problem. However, of the several possible solutions, one that is the most direct (but not necessarily the easiest) is to make use of the equivalent circuit[5] of this electroacoustic configuration and construct a circuit such that an output voltage is obtained that represents the sound pressure in the ear cavity.

Suitably combining the four components, each in itself similar in transient action to the corresponding component in the radio receiver, will give an assembly the output voltage of which will be the same, except for a calibration factor, as the acoustic output of the radio receiver when both are subjected to the same radio interference input signal. There would remain, then, the problem of measuring this output voltage.

Now, as the authors point out, "It has been generally agreed that the interfering effect of frequently occurring impulses, for most applications, is more nearly proportional to the peak value than to the effective or average values." It would seem reasonable, considering all applications of radio receiving equipment (communication, navigation, radar, etc.), to say that even in the case of impulses occurring at low recurrence rates, or in the case of random noise, the peak value is a more acceptable measure of the interfering effect than any other easily defined criterion.

Therefore, the problem is to measure this peak value of the output voltage of the noise meter. Now, in technically significant cases (e.g., electrical control circuits used on aircraft), impulses occurring at recurrence rates as low as once every three minutes are deemed to be radio interference. Obviously, the discharge time constant of a suitable transient-peak voltmeter circuit would be impracticably large. However, a cathode-ray oscilloscope, with a calibrated scale and with or without photographic recording, would accurately indicate the peak value of the transient. Furthermore, such an arrangement would be adaptable to use in semiportable equipment.

This approach to the problem would seem to offer a number of outstanding advantages. The performance of a noise meter can be thus arranged to bear a rational relationship to the behavior of radio receiving equipment. Furthermore, this relationship is capable of being written into a specification in terms of steady-state sine-wave measurements. Finally, the calibration in actual use can be made directly by a standard-signal generator.

On the other hand, some experimental work that the writer has had performed has made it apparent that the maximum allowable limits of radio noise to be used with this method are very different (in general, much higher) than those rather generally accepted for use with the present-day type of noise meter. Thus new limits would have to be determined.

In conclusion, the writer is in full agreement with the result of the experimental study outlined by the authors in the subject paper, and these comments are intended to be supplementary to the conclusions which the authors reached.

[5] H. F Olson, "Elements of Acoustical Engineering," D. Van Nostrand Co., Inc., New York, N. Y.; p. 229; 1940.

Harold E. Dinger:[6] Mr. Watton has further emphasized the difficulties encountered when attempts are made to standardize radio interference-measuring instrumentation and techniques, especially if single-valued indications are desired. Much of the substance of Mr. Watton's discussion was contained in the original manuscript of the subject paper, which was reduced approximately 50 per cent before submission to the Institute and again by 50 per cent, by request of the Editor, prior to publication. Because of these condensations, considerable information of a pertinent nature did not appear in the published paper. The original version described several oscillographic and photographic methods of indication, as well as additional material on each of the items discussed. One subject in particular that did not receive the treatment merited by its importance is that of overloading or dynamic range. This factor is especially prominent in the measurement of impulsive interference.

The general problem of interference measurement must, of course, consider all types of interference, and the units of measurement must be such that they can be related to the subjective effect of the interference on different services, such as a.m. and f.m. broadcasting, television, facsimile, radioteletype, radar, etc.

Since the subject paper was written, considerable work on radio interference measurement has been in progress at the University of Pennsylvania under Navy Department sponsorship. Additional studies are being made by the various service laboratories, the Joint Coordination Committee of Radio Reception, the American Standards Association, and the International Special Committee on Radio Interference. It is anticipated that these studies will result in recommendations for instrumentation and methods of application much more suitable than those currently in use.

One significant feature that is being incorporated into several new measuring equipments is an adaptation of the slide-back voltmeter method of measuring peak values. Although it is somewhat slower than direct-indicating methods and does not lend itself to continuous recording, it has considerable value for use with impulsive interference of relatively low recurrence rates.

The writer is of the opinion that it will probably be desirable to specify two categories of radio-interference

[6] Naval Research Laboratory, Washington, D. C.

measuring equipments; one of which will be a laboratory standard, the other a portable instrument giving indications which have been referenced in some manner to the laboratory standard.

Harold G. Paine:[7] The desirability of eliminating automatic-volume control in radio noise meters, which Mr. Watton mentions, became evident early in the ground work which preceded this paper. Two methods of accomplishing this were discussed at that time: (a) the use of fixed bias in amplifier stages with a suitably designed nonlinear d.c. amplifier in the output to provide the logarithmic scale, and (b) the use of a continuously variable attenuator ahead of a fixed-gain receiver circuit, and adjusting the attenuator for each reading so that a standard matching indication is obtained. Sometime after submission of this paper, Mr. Chappell of Camp Cole Signal Laboratory proposed a device for making noise measurements which utilizes a slide-back voltmeter and a method of comparison against a standard noise generator.

The use of automatic volume control is convenient in the design of light, portable, battery-operated noise meters in which the use of complicated circuits and components is undesirable because of power and weight limitations. The complications arising out of the use of automatic volume control in a radio noise meter, which were pointed out in the above paper, do not necessarily preclude the obtaining of acceptable accuracies in the measurement of noise of a general random character or in the measurement of some impulse noise having a relatively high repetition frequency. It is in the measurement of impulse noise of short duration and relatively low repetition rate that difficulty is encountered, and considerable work remains to be done before suitable techniques and instrumentation are available.

[7] Naval Research Laboratory, Washington, D. C.

Discussion on

"Exact Design and Analysis of Double- and Triple-Tuned Band-Pass Amplifiers"*

MILTON DISHAL

Vernon D. Landon:[1] Dishal has written an excellent summarizing paper on band-pass amplifier design. I believe it is slightly misleading, however, as to the value of Q required for operation of triple-tuned circuits.

On page 620, in speaking of a band-pass filter utilizing three tuned circuits, Dishal says:

"To obtain a flat-topped response with three peaks of equal amplitude in the pass band, all the loading must be removed from the middle tuned circuit . . . . Otherwise, as will be shown later, the outer two peaks of the response will be lower in amplitude than the middle peak."

In the next paragraph he elaborates: "To approach the ideal triple-tuned response curve, the Q of the middle tuned circuit must be of the order of 10 times (or more) the Q of the input and output circuits."

The experimental facts are somewhat at variance to the above, as will be explained. Given three tuned circuits with $Q_1=Q_3$ and with $Q_2=10Q_1$, the circuits may be coupled to obtain the ideal triple-tuned response curve to which Dishal refers. If, for economy, or other reasons, the value of Q_2 must be reduced to only 3 or 4 Q_1, the outside peaks will have lower amplitude than the middle peak (as he states), providing no other circuits constants are changed. However, if at this point Q_1 is reduced somewhat and Q_3 is increased (or the reverse), the equality of the three peaks may be restored. This fact is rather important, as it permits the use of three coupled circuits without having to meet quite as stiff a Q requirement as that proposed by Dishal.

The smallest value of Q that may be employed in the circuit having the highest Q may be found by making use of Dishal's mathematics. In equation (28) on page 623, if the coefficients of F^4 and F^2 are set equal to zero, we have the condition for "maximal flatness"[2]; that is to say, the flattest curve without multiple peaks. This gives the two equations:

$$2K^2-(n_1^2+n_2^2+n_3^2)=0 \tag{1}$$

$$K^4-K^2(n_1^2+n_3^2-n_2(n_1+n_3))+n_1^2n_2^2+n_2^2n_3^2+n_3^2n_1^2=0 \tag{2}$$

where K = the coupling coefficient, and n_1, n_2, n_3 = the inverse of the Q's of the three circuits.

From (1) it appears that, if one of the n's is increased, another must be decreased. Then the largest value, n_0, required for the smallest n, will occur when the two smaller n's have the same value.

Assuming the two smaller n's are n_2 and n_3, we have $n_0=n_2=n_3$, and

$$2K^2-n_1^2-2n_0^2=0 \tag{3}$$

$$K^4-K^2(n_1^2-n_0n_1)+2n_1^2n_0^2+n_0^4=0. \tag{4}$$

Of the three unknowns, K, n_1, and n_0, any one may be assumed fixed, and the other two may be solved for in

* Proc. I.R.E., vol. 35, p. 606–626; June, 1947.

[1] Radio Corporation of America, RCA Laboratories, Princeton, N. J.

[2] V. D. Landon, "Cascade amplifiers with maximal flatness," *RCA Rev.*, Pt. I, pp. 347–363; January, 1941: Pt. II, pp. 481–498; April, 1941.

terms of that one. Cut-and-try methods yield the following solution, which may be checked by substitution:

$$n_0 = 0.236n_1$$
$$K = 0.745n_1.$$

In Dishal's equation (28), the last term of the polynomial under the radical is

$$\left[K^2\left(\frac{n_1+n_3}{2}\right)+n_1n_2n_3\right]^2$$

and is equal to

$$\left(\frac{f_b}{f_0}\right)^6$$

where f_b = the bandwidth at 70 per cent (for the condition of maximal flatness), and f_0 = the resonant frequency.

Then

$$\frac{f_b}{f_0} = \left(K^2\frac{n_1+n_0}{2}+n_1n_0^2\right)^{1/3}$$
$$= 0.737n_1.$$

Now, for Dishal's assumed conditions of

$$n_1 = n_3$$
$$n_2 \doteq 0,$$

we find $K = n_1 = f_b/f_0$ for the maximally flat condition. Dishal assumes that

$$n_2 = \frac{1}{10}\frac{f_b}{f_0}$$

is required.

The present discussion indicates that, when $n_2 = n_3 = n_0$,

$$n_0 = 0.236n_1$$
$$= \frac{0.236}{0.737}\frac{f_b}{f_0}$$
$$= 0.32\frac{f_b}{f_0}$$

is sufficiently small. In other words, the triple-tuned circuit is operable if tuned circuits are available having a Q as high as about

$$3\frac{f_0}{f_b}.$$

Milton Dishal:[3] I would like to take this opportunity to thank Landon for pointing out the fact that it is possible to obtain a triple-tuned response curve having three peaks of equal amplitude and two valleys of equal amplitude, even though the Q of the middle resonant circuit is not infinite. This fact is practically of great importance and I think it is safe to say that when triple-tuned band-pass circuits are used, and a symmetrical band pass is desired when the circuits are correctly resonated, the "Q distribution" pointed out by Landon, i.e., $Q_2 = Q_3 = AQ_1$, should be used rather than the Q distribution mentioned in my paper of $Q_1 = Q_3$ and $Q_2 = \infty$. (A is a number whose value depends on the type of response desired.)

[3] Federal Telecommunication Laboratories, Inc., Nutley, N. J.

In discussing this matter, I think it is important to clearly separate, in the following manner, the two types of useful responses which can be obtained: (a) the type of response having n maxima of equal amplitude and $(n-1)$ minima of equal amplitude within the pass band where n is the number of resonant circuits used; and (b) the type of response having a single maximum which occurs at the middle of the pass band.

Response Type (a)

It should be realized that my paper considered this type of response only. The main reason why I was led to consider the Q distribution $Q_1 = Q_3$ and $Q_2 = \infty$ was that this seemed to be the only distribution which would allow exact design equations to be obtained which were not hopelessly complicated. Unfortunately, this still seems to be the case, and it should be realized that Landon's discussion gives no solution for this type of response. Thus, insofar as the response having peaks and valleys within the pass band is concerned, we have only the qualitative fact that this response can be obtained without the necessity for having $Q_2 = \infty$.

However, insofar as practical design is concerned (where exact final values must be experimentally determined), Landon's equations can be used to obtain the transitional shape condition, and the coefficient of coupling can then be increased very slightly to produce a multiple-peaked response. (It will also be necessary to make $Q_2 = Q_3$ more than $4.24Q_1$; the greater the peak-to-valley ratio desired, the greater will be the required ratio of $Q_2 = Q_3$ to Q_1.)

It may be pointed out here that the following procedure may possibly allow an exact solution to be obtained for the multiple-peaked response, for conditions other than my assumed conditions of $Q_1 = Q_3$ and $Q_2 = \infty$. As pointed out by Landon, the transitional-shape condition or condition of maximal flatness is obtained when, in my (28), the coefficients of F^4 and F^2 equal zero and when the constant term equals $(\Delta f_{3\,db}/f_0)^6$. When the multiple-peaked response is desired, the coefficients, rather than equaling zero, must equal some specific value. These specific values can be obtained by substituting, in my (28a), the required values of K and n as obtained from (34), (35), and (36). We can then set the more general coefficients given in (28) equal to the above values obtained through the medium of (28a). It is possible that a usable solution may then be obtained from the three resulting simultaneous equations. Thus, to obtain a certain percentage bandwidth between outside peaks of $(\Delta f_p/f_0)$ with a certain peak-to-valley ratio defined by γ of (35) we find that the coefficient of F^4

must equal $2(\Delta f_p/f_0)^2$; the coefficient of F^2 must equal $1(\Delta f_p/f_0)^4$; and the constant term must equal $[\gamma(1+\gamma^2)(\Delta f_p/f_0)^3]^2$. Thus, in order to find the required ratio of

$$\frac{K}{(\Delta f_0/f_0)}, \quad \frac{n_1}{(\Delta f_0/f_0)} \quad \text{and} \quad \frac{n_0}{(\Delta f_p/f_0)}$$

Thus, using Landon's Q distribution, it will be necessary to solve the three simultaneous equations given below

$$\begin{cases} 2K^2 - (n_1^2 + 2n_0^2) = 2\left(\dfrac{\Delta f_p}{f_0}\right)^2 \\ K^4 - K^2 n_1(n_1 - n_0) + n_0^2(n_0^2 + 2n_1^2) = 1\left(\dfrac{\Delta f_p}{f_0}\right)^4 \\ K^2 \dfrac{n_1 + n_0}{2} + n_1 n_0^2 = \gamma(1+\gamma^2)\left(\dfrac{\Delta f_p}{f_0}\right)^3. \end{cases}$$

Response Type (b)

In his discussion, Landon has given the exact solution for the constants required to give the limiting case of this single-peaked type of response. (It should be noted that the required conditions for maximal flatness could also be obtained by equating simultaneously to zero the two parts of (30).)

As mentioned previously, Landon's solution is of great practical importance because of the relatively small value of Q required in the two high-Q circuits. When identical circuits are cascaded, the required Q_1 for a given $(\Delta f_{3\,db}/f_0)$ will be even smaller than $0.737(\Delta f_{3\,db}/f_0)$, and, therefore, the high-Q circuits $(Q_2 = Q_3)$ whose Q must equal $4.24Q_1$ will require a necessary Q even less than $3(f_0/\Delta f_{3\,db})$. For example, for five cascaded triple-tuned circuits, the required Q_1 is $Q_1 = 0.54(f_0/\Delta f_{3\,db})$ and, therefore, the required $Q_2 = Q_3$ is only $Q_{2,3} = 2.3(f_0/\Delta f_{3\,db})$.

Since the equations are quite simple, I think it would be helpful to tabulate the equations which enable the complete and exact design to be accomplished for cascaded single-, double-, and triple-tuned band-pass circuits using the maximally flat type of response.

N-Cascaded Triple-Tuned Circuits

$$Q_2 = Q_3 = 4.24Q_1$$

$$K = \frac{0.745}{Q_1}$$

$$\frac{Q_1}{f_0/\Delta f_{3db}} = 0.737[2^{1/N} - 1]^{1/6}$$

$$\frac{V_0}{V} = \left[(2^{1/N} - 1)\left(\frac{\Delta f}{\Delta f_{3db}}\right)^6 + 1\right]^{N/2}$$

or

$$\frac{\Delta f}{\Delta f_{3db}} = \frac{1}{[2^{1/N} - 1]^{1/6}}\left[\left(\frac{V_0}{V}\right)^{2/N} - 1\right]^{1/6}$$

$$\frac{\text{Gain}_{(\text{per stage})}}{G_m/4\pi\Delta f_{3db}\sqrt{C_1C_3}} = 1.03[2^{1/N} - 1]^{1/6}$$

$$\tan\theta_{(\text{per stage})} = \frac{-\left(\dfrac{\pm\Delta f}{\Delta f_{3db}}\right)\left[\left(\dfrac{\Delta f}{\Delta f_{3db}}\right)^2 - \dfrac{2.01}{[2^{1/N}-1]^{1/3}}\right]}{\left[\dfrac{1.93}{[2^{1/N}-1]^{1/6}}\left(\dfrac{\Delta f}{\Delta f_{3db}}\right)^2 - \dfrac{1}{[2^{1/N}-1]^{1/2}}\right]}.$$

N-Cascaded Double-Tuned Circuits

$$Q_1 = Q_2$$

$$K = \frac{1}{Q}$$

$$\frac{Q}{f_0/\Delta f_{3db}} = 1.414[2^{1/N} - 1]^{1/4}$$

$$\frac{V_0}{V} = \left[(2^{1/N} - 1)\left(\frac{\Delta f}{\Delta f_{3db}}\right)^4 + 1\right]^{N/2}$$

or

$$\frac{\Delta f}{\Delta f_{3db}} = \frac{1}{[2^{1/N} - 1]^{1/4}}\left[\left(\frac{V_0}{V}\right)^{2/N} - 1\right]^{1/4}$$

$$\frac{\text{Gain}_{(\text{per stage})}}{G_m/4\pi\Delta f_{3db}\sqrt{C_1C_2}} = 1.414[2^{1/N} - 1]^{1/4}$$

$$\tan\theta_{(\text{per stage})} = \frac{\mp\left[\left(\dfrac{\Delta f}{\Delta f_{3db}}\right)^2 - \dfrac{1}{[2^{1/N} - 1]^{1/2}}\right]}{\pm\left[\dfrac{1.414}{[2^{1/N} - 1]^{1/4}}\left(\pm\dfrac{\Delta f}{\Delta f_{3db}}\right)\right]}.$$

N-Cascaded Single-Tuned Circuits

$$\frac{Q}{f_0/\Delta f_{3db}} = [2^{1/N} - 1]^{1/2}$$

$$\frac{V_0}{V} = \left[(2^{1/N} - 1)\left(\frac{\Delta f}{\Delta f_{3db}}\right)^2 + 1\right]^{N/2}$$

or

$$\frac{\Delta f}{\Delta f_{3db}} = \frac{1}{[2^{1/N} - 1]^{1/2}}\left[\left(\frac{V_0}{V}\right)^{2/N} - 1\right]^{1/2}$$

$$\frac{\text{Gain}_{(\text{per stage})}}{G_m/2\pi\Delta f_{3db}C} = [2^{1/N} - 1]^{1/2}$$

$$\tan\theta_{\text{per stage}} = \left(\pm\frac{\Delta f}{\Delta f_{3db}}\right)[2^{1/N} - 1]^{1/2}.$$

Corrections

I would like to take this opportunity to note that the sign before the radical of equation (30) should be $\pm$.

It should also be noted that in Fig. 6, Chart I, and equation (27), the subscripts 1 and 2 refer to the input and output resonant circuits, 3 referring to the middle circuits; whereas in equations (28), (29), and (30) the subscripts 1 and 3 refer to the end resonant circuits and 2 refers to the middle circuit. In this discussion, both Landon and I have used the latter notation.

In the pi equivalent for the transformer in Fig. 3, the denominators for the vertical legs should have the sign in front of M reversed; i.e., it should be $\mp$. In the tee equivalent for the transformer the dot should be removed from the equality sign of the vertical leg.

Equation (1) should be referrred to Fig. 1 instead of Circuit I of Fig. 2; (2) should be referred to Fig. 4 instead of Circuit A of Fig. 5; and (3) should be referred to Fig. 1 instead of Fig. 5.

In (15), the last bracketed term under the second square-root sign in the denominator should be squared, and the fifth line from the bottom of page 617, in the second column, should read (19) instead of (18).

Discussion on

"The Cathode-Coupled Amplifier"*

KEATS A. PULLEN, JR.

John R. Clark:[1] In reading the recent paper, by Keats A. Pullen, Jr., I was unable to follow the reasoning leading to the choice of a low value of cathode-coupling resistor, as well as the expression in the Appendix which would indicate that the over-all gain was directly proportional to the load impedance. Hence, I submit another analysis which I believe is more conventional and yields results which check more closely with Mr. Pullen's measured voltage-gain curves.

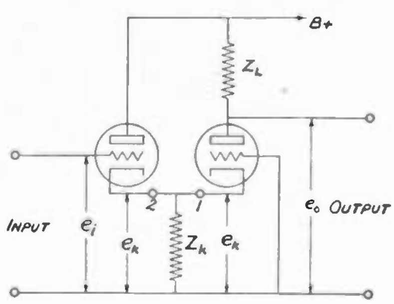

Fig. 1

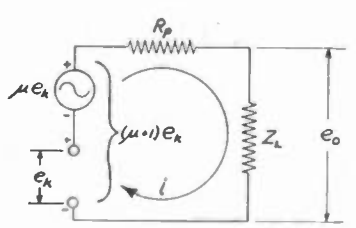

Fig. 2

Fig. 1 shows the basic cathode-coupled amplifier. To the right of point 1 (Fig. 1), we find a conventional cathode-driven amplifier, its equivalent circuit being shown in Fig. 2. From Fig. 2 it is evident that the voltage gain of this stage is

$$VG = \frac{e_0}{e_k} = (\mu + 1)\frac{Z_L}{R_p + Z_L},$$

and the input impedance,

$$Z_{in} = \frac{e_k}{i} = \frac{R_p + Z_L}{\mu + 1}.$$

Again referring to Fig. 1, we find a conventional cathode follower to the left of point 2 whose load impedance consists of Z_{in} shunted by Z_k. The over-all gain expression yields little additional information if only we remember that the gain of the cathode-follower stage increased from zero to $\mu/\mu+1$ as the load impedance increases from zero without limit. Hence, it is apparent that the higher the value of Z_k becomes, the greater will be the over-all gain, and that the real function of Z_k is to provide direct-current continuity to ground, and possibly, in wide-band applications, to flatten the frequency-response characteristic at the expense of gain.

It would then appear that Z_L should be chosen in a manner similar to that used for conventional triode amplifiers, and that the direct-current value of Z_k should be chosen to provide suitable bias for both triodes. Greater flexibility of design may be had by using the biasing arrangements shown in Fig. 3.

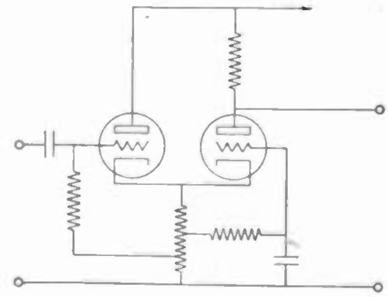

Fig. 3

Adolf Reitlinger:[2] There is obviously an error in the calculation for voltage gain, or effective amplification, for a grounded-grid amplifier. Representing the

input voltage as e_1
input current as i_1
series impedance as Z_1
cathode-circuit impedance as Z_k
plate alternating current as i_2
voltage developed by the tube as e_2
voltage appearing across Z_k as e_3
amplification factor as μ
plate resistance as R_p
load resistance as Z_L
effective impedance to a voltage applied across points a and b of Fig. 4 as Z_{e0}

* Proc. I.R.E., vol. 34, pp. 402–405; June, 1946.
[1] Purdue University, Lafayette, Ind.
[2] 91-01 98 Street, Woodhaven, L. I., N. Y.

current flowing into point a, Fig. 4 as I
output voltage as e_0,

Mr. Pullen states:

$$e_1 = i_1Z_1 + Z_k(i_1+i_2)$$

$$\mu(i_1+i_2)Z_k = i_2(Z_k+R_p+Z_L) + i_1Z_k$$

voltage gain $= e_0/e_1 = \mu Z_L/[Z_1(1+\mu)+(Z_L/Z_k+1)(R_p+Z_1)]$.

No account has been taken for the fact that fractions of e_1 and e_2 appear in phase across R_p and Z_L, and out of phase across Z_k.

The equations for effective amplification and for Z_{eg} are:

The voltage developed by the tube alone is μe_3, and the total current flowing into point a (Fig. 4) is:

$$I = \frac{e_3}{Z_k} + \frac{e_3}{R_p+Z_L} + \frac{\mu e_3}{R_p+Z_L}$$

$$= e_3\left(\frac{1}{Z_k} + \frac{\mu+1}{R_p+Z_L}\right). \quad (1)$$

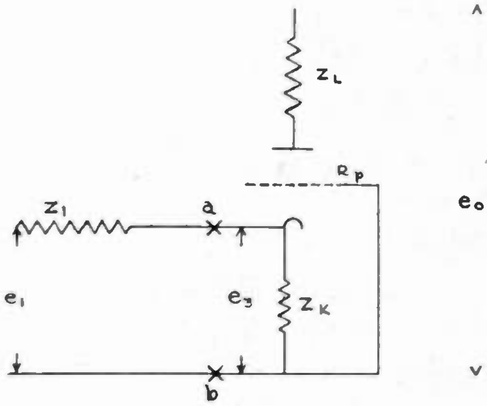

Fig. 4

The equivalent impedance presented to the source across points a and b (Fig. 4) is:

$$Z_{eg} = \frac{e_3}{I} = \frac{1}{\dfrac{1}{Z_k} + \dfrac{\mu+1}{R_p+Z_L}} = \frac{Z_k\left(\dfrac{R_p+Z_L}{\mu+1}\right)}{Z_k + \dfrac{R_p+Z_L}{\mu+1}}$$

$$= \frac{Z_k(R_p+Z_L)}{(\mu+1)Z_k + R_p + Z_L}. \quad (2)$$

$$e_3 = \frac{e_1Z_{eg}}{Z_1+Z_{eg}}$$

$$= e_1\frac{Z_k(R_p+Z_L)}{Z_k(R_p+Z_L)+Z_1[(\mu+1)Z_k+R_p+Z_L]}. \quad (3)$$

$$\text{Gain} = G_2 = \frac{e_3(\mu+1)Z_L}{Z_L+R_p} \times \frac{1}{e_1}$$

$$= \frac{Z_k(R_p+Z_L)(\mu+1)Z_L}{\{Z_k(R_p+Z_L)+Z_1[(\mu+1)Z_k+R_p+Z_L]\}(R_p+Z_L)}$$

$$= \frac{Z_kZ_L}{Z_1Z_k + (Z_1+Z_k)\dfrac{Z_L+R_p}{\mu+1}}. \quad (4)$$

For the circuit of Fig. 5, Z_1 for the second section $=0$. G_2; the gain of second section, becomes

$$G_2 = \frac{Z_L}{\left(\dfrac{Z_L+R_p}{\mu+1}\right)}.$$

Z_{eg} of the second section (2),

$$\frac{Z_k\left(\dfrac{R_p+Z_L}{\mu+1}\right)}{Z_k+\left(\dfrac{R_p+Z_L}{\mu+1}\right)},$$

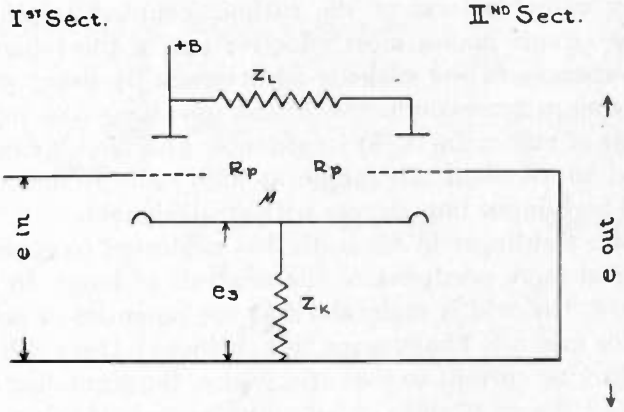

Fig. 5

is the effective Z_k for the first section. G_1 of the first section is, according to Terman,[3]

$$G_1 = \frac{\mu}{\mu+1}\,\frac{Z_{eg}}{Z_{eg}+\dfrac{R_p}{\mu+1}} = \frac{\mu Z_{eg}}{R_p+(\mu+1)Z_{eg}}$$

and the total gain $= G_1G_2$. Provided μ and R_p are equal in both sections, and calling

$$\left(\frac{Z_L+R_p}{\mu+1}\right) = Z_R,$$

and

$$\frac{R_p}{\mu+1} = R_0,$$

$$G_1 = \frac{\mu Z_kZ_R}{(\mu+1)[Z_kZ_R+R_0(Z_k+Z_R)]}$$

$$G_2 = \frac{Z_L}{Z_R}$$

$$G_1G_2 = \frac{\mu Z_LZ_k}{(\mu+1)[Z_kZ_R+R_0(Z_k+Z_R)]}.$$

Keats A. Pullen, Jr.:[4] Mr. Clark's discussion on the Appendix does not appear to provide the clarification

[3] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Co., New York, N. Y., 1943.

[4] The Pullen Laboratories, Brooklyn 17, N. Y.

promised. The solution on each type of stage was only indicated, and no attempt at consolidation of the derivation made because, although such a combination would verify the initially rising gain with rising cathode impedance, the decline of gain at high values of this impedance did not follow. Rather than make a statement of the combined gain, this writer preferred to supply experimental curves. It might be pointed out, however, that Mr. Clark has neglected the input impedance of the cathode circuit of the grounded-grid amplifier. Although the cathode follower admittedly has a low output impendance, it is by no means zero, as is required by his Fig. 2 and his equations.

In Mr. Clark's Fig. 3 he has included an excellent and very valuable form of the cathode-coupled amplifier. This circuit makes most effective use of the inherent advantages of the cathode-input stage by using it at maximum transconductance and operating the input stage at maximum input impedance. This circuit can be used to excellent advantage at high radio frequencies and high-input impedances with small signal level.

Mr. Reitlinger in his study has neglected to observe several facts pertinent to the analysis at hand. In his Fig. 4, the grid is maintained at the potential of point (*b*) or ground. That means that, although there will be a charging current to the capacitance, the grounding action of the grid brings about an isolation between cathode and plate. For this reason, the analysis as performed —namely, injection of signal into a cathode impedance and the use of that as the voltage generator for the voltage to be amplified—is valid.

The curve showing the cathode input voltage on the grounded-grid stage, Fig. 8 of the paper, shows that there is a voltage loss in coupling the grounded-grid stage to the cathode-follower stage. Mr. Reitlinger's analysis does not agree with this experimentally determined curve.

John R. Clark:[1] It appears that I was not too convincing in my first comment, so I will attempt to be more complete, as suggested by Mr. Pullen. To obtain an expression for the over-all gain, let us first assume that Z_k is a parallel-resonant circuit and of such high impedance as to have negligible shunting effect on Z_{in}, the input impedance of the cathode-driven or grounded-grid stage. The total load on the cathode follower would then be Z_{in}, and its gain would be

$$VG_1 = \frac{e_k}{e_i} = \frac{\mu}{\mu+1} \times \frac{Z_{in}}{\dfrac{R_p}{\mu+1} + Z_{in}},$$

and substituting the value of Z_{in} derived in my first comment,

$$VG_1 = \frac{\mu}{\mu+1} \times \frac{R_p + Z_L}{2R_p + Z_L}.$$

The total gain would then be the product of the two individual stage gains:

$$VG = VG_1 \times VG_2 = \frac{\mu}{\mu+1} \times \frac{R_p + Z_L}{2R_p + Z_L} \times (\mu+1)\frac{Z_L}{R_p + Z_L} = \frac{\mu Z_L}{2R_p + Z_L}.$$

This expression assumes identical values of μ, g_m, R_p for both triodes as well as an extremely high value of Z_k.

To see the effect of finite values of Z_k on the over-all gain, it should be noted that the gain of the second stage is independent of Z_k. Therefore, it will be necessary only to find the effect on the gain of the cathode follower. If we set up the ratio F of VG_1 with $Z_k = \infty$ to VG_1 with Z_k finite, the actual gain will then be

$$VG = \frac{\mu Z_L}{2R_p + Z_L} \times F$$

where

$$F = \left[\frac{\dfrac{Z_{in}Z_k}{Z_{in}+Z_k}}{\dfrac{R_p}{\mu+1} + \dfrac{Z_{in}Z_k}{Z_{in}+Z_k}}\right] \times \left[\frac{\dfrac{R_p}{\mu+1} + Z_{in}}{Z_{in}}\right] = \frac{R_p + (\mu+1)Z_{in}}{(\mu+1)Z_{in} + R_p + \dfrac{Z_{in}R_p}{Z_k}}.$$

Substituting,

$$Z_{in} = \frac{R_p + Z_L}{(\mu+1)},$$

and rearranging,

$$F = \frac{\left(\dfrac{Z_L}{R_p} + 2\right)}{\left(\dfrac{Z_L}{R_p} + 2\right) + \dfrac{\left(\dfrac{Z_L}{R_p} + 1\right)}{(\mu+1)Z_k/R_p}}.$$

It is interesting to note that the maximum value of F equals unity, and obtains when Z_k increases without limit. It is of further interest to note that F may be evaluated for the limiting values as Z_L approaches zero and as Z_L increases without limit.

By limiting Z_L and Z_k to real values, the curves shown in Fig. 6 are obtained. Mr. Pullen's choice of $Z_k = 1/gm$ can now be seen to give approximately one-half to two-thirds of the maximum possible gain.

Mr. Pullen did not state the conditions under which he obtained his experimental curves, but I have been able to duplicate them by setting up the circuit as in Fig. 1 of his paper. Under these conditions the quiescent grid and plate voltages soon get out of any normal operating region and the corresponding changes in tube parameters, I believe, account for the falling off of gain at high values of Z_k. To check this point, I have set up a circuit similar to my Fig. 3, but with two separate plate-voltage supplies so that fixed operating points could be maintained with changing values of R_L and R_k. Using tube parameters measured on a General Radio vacuum-tube bridge and decade boxes to obtain accurate values of R_L and R_k, the maximum discrepancy between measured and calculated values of gain over most of the R_k range covered in Fig. 6 was within 5 per cent.

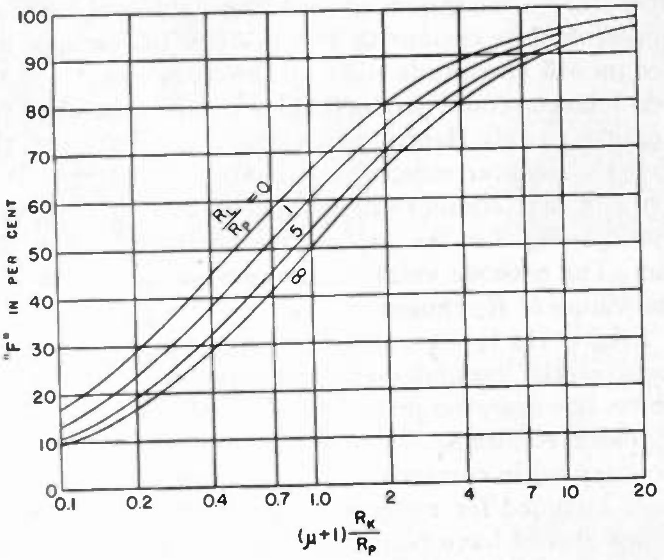

Fig. 6

My conclusions are that Mr. Pullen's criterion for selecting Z_k works very well when using triodes of medium μ and when it is not desired to complicate the circuit in the slightest. However, there are many applications where Z_k and R_L may be easily made to have low values of direct-current resistance and high total impedance, or where the quiescent tube voltages may be maintained easily at normal operating values by relatively simple circuit modifications. Under these conditions, it is advisable to choose a reasonably high value of Z_k.

Mr. Pullen's expression for voltage gain still is beyond my comprehension, particularly since the gain would appear to increase without limit in direct proportion to Z_L. Neither his experimental curves nor mine tend to uphold this conclusion.

Adolf Reitlinger:[2] My analysis was for the voltage gain of a cathode-follower circuit driving a grounded-grid amplifier, and the formula for the gain of such a circuit was derived. No comment was made with respect to Mr. Pullen's voltage-gain measurements.

I did state, however, that there is obviously an error in the calculation for voltage gain or effective amplification for a grounded-grid amplifier, and the exact formula was derived.

Mr. Pullen, in his reply, admits that my analysis—namely, that based upon the injection of a signal into the cathode impedance and the use of that as the voltage generator for the voltage to be amplified—is correct.

Mr. Pullen does not give an explanation of his mesh equations, which, solved for effective amplification, will be:

$$e_0/e_i = \mu Z_L/[Z_i(1+\mu) + (Z_i/Z_k + 1)(R_p + Z_i)].$$

This result differs from my derivation, and I assume that Mr. Pullen agrees that my analysis is the valid one.

Keats A. Pullen, Jr.:[4] Mr. Clark has made the assumption in his discussion that the input power drawn from the cathode-follower stage by the grounded-grid amplifier is negligible. As is known by anyone who has used grounded-grid amplifiers, in some cases it is possible to get more radio-frequency output than the direct-power input on these stages. This results from inclusion of drive power in the output power. It is for this reason that Mr. Clark's derivation is in error.

I regret that typographical errors caused the grounded-grid equations to be in error. The second and third equations should have read

$$-[(i_1 + i_2)Z_k]\mu = i_2(Z_k + R_p + Z_L) + i_1 Z_k \qquad (5)$$

$$e_0/e_i = (\mu+1)Z_L/[Z_i(1+\mu) + (Z_i/Z_k + 1)(R_p + Z_L)]. \qquad (6)$$

The simplest method of getting a correct derivation is as follows:

Take

$$i_{p1} = g_{m1}e_{gk1} + g_{p1}e_{pk1} \qquad \text{(See Fig. 7)}$$

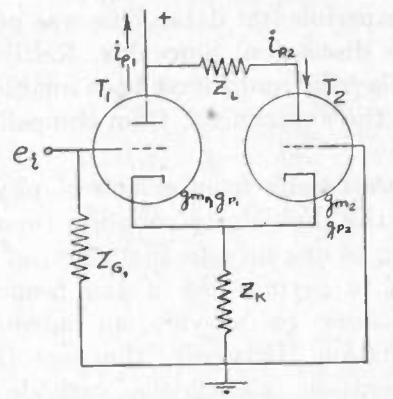

Fig. 7

and

$$i_{p2} = g_{m2}e_{gk2} + g_{p2}e_{pk22}$$

where

$$e_{gk_1} = e_i - Z_k(i_{p1} - i_{p2});\quad e_{pk_1} = -z_k(i_{p1} - i_{p1})$$

$$e_{gk_2} = -Z_k(i_{p_2} - i_{p_1});\quad e_{pk_2} = -Z_L i_{p_2} - Z_k(i_{p_2} - i_{p_1}).$$

Solving for i_{p_2} gives

$$i_{p_2} = \frac{g_{m_1}Z_k(g_{m_2}+g_{p_2})e_i}{[1+Z_k(g_{m_1}+g_{m_2}+g_{p_1}+g_{p_2})+g_{p_2}Z_L+Z_kZ_Lg_{p_2}(g_{m_1}+g_{p_1})]},$$

or

$$V.G._p = \frac{e_0}{e_i} = \frac{i_{p_2}Z_k}{e_i}$$

$$= \frac{g_{m_1}Z_L(g_{m_2}+g_{p_2})Z_k}{[1+Z_k(g_{m_1}+g_{m_2}+g_{p_1}+g_{p_2})+g_{p_2}Z_L+Z_kZ_Lg_{p_2}(g_{m_1}+g_{p_1})]} \qquad (7)$$

and

$$\frac{e_k}{e_i} = \frac{(i_{p_1}-i_{p_2})Z_k}{e_i}$$

$$= \frac{(1+g_{p_2}Z_L)g_{m_1}Z_k}{[1+Z_k(g_{m_1}+g_{m_2}+g_{p_1}+g_{p_2}(+g_{p_2}Z_L+Z_kZ_Lg_{p_2}(g_{m_1}+g_{p_1})]}. \qquad (8)$$

Under normal use, the following approximations may be made

$$g_{p_1} \ll g_{m_1}; \qquad g_{p_2} \ll g_{m_2}$$

$$g_{p_2} Z_L \ll Z_k(g_{m_1} + g_{m_2} + g_{p_1} + g_{p_2})$$

are

$$VG_p = \frac{e_0}{e_i} = \frac{i_{p_2} Z_L}{e_i}$$

$$= \frac{g_{m_1} Z_L (g_{m_2} + g_{p_2}) Z_k}{[1 + Z_k(g_{m_1} + g_{m_2} + g_{p_1} + g_{p_2}) + g_{p_2} Z_L + Z_k Z_L g_{p_2}(g_{m_1} + g_{p_1})]} \tag{7a}$$

$$e_k = (i_{p_1} - i_{p_2}) Z_k$$

$$= \frac{(1 + g_{p_2} Z_L) g_{m_1} Z_k e_i}{[1 + Z_k(g_{m_1} + g_{m_2} + g_{p_1} + g_{p_2}) + g_{p_2} Z_L + Z_k Z_L g_{p_2}(g_{m_1} + g_{p_1}]} . \tag{8a}$$

These reduce approximately to

$$VG_p \approx \frac{g_{m_1} g_{m_2} Z_k Z_L}{1 + (g_{m_1} + g_{m_2}) Z_k} \approx \frac{g_m Z_L}{2} \tag{7b}$$

$$\frac{e_k}{e_i} = VG_k \approx \frac{g_{m_1} Z_k}{1 + (g_{m_1} + g_{m_2}) Z_k} \approx \frac{1}{2} . \tag{8b}$$

It appears that Mr. Reitlinger's equation for the cathode-coupled circuit has only one failing—that of not fitting experimental data. This was pointed out in my previous discussion. Since Mr. Reitlinger has not replied to this point and I have been unable to make his equation fit the experiment, I am compelled to accept this as fact.

This failure results from a lack of physical understanding of the problem of coupling these two tubes. The coupling of one tube to the other, at first glance, would seem to permit use of the properties of the cathode follower to provide an adequate voltage source regulation. However, the fact that cathode-follower operation, giving the cathode current as $i_p = (g_m / [1 + (g_m + g_p) Z_k]) e_c$, actually gets regulation by in effect reducing the equivalent transconductance of the tube to $g_m / [1 + (g_m + g_p) Z_k]$ causes the tube not to do what would appear to occur. The second triode (grounded-grid element) still has full transconductance and, hence, acts as a very heavy load on the input stage.

To drive home this point, consider a voltage of $+1$ volt to appear on the input grid. Normal cathode-follower action would place all this on the cathode. But the effective g_m of the tube has been reduced to $g_m/2$ if $(g_m + g_p) Z_k = 1$. This voltage appears between grid and cathode of the grounded-grid stage. If full g_m is available on the grounded-grid stage, this will cause an opposing plate current to flow through the cathode impedance of magnitude equal to the maximum the cathode follower could produce. This would neutralize the applied signal. Hence, the voltage output from the cathode follower must be automatically reduced. It is evident that equilibrium will occur at half the input voltage. Fig. 8 in the paper experimentally verifies this fact. The cathode voltage-gain curve holds for each of the values of R_L chosen.

I regret the typographical errors in copying the equations of the grounded-grid amplifier from my notes. Since, however, the procedure is one of routine analysis of mesh equations, no difficulties should have been experienced in correcting the equations. The equations were included for reference only and, save for the Z_i which should have read Z_L in the last term of the denominator, would have been sufficiently accurate for most applications. The left-hand term of the second equation obviously should have been negative in sign. The third equation should have read:

$$\frac{e_0}{e_i} = (\mu + 1) Z_L \Big/ \left[Z_i(1 + \mu) + \left(\frac{Z_i}{Z_k} + 1 \right) (R_p + Z_L) \right].$$

The derivation fitting physical facts for the gain of the cathode-coupled stage has now been worked out. The exact resulting equations are given above. The approximate equations are

$$\frac{e_k}{e_i} = VG_k \approx g_{m_1} Z_k / [1 + (g_{m_1} + g_{m_2}) Z_k] \approx \tfrac{1}{2} \tag{8b}$$

$$VG_p \approx g_{m_1} g_{m_2} Z_k Z_L / [1 + (g_{m_1} + g_{m_2}) Z_k] \approx g_m Z_L / 2. \tag{7b}$$

These two expressions obviously satisfy my curves, since g_m drops as Z_k rises.

Contributors to Proceedings of the I.R.E.

George B. Criss (A'47) was born on June 23, 1911, at Schenectady, N. Y. He received the B.S. degree in physics in 1931, and the E.E. degree in 1933 from the College of the City of New York. In 1944 he received the M.S. degree in electrical engineering from the University of Pennsylvania.

In 1942, after a short period in the test department of the General Electric Company at Schenectady, Mr. Criss entered the employ of the War Department at the Frankford Arsenal, Philadelphia, as electrical engineer. During the war he was active in the application of electronic techniques to the development and design of gun-fire-control systems and instruments. Mr. Criss is now engaged in electronic development in the fire-control development division at the Frankford Arsenal.

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For a photograph and biography of C. C. Cutler, see page 1328 of the November, 1947, issue of the Proceedings of the I.R.E.

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A. Gardner Fox (A'40–SM'45) was born on November 22, 1912 at Syracuse, N. Y. He received the S.B. and S.M. degrees in electrical engineering from the Massachusetts Institute of Technology in 1935. From 1935 until early 1936 he was employed in the radio receiver division of General Electric Company, at the end of which time he entered the radio development department of the Bell Telephone Laboratories, Inc. For the next three years he was engaged in the development of mobile and airborne radio transmitters, and in early radar development. In 1939 he was transferred to the radio research department where for three years he conducted research on wave-guide techniques. From 1942 to 1944 he was again occupied with the development of radar systems.

Since 1944 Mr. Fox has been engaged in microwave research as a member of the staff of the Holmdel Radio Laboratory of Bell Telephone Laboratories, Inc. He is currently a member of the committee on Radio Wave Propagation and Utilization of The Institute of Radio Engineers.

George B. Criss

A. S. Gladwin was born at Glasgow, Scotland, on November 26, 1916. He received the B.Sc. degree in electrical engineering from the Glasgow University in 1940. Mr. Gladwin was a member of the scientific staff in the research laboratories of the General Electric Co., Ltd., Wembley, England, from 1940 to 1946. He is now a demonstrator in electrical engineering at King's College, London.

Robert D. Huntoon (A'41–SM'47) was born at Waterloo, Iowa, on July 20, 1909. In 1932 he received the A.B. degree at Iowa State Teachers College, and obtained the M.S. degree in 1935, and the Ph.D. degree in 1938, from the State University of Iowa. He was instructor in physics at New York University from 1938 to 1940, and research physicist for Sylvania Electric Products, Inc., Emporium, Penn., from 1940 to 1941.

Since 1941 Dr. Huntoon has been at the National Bureau of Standards. During 1944 and 1945 he served as expert consultant in the office of Dr. E. L. Bowles, Office of the Secretary of War. He is now chief of the newly-formed electronics section of division 4, and assistant chief of the atomic physics division. Dr. Huntoon is a member of Sigma Xi and the American Physical Society.

A. Gardner Fox

A. S. Gladwin

Edward C. Jordan (S'36–A'39–SM'45) was born in Edmonton, Alberta, Canada, on December 31, 1910. He received the B.Sc. degree in electrical engineering in 1934, and

Robert D. Huntoon

the M.Sc. degree in 1936, from the University of Alberta, and in 1940, the Ph.D. degree from the Ohio State University. He has taught electrical engineering at Worcester Polytechnic Institute and Ohio State University, and was a part-time consultant on the antenna research program at the Ohio State University during the war.

Dr. Jordan is now professor of electrical engineering at the University of Illinois. He is a member of Tau Beta Pi, Sigma Xi, Eta Kappa Nu, and the American Institute of Electrical Engineers.

Edward C. Jordan

Archie P. King

Archie P. King (A'30–SM'45) was born at Paris, France, on May 4, 1901. He received the B.S. degree from the California Institute of Technology in 1927. From 1927 to 1930 he was in the seismological research department of the Carnegie Institution of Washington. Since 1930 he has been with Bell Telephone Laboratories, Inc.

Winston E. Kock (SM'45) was born on December 5, 1909, at Cincinnati, Ohio. He received the E.E. degree from the University of Cincinnati in 1932, and the M.S. degree in physics in 1933. As an Exchange Fellow of the Institute of International Education, he received the Ph.D. degree from the University of Berlin in 1934. Following one year as Teaching Fellow at the University of Cincinnati, he attended the Institute for Advanced Study at Princeton and the Indian Institute of Science at Bangalore, India.

Paul Boris

Winston E. Kock

NILS ERIK LINDENBLAD

Dr. Kock was formerly director of electronic research and development at the Baldwin Piano Co., Cincinnati, Ohio. He is now associated with the Bell Telephone Laboratories, Inc. at Holmdel, N. J., engaged in microwave antenna research. Dr. Kock is a member of the American Physical Society, Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

N. E. Lindenblad (M'34–SM'43) was born on October 30, 1895, in Norrkoping, Sweden. He attended the Norrkoping Technical Evening School during 1911 and 1912, and the Norrkoping Polytechnic Institute from which he received the M.E. degree in 1915. He joined the Swedish Army Signal Corps in 1915. From 1916 to 1919 he studied electrical engineering at the Royal Institute of Technology in Stockholm, after which he came to the United States.

Mr. Lindenblad was associated with the General Electric Company until September, 1920, when he joined the Radio Corporation of America, where he is now located. His major activity has been antenna design, development, and research. He was the recipient of the Modern Pioneer Award of the National Association of Manufacturers in 1940, and was expert consultant to the Secretary of War during World War II.

J. O. McNALLY

J. O. McNally (J'24–A'26–SM'44) was born in Fredericton, New Brunswick, Canada, in 1903. He received the B.S., degree in electrical engineering from the University of New Brunswick in 1924, and joined the technical staff of the Bell Telephone Laboratories, Inc., the same year. Since then, he has been engaged in the development of electron tubes of various types.

For a biography and photograph of JOHN W. MILES, see page 1331 of the November, 1947, issue of the PROCEEDINGS OF THE I.R.E.

Greenleaf W. Pickard (M'12–F'15) was born on February 14, 1877, in Portland, Maine. He has been associated with radio since 1901, when he became engineer for the American Wireless Telegraph and Telephone Company. In 1902, he was made chief engineer for the Federal Wireless Telegraph and Telephone Company. Later, he joined

GREENLEAF W. PICKARD

the American Telephone and Telegraph Company, remaining until 1907, when he organized the Wireless Specialty Apparatus Company, which became the R.C.S. Victor Company of Massachusetts. From 1942 to 1945 Mr. Pickard was director of research for the American Jewels Corporation. He is now associated with the firm of Pickard and Burns, consulting engineers.

Mr. Pickard received the I,R.E. Medal of Honor in 1926 for his "contributions as to crystal detectors, coil antennas, wave propagation and atmospheric disturbances." He also was the recipient of the Armstrong Medal of the Radio Club of America in 1940. He has served on numerous I.R.E. committees, including the Board of Editors, Constitution and Laws, Wavelength Regulation, and Wave Propagation, and was actively associated with the organization of the Boston Section, in 1914. Mr. Pickard was a member of both the Wireless Institute and the Society of Wireless Telegraph Engineers, when these two organizations fused into the present I.R.E. on May 13, 1912.

Mr. Pickard is a Fellow of the American Academy of Arts and Sciences, as well as of the American Institute of Electrical Engineers and the Radio Club of America. He

R. V. POUND

also holds membership in the American Geophysical Union, the American Meteorological Society, and is a life member of the *Société des Radoélectriciens.*

R. V. Pound was born on May 16, 1919, at Ridgeway, Ontario, Canada. In 1941 he received the B.A. degree in physics from the University of Buffalo. From 1942 to 1946 he was a staff member of the Radiation Laboratory at the Massachusetts Institute of Technology, engaged in the development of microwave circuits, especially those for the use of crystal rectifiers as frequency convertors. In 1945 Mr. Pound was elected a Junior Fellow of the Society of Fellows at Harvard University, which appointment he now holds.

William G. Shepherd (A'42) was born on August 28, 1911, at Fort William, Ontario, Canada. He received the B.E.E. degree in 1933, and the Ph.D. degree in physics in 1937, from the University of Minnesota.

From 1933 to 1937, Dr. Shepherd was a teaching fellow in physics at the University of Minnesota. Since 1937 he has been a member of the technical staff of the Bell Telephone Laboratories, Inc, engaged in nonlinear circuit research until 1939, and since in electronics research and develop-

WILLIAM G. SHEPHERD

George Sinclair

ment. He is a member of the American Physical Society and Sigma Xi.

George Sinclair (A'37–SM'46) was born in Hamilton, Ontario, Canada, on November 5, 1912. He received the B.Sc. degree in electrical engineering in 1933 and the M.Sc. degree in 1935 from the University of Alberta, and the Ph.D. degree in 1946 from the Ohio State University. Dr. Sinclair was an instructor in electrical engineering at the University of Alberta for one year, and engineer for the Northern Broadcasting Corporation for two years.

From 1941 to 1947 Dr. Sinclair was a research associate in the department of electrical engineering of the Ohio State University, supervising the research program of the Antenna Laboratory. He is now an assistant professor of electrical engineering at the University of Toronto.

Harlan T. Stetson (A'31) was born in Haverhill, Mass., on June 28, 1885. After graduating from Brown University, he received the Sc.M. degree from Dartmouth College in 1910, and the Ph.D. degree from the University of Chicago in 1915. He was associated with the physics department at Dartmouth for four years, and later taught astronomy and mathematics at Northwestern University. From 1916 to 1929, while he was an assistant professor of astronomy at Harvard University, Dr. Stetson became associated with Mr. Pickard in the investigation of the effect of sunspots on radio reception. From 1929 to 1934 Dr. Stetson was Perkins professor of astonomy at Ohio Wesleyan University and director of the Perkins Observatory, as well as lecturer at Ohio State University. He returned to Harvard in 1934 as a research associate in geophysics. He joined the Massachusetts Institute of Technology in 1936 and is the director of the Cosmic Terrestrial Research Laboratory at Needham, Mass.

Harlan T. Stetson

Eric W. Vaughan

Dr. Stetson is the author of numerous papers on solar activity, radio reception, and ionization of the upper atmosphere. He has been chairman of the Special Committee on Cosmic Terrestrial Relationships of the American Geophysical Union, National Research Council, since 1938. He is the author of "Man and the Stars," "Earth, Radio and the Stars," "Sunspots and Their Effects," and a forthcoming book, "Sunspots in Action."

Eric W. Vaughan (A'40–M'44) was born in Rangely, Maine, on November 1, 1916. He received the A.B. degree in physics in 1938, and the A.M. degree in 1940, from Dartmouth College. He was a graduate student at the Ohio State University from September, 1940, to January, 1942, and then an instructor in the department of electrical engineering for one semester.

Mr. Vaughan became a research associate in June, 1942, working on wartime research projects of the Ohio State University Research Foundation until December, 1945. Mr. Vaughan is now with the Superior Electric Company, Bristol, Conn.

Albert Weiss

Albert Weiss was born on February 5, 1912, in New York, N. Y., and studied at George Washington University, in Washington, D. C. In 1935 Mr. Weiss was employed by the United Transformer Company, and later by the White Sound Company. Since 1941 he has been serving as radio engineer with the ordnance development division of the National Bureau of Standards.

Harold A. Wheeler (A'27–M'28–F'35) was born in St. Paul. Minn., on May 10, 1903. He received the B.S. degree in physics from George Washington University in 1925. From 1925 to 1928 he pursued postgraduate studies in the physics department of Johns Hopkins University, and lectured there during 1926 and 1927. He was employed as laboratory assistant in the radio section of the National Bureau of Standards in 1921, leaving in 1923 to assist Professor Hazeltine and later to join the Hazeltine Corporation in 1924. He was in charge of their Bayside laboratory from 1930 to 1937, and advanced to the position of vice-president and chief consulting engineer of Hazeltine Electronic Corporation.

In 1946 Mr. Wheeler opened his own consulting office in Great Neck, N. Y. He is now also president of Wheeler Laboratories, Inc., an engineering organization engaged in consultation and the construction of special equipment. Mr. Wheeler is a Fellow of the American Institute of Electrical Engineers, and a member of Sigma Xi. He received the Morris Liebmann Memorial Prize in 1940, and was a member of the Board of Directors of the I.R.E. in 1934, and from 1940 to 1945.

Harold A. Wheeler

Correspondence

Multifrequency Bunching in Reflex Klystrons*

During the development of wide-tuning-range reflex klystrons at the Raytheon Manufacturing Company, it has recently been observed that spurious higher-frequency oscillations may occur simultaneously with the desired lower-frequency oscillation. At first, these spurious oscillations were believed to be one of the ordinary $(n+\frac{3}{4})$ repeller modes. However, a more careful examination of the phenomenon indicated the following peculiarities:

(a) The spurious oscillations occurred only in the presence of a vigorous oscillation of the desired frequency.

(b) The spurious oscillations occurred at a repeller voltage that is roughly midway between the mode voltages ordinarily required to sustain the spurious-frequency oscillations in the absence of the low-frequency oscillation.

In an effort to explain this phenomenon, an analysis was made of the bunching action in a reflex klystron when several sinusoidal voltages of different frequencies exist simultaneously across the interaction gap. The analysis shows that the presence of a vigorous low-frequency oscillation may result in a sign reversal of the electronic admittance at a higher frequency, and that, as a result, stable higher-frequency oscillations may simultaneously be obtained when the reflex transit time corresponds to $(n+\frac{1}{4})$ cycles (at the higher frequency). That is, the change in sign resulting from the compression effect of the low-frequency oscillation requires that the transit angle be altered by 180 electrical degrees. This explains the necessity for the repeller voltage being midway between the mode voltages ordinarily required in the absence of the compression.

The expressions for the electronic admittances were found to be

$$
\begin{aligned}
Y_1 &= j\epsilon^{-j2\pi N_1} M_1 J_0\left(\frac{N_1}{N_2}X_2\right)\frac{2J_1(X_1)}{X_1} \\
Y_2 &= j\epsilon^{-j2\pi N_2} M_2 J_0\left(\frac{N_2}{N_1}X_1\right)\frac{2J_1(X_2)}{X_2}
\end{aligned}
\qquad (1)
$$

where N is the transit time in cycles, X is the bunching parameter, and M is the zero-signal admittance—each at its respective frequency. Starting with (1), it may be shown that, even though the electron stream simultaneously presents a negative electronic conductance at both frequencies, the modes of oscillation may be dynamically unstable. For example, the $N_2=4\frac{3}{4}$-cycle mode may be shown to be unstable and ordinarily recessive to the $N_1=2\frac{3}{4}$-cycle mode. It is suggested that the excessive noise sometimes found in klystrons may be attributed to a "fighting action" between a dominant and a recessive unstable mode of oscillation.

The general expression for the admittance at the first frequency may be shown to be

$$
Y_1 = -j\epsilon^{-j2\pi N_1}\frac{M_1}{X_1}\sum_{(k,n)} J_k(X_1)J_n(N_1X_2/N_2)\epsilon^{jn\phi} \qquad (2)
$$

where ϕ is the initial phase angle between the two sinusoidal voltages, and the integer-pairs (k, n) are all values (positive and negative) that satisfy the equation

$$
(1+k)N_1 = -nN_2. \qquad (3)
$$

When N_1 and N_2 are incommensurable, the only integer pair that satisfies (3) is $k=-1$ and $n=0$, and equations (1) result. When N_2 and N_1 are exactly in the ratio of two integers, other pairs arise and the corresponding terms in (2) must be included. By symmetry, the electronic admittance at the other frequency of oscillation may be found by interchanging subscripts in (2) and (3). The expressions derived from (2) for the case where $N_2/N_1=2$ have been found to agree with the experimental data on the generation of second-harmonic power in a reflex klystron.

W. H. Huggins
Air Matériel Command
Watson Laboratories
Cambridge, Mass.

* Received by the Institute, August 28, 1947. A comprehensive paper on the same subject has been concurrently submitted to the Institute for publication consideration.

Comparison of Primary and Secondary Radar System*

I have read the recent paper by Hultgren and Hallman[1] and would like to comment on certain phases of the mathematics used in this paper. I refer particularly to Part I, Section 13 A.

First, the use of units or dimensions of the various terms used in the equations is not consistent. It should be remembered that the dimensions of the various terms of an equation may be handled as a parallel auxiliary equation. Thus, in the equation for a tangential signal, $A_bS_t=3N$, $A_b=$ meter² and $S_t=$ watts/meter². The dimension of the term N is not defined in the text, but may be derived from the known terms of the equation. Thus

$$
A_b\cdot S_t = 3N
$$

$$
\frac{\text{meters}^2}{1}\cdot\frac{\text{watts}}{\text{meters}^2} = \text{watts}.
$$

Turning now to (1a) and (1b) of the Hultgren and Hallman paper, and applying dimension equations to each as was done above, it becomes evident that the term $2N$ is not watts/meter² but simply watts. Since N is not clearly defined by the authors, this discrepancy is confusing until the reader peruses the paper further and definitely determines that the dimension of N is watts.

Another point I should like to make is that the expression for the average absorption cross section of resonant dipoles oriented at random is given as $\lambda^2/4\pi^2$, instead of $\lambda^2/4\pi$ as stated in the paper. Equation (5) is, therefore, incorrect, the correct equation being $A_b=\lambda^2G_b/4\pi$. Substituting this value for A_b in the equation for a tangential signal, it is found that $G_b=3N(4\pi r)^2/\lambda^2G_0P_t$, which would not have been the case had the term $\lambda^2/4\pi$ been used to derive the expression for A_b.

Maurice V. Gowdey
1135 Trenton Avenue
Bremerton, Wash.

* Received by the Institute, August 25, 1947.
[1] R. D. Hultgren and L. B. Hallman, Jr., "The theory and application of the radar beacon," Proc. I.R.E., vol. 35, pp. 716–730; July, 1947.

Reprints Available

In most cases, The Institute of Radio Engineers does *not* have available reprints or preprints of papers published in the Proceedings of the I.R.E., papers presented at Conventions, or papers presented at Section meetings. Reprints of Proceedings can be obtained *only* if ordered in advance of publication and in quantities of fifty or more copies. However, the following three papers *are* available in reprint form and may be obtained by writing to the Institute.

"Radar," by Edwin G. Schneider, published in the August, 1946, issue of the Proceedings of the I.R.E. Price, $0.50.

"The Presentation of Technical Developments Before Professional Societies," by William L. Everitt, published in the July, 1945, issue of the Proceedings of the I.R.E. Obtainable on request without charge.

"Preparation and Publication of I.R.E. Papers," by Helen M. Stote, published in the January, 1946, issue of the Proceedings of the I.R.E. Obtainable on request without charge.

Please address your inquiries to:
The Institute of Radio Engineers, Inc.
1 East 79 Street
New York 21, N. Y.

It would be appreciated if a large stamped, self-addressed envelope accompanied each request.

Notice

The new I.R.E. television standard, "Standards on Television: Methods of Testing Television Transmitters—1947," is now available. The price is $0.75 per copy, including postage to any country.

Orders may be sent to The Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y., enclosing remittance and the address to which copies are to be sent.

Institute News and Radio Notes

Board of Directors

October 8, 1947

Final Report of Office Quarters Committee. Mr. Heising, Chairman of the Office Quarters Committee, presented the final report of the Committee, a copy of which had been distributed. Following a discussion, Dr. Everitt moved that the report of the Office Quarters Committee, dated October 6, 1947, be accepted and that the Board express its deep appreciation to Chairman Heising and the Committee which he so ably led, for so successful a completion of their task. (Unanimously approved.)

Report of Planning Committee. Chairman Heising presented the report of the Planning Committee, dated October 7, 1947, a copy of which had been distributed. This report set forth the plan of having the Institute members belong to technical divisions called "Groups" according to the members' interests, and outlined the organizational plan for these groups. Following a discussion, Mr. Lack moved that the Board accept the recommendation of the Planning Committee and direct the Planning Committee to formulate and submit a concrete plan to inaugurate the group system. (Unanimously approved.)

Proposed Special Member Bylaw. Dr. Shackelford, Chairman of the Constitution and Laws Committee, reported the result of the survey made of Board members regarding the proposed Special Member Bylaw. After discussion, the following actions were taken:

a. *Age Requirement.* Mr. Pratt moved that the age limit for the Special Member Grade, be 32 years. (Approved.)

b. *Procedure for Approval of Special Member.* Mr. Lack moved that the name of a candidate for Special Member, presented at any Board meeting, shall be voted on at the following meeting of the Board, and that the proposed Special Member shall be invited to become a Special Member if two-thirds of the Board members present vote in the affirmative. (Unanimously approved.)

Executive Committee

October 7, 1947

Mr. Lack moved that the Office Quarters Committee be discharged with an expression of grateful appreciation on the part of the Executive Committee. This was unanimously approved. Mr. Lack also reported that six technical committee meetings were held during the month of September and a number of tentative definitions issued. Technical Committee work is going forward with energy. It is expected that a large number of standards and new definitions will be brought out in 1948.

Dr. Goldsmith moved that the educational directory, part III, Office of Education, Federal Security Agency, Washington, D. C., be used for determining the standings of colleges and universities as applied to the schools of recognized standing, as required by the I.R.E. bylaws, with the understanding that schools already approved by the Board of Directors not appearing in this directory, shall continue to have the approval of the Institute, and that accredited representatives be accepted from all schools included in the above categories. This was unanimously approved.

It was moved by Mr. Henney that the National Chiao-Tung University, Shanghai, China, be approved as a school of recognized standing. This received unanimous approval.

The final report of the 1947 National I.R.E. Convention Committee was sent to the members of the Executive Committee with the suggestion that it be used as a guide for future convention committees. It was proposed that a letter of thanks be sent to Mr. Bassett of the Sperry Gyroscope Company, thanking him for his co-operation and pointing out the magnificent job done by Dr. J. E. Shepherd.

The following candidates were unanimously approved for membership on the I.R.E. Nuclear Studies Committee: W. R. G. Baker, R. M. Bowie, chairman (alternate: P. R. Bell), W. F. Davidson, J. B. Fiske (alternate: L. E. Rassmussen), H. H. Goldsmith, Andrew Haeff, Keith Henney, M. M. Hubbard, W. H. Jordan, Thomas Killian, Serge Korff, R. A. Krause, J. B. H. Kuper, F. R. Lack, W. K. Parsons (alternate: C. B. Laning), J. E. Rose (alternate: F. R. Shonka), S. M. Van Voorhis, John Victoreen, R. S. Warner, Jr., J. R. Weisner.

I.R.E. Subsection for Northern New Jersey

The Northern New Jersey Subsection of the New York Section of The Institute of Radio Engineers was formally organized Wednesday, October 8, 1947, in the Boonton, N. J., High School auditorium. J. E. Shepherd, Chairman of the New York Section, spoke to the group and informed them that the objective was in line with the I.R.E. policy of greater decentralization. The new subsection would make it more convenient for the more than 1000 radio engineers in the northern New Jersey area to attend the meetings. It would also increase the number of papers that could be delivered in the course of the year and improve the opportunity for discussion after the papers. Mr. Shepherd appointed Jerry B. Minter, Chief Engineer of Measurements Corporation, Boonton, N. J., Chairman of the Subsection. John H. Redington of Technical Devices, Roseland, N. J., Vice-Chairman, and A. W. Parkes, Jr., of Aircraft Radio Corporation, Boonton, N. J., as secretary, pending formal elections at a later date. The I.R.E. Executive Committee responsible for the new organization consists of three appointed officers plus Murray G. Crosby of Paul Godley, Inc., C. J. Franks, Consulting Engineer, H. W. Houck of Measurements Corporation, W. D. Loughlin of Boonton Radio Corporation, John F. Morrison of Bell Telephone Laboratories, Inc., and A. G. Richardson of Federal Telecommunication Laboratories, Inc.

J. R. Pierce of Bell Telephone Laboratories described recent developments in the traveling-wave tube, a new electronic tube

Temporary Executive Committee of the Newly-Organized Northern New Jersey Subsection of the New York Section of The Institute of Radio Engineers

Top row, left to right: H. W. Houck, C. J. Franks, John H. Redington.
Bottom row, left to right: John F. Morrison, Jerry B. Minter, A. W. Parkes.

capable of amplification in one of its models of 20 decibels at 4000 megacycles. The tube is expected to be valuable in the problem of sending television signals over long distances by radio relays. Radio engineers throughout the world are watching its development with considerable interest.

Administration of Research Conference

A conference on the administration of research, sponsored by the school of engineering of Pennsylvania State College, was held at this institution on October 6 and 7, 1947.

Two hundred leaders of research management from industrial, educational, and governmental laboratories in the United States and Canada registered for the conference. The first meeting was called to order by Dean Hammond of the school of engineering, who introduced Dr. J. A. Hutcheson (A'28–M'30–SM'43), associate director of the research laboratories of Westinghouse Electric Corporation. Among the speakers, Mr. Maurice Holland presented the subject, "The Place of Research in the Corporate Structure," and Dr. R. L. Jones spoke on "Organization by Scientific Division." This was followed by a paper by Dr. G. H. Young, "Organization by Individual Projects."

Dr. Philip M. Morse, director of Brookhaven National Laboratories, led the afternoon meeting on October 6. The speakers were: Dr. Dwight E. Gray, Dr. L. Warrington Chubb, and Dr. Edward U. Condon. Col. Leslie E. Simon, director of Ballistic Research Laboratories, Aberdeen Proving Ground, spoke at the Monday evening dinner on "German Research in World War II" from personal investigation of German research laboratories made since the war.

Speakers on October 7 were Dr. Jesse E. Hobson, Commodore Henry A. Schade, and Dr. Blaine B. Wescott. Dr. Paul D. Foote used Dr. Wescott's presentation as a basis for a further discussion of Analyses of Research Costs. "Selection and Training of Research Personnel" was the subject of a paper by Dr. Albert W. Hull.

Proceedings of this conference will be published. Those who desire a copy of these proceedings may order it from Prof. Kenneth L. Holderman, Pennsylvania State College, State College, Pa., Price $3.00.

West Coast I.R.E. Convention

Climaxed by a banquet at the Rose Room, Palace Hotel, San Francisco, at which Frederick E. Terman, Past President, I.R.E. was guest speaker, the postwar West Coast I.R.E. Convention was brought to a successful conclusion on September 26, 1947.

With an attendance of 753, including not only West Coast members, but engineers from all parts of the United States, some twenty-five papers were presented which covered the general field of radio and electronics. The convention, which was held September 24, 25, and 26, was effectively supplemented by exhibits of the West Coast Electronic Manufacturer's Association which were open to I.R.E. convention registrants.

A varied program of papers covered the subjects of frequency modulation, instrumentation, television, electronic devices, and the application of electronics to military needs. Important developments in the use of electronics in the field of atomic energy were discussed by representatives of the University of California and Stanford University. The military described some of the technical problems of communication with which they are presently confronted and, among other things, reported on the telemetering of guided missiles.

Frequency modulation was covered by several papers which discussed the problems of detection and interference of signals, and also the generation of high power at the frequencies presently allocated. A newly-developed method of monitoring f.m. stations was also described which measures the mean frequency of the carrier, distortion, frequency response, and provides for over-modulation alarm.

Some of the varied applications of electronic tubes and circuits were described including a method to determine the velocity of a shell as it leaves the gun barrel, and the detection of flaws in metal castings and forgings, both by methods reminiscent of radar techniques applied during the war.

The Bell Telephone Laboratories presented a report on the progress of their New York to Boston radio relay experiment. Operating in the 3700- to 4200-megacycle band, seven repeater stations are spaced about thirty miles apart, between the two terminals. With a radiated beam width of but a few degrees, two two-way channels will be provided and will be capable of accommodating several hundred telephone conversations, or a television broadcast in each direction.

It is hoped that some of the papers presented before the convention will be published in future issues of the Proceedings.

A number of inspection trips were included in the convention program. Visits to Naval installations, the University of California cyclotron, laboratories of Stanford University, National Advisory Committee for Aeronautics, Eitel-McCullough, and Radio Stations KWID and KWIX were well attended and offered opportunities to observe their general operation.

An interesting program of entertainment for the ladies included a welcoming tea, sightseeing tours, and radio broadcast.

NAB Holds Engineering Conference and Roundtable

A day-long engineering conference marked the opening of the National Association of Broadcaster's annual convention which took place at Atlantic City on September 15. Judge Justin Miller, president of the NAB, gave the welcoming address. Both industry and government presented papers.

The morning session opened with a discussion entitled, "Recent Television Development," with particular emphasis on photography of kinescope images, and with a description of the Washington and New York NBC television stations. Paul A. de Mars, f.m. pioneer, spoke on "Frequency Modulation Broadcast Station Construction." The final paper of the morning brought John D. Colvin, audio facilities engineer of ABC, to the podium with an illustrated talk on "Audio Consideration for Broadcast Stations."

One of the major problems facing engineers in modern radio allocation was brought to the fore in the afternoon session when Dixie B. McKey presented his lecture on "Directional Antennas, Their Care and

The Speaker's table at the West Coast I.R.E. Convention banquet, held in the Rose Room at the Palace Hotel, San Francisco, on September 26, 1947. Left to right: Laurence G. Cumming, technical secretary, I.R.E., Earl Scott, Portland, Oregon Section, I.R.E., Wallace Wahlgren, president, West Coast Electronic Manufacturer's Association, Captain Rawson Bennett, chairman, San Diego Section, I.R.E., Dr. F. E. Terman, Past President I.R.E., dean of engineering, Stanford University, Professor Karl Spangenberg, convention chairman, department of electrical engineering, Stanford University, Rear Admiral J. R. Redman, U.S.N., deputy commander, Western Sea Frontier, Col. L. C. Parsons, signal officer, Sixth Army Presidio, San Francisco, George W. Bailey, executive secretary, I.R.E., and Bernard Wailey, secretary, Los Angeles Section, I.R.E.

Maintenance." George P. Adair, former chief engineer of F.C.C. and a radio engineering consultant in Washington, spoke on the "Technical Regulation of Radio." He included in this paper the problem of operator licensing requirements.

The final session was the F.C.C.-industry engineering roundtable. The commission representatives, headed by chief engineer George E. Sterling, were: Dr. John A. Willoughby, assistant chief engineer; James E. Barr, chief, standard broadcast division; Cyril M. Braum, chief, f.m. broadcast division; and Curtis B. Plummer, chief, television broadcast division. Those appearing on the roundtable answered regulatory engineering problems presented by the engineers in attendance.

Annual Report
June 7, 1947

Canadian Council of Institute of Radio Engineers Standing Committee on "Membership and Admission Standards"

1. The Committee was formed by Dr. F. S. Howes, Chairman of the Council on July 8, 1946, to give consideration to our standards of membership and admission.

The Committee is composed of the Chairmen of the Membership Committees of the individual sections: J. A. Collins, Montreal; R. A. H. Galbraith, Ottawa; H. Langford, London; W. F. Choat, Toronto; and F. H. R. Pounsett, Chairman.

2. No meetings were held during the year, all business being carried on by correspondence. The co-operation of the members has been very much appreciated by the chairman and a considerable number of points have been covered.

3. *Admission Standards*

3.1 Admissions Committee Manual, November 7, 1945. In general, the definitions and explanations in this manual issued by Headquarters regarding grades of membership and the required qualifications for same appear to be as complete as can be expected, considering the wide field covered by the Institute and the multiplicity of types of individuals from which we recruit our membership. (At least one member of the Committee was not aware of the existence of this manual.) This manual has helped considerably to clarify several points which were previously in doubt regarding qualifications for membership, but the following items could be noted with regard to their application in Canada.

3.2 *Physicists.* Physicists who have training and professional experience in radio or allied fields are qualified. The Institute is not now limited to engineers by the very wording of our "Aims and Objects."

3.3 A teacher, to qualify, must have taught in a school of recognized standing and not in a trade school, nor should we accept teaching experience in a military or war emergency school.

3.4 *Schools of recognized standing.* The list given in the Admissions Manual includes only American Institutions and one, added by the Board of Directors, in South Africa. For purposes of professional standing and teaching experience we recommend the addition of the following:

- Nova Scotia Technical College
- Dalhousie (physics)
- University of New Brunswick
- Laval University
- McGill University
- Ecole Polytechnique
- University of Montreal (physics)
- Queens University
- University of Toronto
- University of Western Ontario (physics)
- University of Manitoba
- University of Saskatchewan
- University of Alberta
- University of British Columbia
- McMaster University (physics)
- Sir George Williams College (physics)

Professorship in St. Mary's College, Acadia College, and St. Francis Xavier College, is considered adequate professional standing for admission to the Institute grades.

Registration in one of the eight professional engineering bodies in Canada is considered adequate standing for admission to the Institute grades.

Calendar of
COMING EVENTS

1948 I.R.E. National Convention
March 22–25, 1948.

Action by the Council is requested in order to ratify all or any of the above sixteen schools so that we may formally advise Headquarters.

3.5 *Professional Associations.* It is recommended that membership in one of the Provincial Professional Engineering Associations registering bodies be considered equivalent for the purposes of admission to the Institute, to graduation from a school of recognized standing.

3.6 *Examinations.* As far as can be ascertained, no engineering associations in Canada require examinations for admission.

4. *Membership*

4.1 *Up-grading.* A drive to up-grade membership has been carried on in all sections, but the results have been none too encouraging. The Committee feels that it is most important to raise the professional level of our Institute, but it must also be borne in mind that up-grading should be compatible with the necessary qualifications. It appears that one sound method of raising the level is the careful selection of new members. Our Associate grade is practically wide open, whereas this is not the case in all other engineering societies.

4.2 The Membership status of the Sections as of May, 1947, is given below; the up-grading is also shown.

F. H. R. Pounsett
Chairman

	F	SM	M	VA	A	S	Total	
Ottawa	2	7	22	5	51	16	103	Decrease 4 Upgrading by transfer to SM 1 to M 1
London	0	4	8	1	69	75	157	Increase 32
Ontario*	3	23	31	21	209	78	365	Increase 35 Upgrading by transfer SM to F 2 A to SM 1 A to M 2
Winnipeg S. Section		1	3	1	22		27	
BC, Alta and Sask.		2	8	5	62	19	96	
Montreal		12	37	26	91	37	203	Decrease 16
							951	Net gain of 61 for year 1946–1947.

* Including Toronto Section and Hamilton Subsection.

Industrial Engineering Notes[1]

Army Electronic Research Furthered by Industry and Colleges

Radio and electronic research is being carried on by the following industrial concerns and universities in connection with the Signal Corps' broad research program:

Columbia University: research in connection with the generation and control of electromagnetic radiation in the centimeter and millimeter regions of the spectrum.

University of Michigan: study of continuous-wave and pulsed magnetrons for communication purposes.

General Electric Company: high-power continuous-wave magnetrons.

Sylvania Electric Products, Inc.; tunable continuous-wave magnetron tube and low drain secondary-emission amplifier tubes and filamentary alloys for electron tubes.

Purdue University: semiconductors for use as rectifiers.

Westinghouse Electric Corporation: cold-cathode signaling lamp.

Galvin Manufacturing Corp., (Motorola,

[1] The data on which these NOTES are based were drawn, by permission, from "Industry Report," issues of October 3, and 10, 1947, published by the Radio Manufacturers Association, whose courteous co-operation in this matter is gratefully acknowledged.

Inc.): intermediate-frequency systems with a high degree of stability and minimum band pass.

Philco Corporation: investigation of frequency-modulated detector circuits and automatic relaying in radio relay circuits.

Armour Research Foundation: techniques and methods for producing improved microwave equipment surfaces.

Polytechnic Research and Development Company, Inc.: wave-guide mode filters and broad-band couplings for wave-guide application.

DeMornay-Budd, Inc.: lightweight wave guides.

Sperry Gyroscope Company, Inc.: broad-band wave guides and wave guide components in the "X" and "K" frequency bands.

Northern University: wave-guide mode filters.

Ohio State University: technique of using models to determine the characteristics of low-frequency antennas.

Washington University: diversity-type antenna systems.

Federal Telecommunications Laboratories, Inc.: thermosetting molding plastics and mold inhibitor for plastics and rubbers.

General Research Laboratories: thin self-supporting insulation films for use in capacitors.

Radio Corporation of America: high-speed facsimile.

Stromberg-Carlson Company: magnetic recording systems.

New Division of National Bureau of Standards

The National Applied Mathematics Laboratories, a new division of the National Bureau of Standards, will include the Institute of Numerical Analysis, the Computation Laboratory, the Statistical Engineering Laboratory, and the Machine Development Laboratory.

Dr. E. U. Condon, director of the National Bureau of Standards, has stated that the aim of the new division is to conduct research, both government and private, and provide services in the field of applied mathematics, substituting relatively inexpensive calculating for the more costly trial-and-error experimentation.

Discussions on Radio-Related Subjects

Printed circuits and NBS casting resin, which were developed by scientists of the National Bureau of Standards, were discussed on October 15 and 16 in Washington, D. C. Eleven technical papers were presented by government and industry representatives. The Navy's Aircraft Radio and Electronics Committee sponsored the printed-circuit meeting on October 15, evaluating the techniques, applications and limitations of printed circuits, and the Standards Bureau sponsored the casting resin symposium on October 16.

The NBS resin, for which the Bureau of Standards claims ruggedness, moisture-proofing, circuit stability, and specialized mechanical and electrical properties, has suggested many peacetime uses. The focusing of industrial interest on the NBS resin prompted the decision to hold the symposium. Among the speakers were Harry Diamond, chief of the Ordnance Development Division, in which the resin was developed, and P. J. Franklin and M. Weinberg, who were active in its formulation.

U. S. Develops Mechanical Mica Splitter

Wartime research is responsible for a mechanical mica splitter which speeds up the processes and reduces the period required to train skilled splitters. A detailed description of the machine was published in the November issue of the "Technical News Bulletin" of the National Bureau of Standards, and may be obtained by sending 10 cents to the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C.

Atomic Physics Division

Dr. E. U. Condon, director of the National Bureau of Standards has announced the formation of a division of atomic physics, including an electronics section and five others, in which the Bureau activities relating to atomic and molecular physics have been grouped.

Functions of the new division include the promotion of fundamental fact-finding research and determination of fundamental standards in the field of atomic physics. Dr. Condon will head the new division, with Dr. Robert D. Huntoon, former chief of the electronics section in the ordnance development division, as assistant chief. The six sections which make up the new division are: Spectroscopy, Electronics, Mass Spectrometry, Radioactivity, X-rays, and Atomic Physics.

Details Available on New German Magnetophone

The office of Technical Services of the Department of Commerce, released additional data on the German Magnetophone. This report, which is now on sale, contains a full description of the amplifier unit, principles of operation, care and handling of tapes, lubrication, and technical and other data (including new features) for the "K4" and "K7" Magnetophones, and includes a list of their parts with circuit diagrams. An appendix contains a description of the manufacturing processes.

Orders for the report (PB-79558; mimeographed, $3.50) should be addressed to the Office of Technical Services, Department of Commerce, Washington 25, D. C., and should be accompanied by check or money order, payable to the Treasurer of the United States.

Supplement For Commodity Specification Directory

The 1945 edition of the National Directory of Commodity Specifications has been enlarged by a supplement issued in September, 1947. The Directory and Supplement combined now list by name, number, and issuing or sponsoring organization, all standards, specifications, and methods of test in general use for commodities produced in or purchased by this country.

Copies of the Supplement and Directory (Miscellaneous Publication M178) may be obtained from the Superintendent of Documents, Government Printing Office, Washington 25, D. C., for $4.00 and $2.25, respectively.

To Replace Cairo Convention

F.C.C. chairman Charles R. Denny headed the delegation of the United States which was one of the 78 signatory nations to promulgate new world radio regulations at Atlantic City this past summer. These regulations cover all phases of international radio communications. The Atlantic City Convention, when ratified by the U. S. Senate and other signatory nations, will replace the Cairo Convention as the radio law of the world.

Signal Corps' 21 Million for Electronics

A breakdown of the Signal Corps' 21-million-dollar fund for electronics is planned as follows: $6,400,000 allocated for field radio, $210,000 for radio and radar equipment for army boats, and $2,000,000 for meteorological equipment. Of its $8,000,000 allotment for research and development 70 per cent will be spent in the electronic field: $1,287,000 for transmitters and diversity receivers in the administrative radio network with an added $1,852,002 for the system's fixed plant. The proposed outlay for electronic parts will be $498,000, and $920,000 for the army airways communication system. Because of the delay in Congressional approval of funds for the 1948 fiscal year, only a small portion of these amounts has thus far been obligated.

84 Million in Navy Electronic Purchases

The Electronic Division of the U. S. Navy has been allocated over 84 million dollars for procurement of radio and electronic equipment, and parts for the 1948 fiscal year. Proposed allotments are as follows: $2,500,000 for ship radio; $8,540,000 for ship radar; $5,000,000 for sonar; $2,260,000 for countermeasures; $1,200,000 for cryptographic and analytical developments; $14,272,000 for the Marine Corps; $7,700,000 for shore radio and radar; $1,200,000, nancy; $300,000 for patents; $9,128,000 for electron tubes; and $11,500,000 for component parts. Under "shore funds," the Division will spend $2,500,000 for shore improvements; $875,000 for schools, and $17,100,000 for investigations and tests.

This program of expenditures, which got under way the latter part of September, 1947, does not include the program of the Bureau of Aeronautics or the Ordnance Bureau, which are receiving separate funds for their radio and electronic needs.

Navy Revises List of Radio Bidders

The Bureau of Ships of the U. S. Navy Department has revised its "master list" of all interested, qualified contractors to give each such contractor an equal opportunity to submit bids for electronics equipment specified by the Electronics Division.

To be placed on the list, contractors must request action and submit proof as to their (1) technical ability, (2) security of plant and personnel (if work on classified contracts is desired), (3) facilities, and (4) financial responsibility.

Airline Radio Equipment Standards

By request of the Civil Aeronautics Administration, minimum standard performance requirements for communications and radio navigational equipment for air carriers will be adopted as a technical standard order. This order will supplement former requirements for individual-type certifications. Manufacturers will not be required to submit any data to the C.A.A., but will have to certify that their products comply with the technical standard order. Study and preparation of this has been placed in the hands of the Radio Technical Commission for Aeronautics, of which RMA is a member.

Television, Mobile Channel Argument

Oral argument on the F. C. C.'s proposal to reallocate certain frequencies in the bands 44–50 and 72–78 megacycles, which has elicited considerable interest on the part of the television broadcasters, general mobile communications operators, and equipment manufacturers, got underway on October 13.

Unique Television Plan In Dallas

Roger Lacey and Tom Potter, Texas oil men, were granted a permit by the F.C.C. on September 11 for the construction of a commercial television station in downtown Dallas where they plan to build a 47-story hotel. Besides being the site for the television station, the hotel will be equipped with a television receiver in each room.

Radio and Television Receiver Production Increase

The month of August marked the first increase in radio and television receiver production since the peak was reached last April. The number of receivers manufactured by the RMA member-companies from January to August inclusive was 11,031,935; f.m.-a.m., 588,226; television, 68,669. Production of television receivers in August surpassed the June record by 719 sets, bringing the total to 12,203 sets for that month.

July Radio Exports

The export figures for radio equipment and parts dropped to $8,862,325 in July, 1947, from an 11-million-dollar mark in June. It rose slightly, however, in point of units shipped, from 6.3 million in June to 6.8 million in July, The accumulated export quantities for the first seven months of 1947 was 55,219,978 items, with a value of $69,049,289.

Test Radio Direction For Television Shows

Under its new construction permit and license for an experimental Class 2 station, the National Broadcasting Company will use industrial, scientific, and medical frequencies for testing radio direction in producing television plays. The directors will receive instructions from the control booth through lightweight receivers.

Broadcasters Urged to Enter Television F.M. Field

Speaking at a National Association of Broadcasters luncheon September 17, F.C.C. Chairman Charles R. Denny outlined some of the unlimited potentialities of television. He urged broadcasters who have not applied for f.m. facilities to "reexamine their position" and to take note of the new Continental Network of f.m. stations as a "spot on the horizon" well worth watching. He then went on to describe an imaginary f.m. set of the future with ten push buttons, four of which might provide established network programs, and two carrying independent programs via f.m., while the other buttons might be labeled classical music, dance music, features, and news. The "news" button could be pushed at any hour of the day to get a 15-minute news summary.

Television and High Speed Facsimile

Brigadier-General David Sarnoff (M'25–F'30), president and chairman of the board of the Radio Corporation of America, Fellow and former Secretary of the I.R.E., addressed the Chicago Council on Foreign Relations on September 12, 1947. In his address, General Sarnoff pointed out that the development of television has brought about a new problem in the field of human rights.

"This extension of television is nearer than most people may realize," he said. "When nation-wide broadcasting began, it was only five years before listeners overseas were picking up the broadcasts, and before long, regularly scheduled international broadcasts became an established fact. Therefore, in looking ahead, we may reasonably expect that international television will follow much the same pattern of progress. In fact, it may develop more rapidly because the foundation is laid by international sound broadcasting. Already the scientific principles and means for world-wide television are known. No technical problem is involved that money cannot solve."

He stressed the awareness of the new human right, "Freedom to look," which he believes will be as important as "Freedom to listen." International television will mean intracontinental connections as well. "Such television," he affirmed, "has broad possibilities in portraying the way of life of one nation to another. For example, discussion in the press or on the radio of a food shortage is one way of imparting information, but to be able to see hungry men, women, and children in breadlines would help more fortunate people to visualize instantly the dire circumstances and basic needs of their fellow man."

Continuing his remarks, General Sarnoff described a novel method of facsimile communication based on the utilization of television principles. He said, "In our lifetime we have witnessed the evolution of international radio in its various forms of service; we have seen the manually operated telegraph key give way to high-speed automatic printers. Words no longer travel at 25 words a minute, but at 600, and next month, for the first time in a public demonstration, a new and revolutionary system of radio communication, 'ultrafax,' capable of handling a million words a minute, will be revealed by the RCA, in Washington, D. C.

"Ultrafax is a combination of radio and television. It is essentially a radio mail bag to flash documents, newspaper pages, letters, maps, drawings, balance sheets, or, in fact, any written message, in any language. It will be received at its destination as an error-free facsimile of the original.

"Nothing else known to man can span the world as fast as a radio wave for it travels with the speed of light: 186,000 miles a second! Ultrafax is capable of transmitting the equivalent of 40 tons of airmail, coast-to-coast, in a single day; a 500-page book in half a minute and a Sunday metropolitan newspaper including the comics in one minute!

"Indeed, the radio of today will not be the radio of tomorrow. The opportunities of radio as we now see them on the international horizon will change with even greater speed than they did when the first feeble transatlantic wireless signal in 1901 served as the thread out of which a global communication system has been woven.

"Today, science makes it possible for radio to serve all parts of the world instantly. Therein lies the greater responsibility for the leaders of all nations to encourage its proper use and to serve the peoples of the world whose yearning is for peace."

302 F.M. Stations and 12 Television Stations

Six new f.m. stations went on the air the early part of September, 1947, bringing the total number of f.m. stations to 302 as of October 2. The new stations are: WKYC, Paducah, Ky.; WTFM, Tiffin, Ohio; WJBY-FM, Gadsden, Ala.; KSFH, San Francisco, Calif.; KRJM, Santa Maria, Calif.; and WJPG-FM, Green Bay, Wis. Other recently established stations are: WKAT-FM, Miami, Fla.; WBAM, New York, N. Y.; WEHS, Chicago, Ill.; KOKY-FM, Keokuk, Iowa; WHCU-FM, Ithaca, N. Y.; WRRN-FM, Warren, Ohio; WRLD-FM, Lanett, Ala.; WXNJ, Greenbrook Township, N. J.; WVAW, Cheviot, Ohio; KSEO-FM, Durant, Okla; WKIL, Kankakee, Ill.; WEAM-FM, Eau Claire, Wis.; KUGN-FM, Eugene, Ore. Conditional grants were issued for f.m. stations to be located at Niagara Falls, N. Y.; Decatur, Ga.; Clayton, Mo.; Washington, Ind.; Flint, Mich. Conditional grants for five more f.m. stations were authorized early in October, and a construction permit for a commercial television station at Boston, Mass., was authorized by the F.C.C.

There are six licensed television stations on the air and six operating under temporary authority. Fifty-six more stations are authorized and under construction, while thirteen are pending before the F.C.C.

August Excise Taxes

The U. S. Bureau of Internal Revenue reported an excise-tax increase of over a half-million dollars for radio sets, phonographs, components, and the like, for August, 1947, as against the same month a year ago. The July, 1947, excise collections, however, topped by more than a million those of August, reaching a figure of $6,450,451.19. The August figure was $5,084,018.07.

Wholesale Radio Sales up for August

The Census Bureau reported that sales by wholesale appliance and specialty dealers during August totaled $8,437,000, bringing the total sales of these wholesalers during the eight months of 1947 to $54,857,000. This was an increase of 74 per cent over the same period in 1946.

Manufacturers' Inventories Show Increased Sales

Preliminary estimates by the Department of Commerce placed manufacturers' sales for August at $13.4 billion, representing a two per cent rise over July. More than two-thirds of this increase was in the durable goods group, sales of which rose to $5.9 billion. Book value of manufacturers' inventories for August increased to an estimated $22.9 billion, or a gain over July of $200,000,000.

Canada-U. S. Customs Curbs Relaxed

Mobile radio transmitting equipment licensed in either the Dominion of Canada or the United States may now enter both countries, subject to the sealing of the transmitter by customs officials at port of entry. The new arrangement, which was announced by the Federal Communications Commission, went into effect the latter part of September. The seal, to be removed at the port of exit, must not be tampered with during the visitor's stay in either country, under penalty of seizure of the vehicle.

RMA Engineering Department Offers Television Aids

A recently published report by the RMA engineering department entitled "Apartment House Television Antenna," offers a solution to the problem of apartment house owners and their tenants who want good television without spoiling the appearance of their residential building. The report, which was prepared by a special subcommittee, headed by W. P. Short of the committee on television receivers, states that "the solution to the problem has been found in a distribution system which uses an antenna or a combination of antennas, amplifier, cables, and an outlet box for each apartment." The antennas are mounted on rooftops located and oriented or sited for best reception when installed. The individual apartments are connected via a low-loss transmission line connected through conduit to the various apartments, and each apartment is equipped with a connection box similar to an ordinary wall outlet. The cost of installing this system is determined by the cost of cable installation. The number of receivers that can be connected to it is practically unlimited since additional amplifiers can be added when required.

The other aid to television offered by the RMA is the Resolution Chart 1946 intended to standardize resolution measurements, and for checking television equipment. Detailed information on this chart is available from L.C.F. Horle, RMA Data Bureau, 90 West Street, New York 6, N. Y.

Industrial Angles At RMA Fall Meetings

Unusual importance was attached to the annual fall meetings of the RMA which were held between October 13 and 16, 1947.

President Max F. Balcom of the Radio Manufacturers Association presided at the meeting of its board of directors on October 15, which was held at the new headquarters of The Institute of Radio Engineers at 1 East 79th Street, New York City, on the joint invitation of I.R.E. President, Dr. W. R. G. Baker, who is also director of the RMA Engineering Department, and of the I.R.E. Board of Directors. The board planned the Association's program of activities for 1947–48. In the promotion of television, it considered a new RMA resolution chart to facilitate television broadcasting transmission and also public reception, as well as to promote production. The transmitter division approved new activities and services for transmitter manufacturers.

Two publications were ready for the New York meetings, a report prepared by the engineering department on television antennas for apartment houses, and a brochure establishing basic standards for school sound-recording and playback equipment. The latter, prepared by a joint RMA and U. S. Office of Education committee, is expected to promote sales of this apparatus to schools and other markets.

On October 16 the executive committee and all section chairmen met under the chairmanship of S. P. Taylor of the transmitter division to discuss intensified projects for the various sections of the transmitter and parts divisions.

RMA Meetings for October 13 and 14

On October 13, the following meetings were held: coil section—chairman, Edwin I. Guthman; metal stampings and metal specialties section—chairman, S. L. Gabel; record changers and phonomotor assemblies section—chairman, Allan W. Fritzsche; special products section—chairman, William R. MacLeod; wire-wound register section—alternate chairman, Roy S. Laird.

On October 14, the following meetings were held: set division executive committee—chairman, Paul V. Galvin; parts division executive committee and section chairmen—chairman, J. J. Kahn.

Repairmen License Bill Opposed by RMA

On October 16 in the New York City Hall a conference was held on proposed municipal legislation to license radio repairmen in New York. City Councilman Stanley H. Isaacs is the author of the bill.

The RMA board of directors participated in the meeting and vigorously opposed the ordinance and also any discrimination in electric rates against television receivers.

The RMA parts division, in co-operation with radio parts distributors, will sponsor experimental clinics for radio servicemen to raise their standards of service and stabilize their business operation.

School Equipment Committee

Organizational changes in the RMA school equipment committee include the formation of a classroom receiver section, with Sidney Jurin of New York as chairman. The school equipment committee consists of: Lee McCanne (A'36–SM'45) of the Stromberg-Carlson Company, as chairman, and A. K. Ward of the RCA Victor Division as vice chairman.

Staff Assistants Appointed

Early in October, Bond Geddes, executive vice-president of the RMA, announced the appointment of Ralph M. Haarlander as staff assistant to S. P. Taylor, chairman of the transmitter division, and the appointment of James D. Secrest, RMA director of publications, as staff assistant to J. J. Kahn, chairman of the parts division.

RMA Activities

The following RMA Engineering meetings were held:

September 19—Subcommittee on Transformers and Reactors

September 25—Subcommittee on Propagation

September 26—Subcommittee on Gas-filled Microwave Transmission Lines

September 29–30—Transmitter Tube Section

September 30—Subcommittee on Electron Tube Sockets

October 8—Committee on Thermoplastic Hookup Wire.

October 14—Subcommittee on Systems Standards of Good Engineering Practice

October 14—Subcommittee on Geiger Counter Tubes

October 15—Subcommittee on Antennas and R.F. Lines

October 15—Committee on Audio Facilities.

October 17—Committee on Vacuum Sealed Devices

October 21—Committee on Sound Systems.

October 21—Committee on Speakers

October 21—Committee on Intercommunicating Systems

October 21—Executive Committee, Sound Equipment Section

October 21—Subcommittee on UHF Television Systems

October 22—Committee on Amplifiers

October 22—Committee on Microphones

October 22—Executive Committee, Sound Equipment Section.

Sections

Chairman		Secretary
. H. Herndon /o Dept. in charge of ederal Communication 11 Federal Annex tlanta, Ga.	ATLANTA December 19	M. S. Alexander 2289 Memorial Dr., S.E. Atlanta, Ga.
. W. Fischer 14 Beechfield Ave. altimore 29, Md.	BALTIMORE	E. W. Chapin 2805 Shirley Ave. Baltimore 14, Md.
V. H. Radford Massachusetts Institute f Technology ambridge, Mass.	BOSTON	A. G. Bousquet General Radio Co. 275 Massachusetts Ave. Cambridge 39, Mass.
. T. Consentino an Martin 379 Buenos Aires, Argentina	BUENOS AIRES	N. C. Cutler San Martin 379 Buenos Aires, Argentina
. G. Rowe 237 Witkop Avenue Niagara Falls, N. Y.	BUFFALO-NIAGARA December 17	R. F. Blinzler 558 Crescent Ave. Buffalo 14, N. Y.
. A. Green Collins Radio Co. Cedar Rapids, Iowa	CEDAR RAPIDS	Arthur Wulfsburg Collins Radio Co. Cedar Rapids, Iowa
Karl Kramer Jensen Radio Mfg. Co. 601 S. Laramie St. Chicago 38, Ill.	CHICAGO December 19	D. G. Haines Hytron Radio and Electronics Corp. 4000 W. North Ave. Chicago 39, Ill.
J. F. Jordan Baldwin Piano Co. 1801 Gilbert Ave. Cincinnati, Ohio	CINCINNATI December 16	F. Wissel Crosley Corporation 1329 Arlington St. Cincinnati, Ohio
W. G. Hutton R.R. 3 Brecksville, Ohio	CLEVELAND December 11	H. D. Seielstad 1678 Chesterland Ave. Lakewood 7, Ohio
C. J. Emmons 158 E. Como Ave. Columbus 2, Ohio	COLUMBUS December 12	L. B. Lamp 846 Berkeley Rd. Columbus 5, Ohio
L. A. Reilly 989 Roosevelt Ave. Springfield, Mass.	CONNECTICUT VALLEY December 18	H. L. Krauss Dunham Laboratory Yale University New Haven, Conn.
Robert Broding 2921 Kingston Dallas, Texas	DALLAS-FT. WORTH	A. S. LeVelle 308 S. Akard St. Dallas 2, Texas
E. L. Adams Miami Valley Broadcasting Corp. Dayton 1, Ohio	DAYTON December 18	George Rappaport 132 E. Court Harshman Homes Dayton 3, Ohio
P. O. Frincke 219 S. Kenwood St. Royal Oak, Mich.	DETROIT December 19	Charles Kocher 17186 Sioux Rd. Detroit 24, Mich.
N. J. Reitz Sylvania Electric Products, Inc. Emporium, Pa.	EMPORIUM	A. W. Peterson Sylvania Electric Products, Inc. Emporium, Pa.
F. M. Austin 3103 Amherst St. Houston, Texas	HOUSTON	C. V. Clarke, Jr. Box 907 Pasadena, Texas
R. E. McCormick 3466 Carrollton Ave. Indianapolis, Ind.	INDIANAPOLIS	M. G. Beier 3930 Guilford Ave. Indianapolis 5, Ind.
C. L. Omer Midwest Eng. Devel. Co. Inc. 3543 Broadway Kansas City 2, Mo.	KANSAS CITY	Mrs. G. L. Curtis 6003 El Monte Mission, Kansas
R. C. Dearle Dept. of Physics University of Western Ontario London, Ont., Canada	LONDON, ONTARIO	E. H. Tull 14 Erie Ave. London, Ont., Canada
C. W. Mason 141 N. Vermont Ave. Los Angeles 4, Calif.	LOS ANGELES December 16	Bernard Walley RCA Victor Division 420 S. San Pedro St. Los Angeles 13, Calif.
O. W. Towner Radio Station WHAS Third & Liberty Louisville, Ky.	LOUISVILLE	D. C. Summerford Radio Station WHAS Third & Liberty Louisville, Ky.

Chairman		Secretary
E. T. Sherwood Globe-Union Inc. Milwaukee 1, Wis.	MILWAUKEE	J. J. Kircher 2450 S. 35th St. Milwaukee 7, Wis.
R. R. Desaulniers Canadian Marconi Co. 211 St. Sacrement St. Montreal, P.Q., Canada	MONTREAL, QUEBEC January 14	R. P. Matthews Federal Electric Mfg. Co. 9600 St. Lawrence Blvd. Montreal 14, P.Q., Canada
J. E. Shepherd 111 Courtenay Rd. Hempstead, L. I., N. Y.	NEW YORK January 7	I. G. Easton General Radio Co. 90 West Street New York 6, N. Y.
L. R. Quarles University of Virginia Charlottesville, Va.	NORTH CAROLINA-VIRGINIA	J. T. Orth 4101 Fort Ave. Lynchburg, Va.
K. A. Mackinnon Box 542 Ottawa, Ont. Canada	OTTAWA, ONTARIO December 18	D. A. G. Waldock National Defense Headquarters New Army Building Ottawa, Ont., Canada
P. M. Craig 342 Hewitt Rd. Wyncote, Pa.	PHILADELPHIA January 8	J. T. Brothers Philco Radio and Television Tioga and C Sts. Philadelphia 34, Pa.
E. M. Williams Electrical Engineering Dept. Carnegie Institute of Tech. Pittsburgh 13, Pa.	PITTSBURGH January 12	E. W. Marlowe 560 S. Trenton Ave. Wilkinburgh PO Pittsburgh 21, Pa.
Francis McCann 4415 N.E. 81 St. Portland 13, Ore.	PORTLAND	A. E. Richmond Box 441 Portland 7, Ore.
N. W. Mather Dept. of Elec. Engineering Princeton University Princeton, N. J.	PRINCETON	A. E. Harrison Dept. of Elec. Engineering Princeton University Princeton, N. J.
A. E. Newlon Stromberg-Carlson Co. Rochester 3, N. Y.	ROCHESTER December 18	J. A. Rodgers Huntington Hills Rochester, N. Y.
E. S. Naschke 1073-57 St. Sacramento 16, Calif.	SACRAMENTO	G. W. Barnes 1333 Weller Way Sacramento, Calif.
R. L. Coe Radio Station KSD Post Dispatch Bldg. St. Louis 1, Mo.	ST. LOUIS	N. J. Zehr Radio Station KWK Hotel Chase St. Louis 8, Mo.
Rawson Bennett U. S. Navy Electronics Laboratory San Diego 52, Calif.	SAN DIEGO January 6	C. N. Tirrell U. S. Navy Electronics Laboratory San Diego 52, Calif.
W. J. Barclay 955 N. California Ave. Palo Alto, Calif.	SAN FRANCISCO	F. R. Brace 955 Jones San Francisco 9, Calif.
J. F. Johnson 2626 Second Ave. Seattle 1, Wash.	SEATTLE December 11	J. M. Patterson 7200—28 N. W. Seattle 7, Wash.
C. A. Priest 314 Hurlburt Rd. Syracuse, N. Y.	SYRACUSE	R. E. Moe General Electric Co. Syracuse, N. Y.
C. A. Norris J. R. Longstaffe Ltd. 11 King St., W. Toronto, Ont., Canada	TORONTO, ONTARIO	C. G. Lloyd 212 King St., W. Toronto, Ont., Canada
O. H. Schuck 4711 Dupont Ave. S. Minneapolis 9, Minn.	TWIN CITIES	B. E. Montgomery Engineering Department Northwest Airlines Saint Paul, Minn.
R. M. Wainwright Elec. Eng. Department University of Illinois Urbana, Illinois	URBANA	M. H. Crothers Elec. Eng. Department University of Illinois Urbana, Illinois
L. C. Smeby 820—13 St. N. W. Washington 5, D. C.	WASHINGTON January 12	T. J. Carroll National Bureau of Standards Washington, D. C.
J. C. Starks Box 307 Sunbury, Pa.	WILLIAMSPORT January 7	R. G. Petts Sylvania Electric Products, Inc. 1004 Cherry St. Montoursville, Pa.

SUBSECTIONS

Chairman		Secretary
J. D. Schantz Farnsworth Television and Radio Company 3700 E. Pontiac St. Fort Wayne, Ind.	FORT WAYNE (Chicago Subsection)	S. J. Harris Farnsworth Television and Radio Co. 3702 E. Pontiac Fort Wayne 1, Ind.
F. A. O. Banks 81 Troy St. Kitchener, Ont., Canada	HAMILTON (Toronto Subsection)	E. Ruse 195 Ferguson Ave., S. Hamilton, Ont., Canada
A. D. Emurian HDQRS. Signal Corps Engineering Lab. Bradley Beach, N. J.	MONMOUTH (New York Subsection)	Ralph Cole Watson Laboratories Red Bank, N. J.
A. R. Kahn Electro-Voice, Inc. Buchanan, Mich.	SOUTH BEND (Chicago Subsection) December 18	A. M. Wiggins Electro-Voice, Inc. Buchanan, Mich.
W. M. Stringfellow Radio Station WSPD 136 Huron Street Toledo 4,Ohio	TOLEDO (Detroit Subsection)	M. W. Keck 2231 Oak Grove Place Toledo 12, Ohio
W. A. Cole 323 Broadway Ave. Winnipeg, Manit., Canada	WINNIPEG (Toronto Subsection)	C. E. Trembley Canadian Marconi Co. Main Street Winnipeg, Manit., Canada

I.R.E. People

RUDOLFO M. SORIA

Rudolfo M. Soria (S'38–A'43–M'46) is now associated with the American Phenolic Corporation, Chicago, Ill., as project engineer in charge of special development work on antennas and r.f.-transmission lines.

Mr. Soria obtained the bachelor and master degrees in communication engineering from the Massachusetts Institute of Technology. He was formerly instructor in electrical engineering at Illinois Institute of Technology, where he received the Ph.D. degree in June, 1947.

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M. W. SCHELDORF

M. W. Scheldorf (A'26–SM'46) has joined the Andrew Company as head of the engineering research department.

Mr. Scheldorf, who is the co-inventor of the circular-loop antenna, was born on February 15, 1902, at Westside, Iowa. He received the B.S. degree in electrical engineering from Iowa State College in 1923, and joined the radio department of the General Electric Company at Schenectady, N. Y. From 1930 to 1935 he was with the Radio Corporation of America. He then returned to General Electric, where for the last five years he was a specialist in antennas for the electronics department.

M. W. SCHELDORF

WILLIS LAURENS EMERY

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WILLIS LAURENS EMERY

Beginning with the autumn quarter, Dr. W. L. Emery (A'41–SM'46) will assume his new duties as associate professor of electrical engineering at the University of Utah.

A native of Salt Lake, Dr. Emery received his B.S. degree from the University of Utah in 1936 and was an engineering instructor there for the following two years. He received the M.S. in 1940 and the Ph.D. degrees in 1947 from Iowa State College, where he instructed in electronics, and organized and directed the ultra-high-frequency radio laboratory. In 1942 he was given a leave of absence to become radio engineer at the Naval Research Laboratory, Washington, D. C., where he directed work on countermeasures and investigation of enemy equipment.

Dr. Emery is a member of the American Institute of Electrical Engineers, the American Association for the Advancement of Science, Tau Beta Pi, Phi Kappa Phi, and Sigma Xi. His book, "Ultra High Frequency Radio Engineering," which was published by the Macmillan Company in 1944, was a pioneer in this field.

W. H. DOHERTY

William H. Doherty (A'29–M'36–SM'43–F'44), on the invitation of the Italian National Council of Research, attended the celebration of the 50th anniversary of Marconi's discovery of radio, which was held in Rome, Italy, September 28 to October 5. Mr. Doherty, who is a radio development engineer of the Bell Telephone Laboratories, Inc., presented a paper discussing "Linear Power Amplifiers in American Broadcasting."

Mr. Doherty was born in Cambridge, Mass., on August 21, 1907. He received the B.S. degree in electrical communication engineering from Harvard in 1927 and the M.S. degree in engineering in 1928. From 1928 to 1929 he was research associate, radio section, at the National Bureau of Standards. From 1929 to date he has been connected with the Bell Telephone Company.

In 1937 he received the Morris Liebmann Memorial Prize for his improvement in the efficiency of radio-frequency power amplifiers. This was presented to him during the Silver Anniversary banquet of the I.R.E. held on May 12 of that year in the Hotel Pennsylvania.

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W. H. DOHERTY

Stanley Rosenberg

Stanley Rosenberg

Stanley Rosenberg (A'46) has recently been appointed engineer in charge of purchasing and materials control at the Espey Manufacturing Co., Inc.

Mr. Rosenberg received the B.E.E. degree at the College of the City of New York in 1939. During the war, he held the position of electrical engineer with the United States Signal Corps and specialized in radar test equipment. Before working at Espey, where he was a project engineer in charge of an u.h.f. signal generator manufactured for the Signal Corps, he was associated with the Hub Engineering Company.

Mr. Rosenberg is a member of the American Institute of Electrical Engineers and is a professional engineer licensed by the State of New York.

Orrin E. Dunlap, Jr.

Orrin E. Dunlap, Jr. (M'44–SM'44) was recently elected vice-president in charge of advertising and publicity of the Radio Corporation of America.

Mr. Dunlap, who was chief operator of the Marconi Wireless Telegraph Company of America aboard the S.S. *Octorora* in 1917, served during World War I as a radio operator in the United States Navy. After graduation from Colgate University in 1920, he attended Harvard Graduate School of Business, specializing in advertising and marketing, then joined the staff of the Hanff-Metzger Advertising Agency. A year later, he was invited by Carr V. Van Anda, Managing Editor of the *New York Times*, to organize a radio section and direct the coverage of radio news. He served in this capacity for eighteen years.

In 1940, Mr. Dunlap joined RCA as manager of the department of information, and on January 1, 1944, became director of advertising and publicity.

Mr. Dunlap is the author of numerous books on radio and radio advertising. He was among the first to become a member of the American Radio Relay League and is a Life member of the Veteran Wireless Operators' Association.

P. B. Reed

P. B. Reed (A'30–M'45) was recently appointed field sales manager in the Eastern Central Region for the RCA Victor's engineering products department.

Prior to his appointment, Mr. Reed represented RCA in Washington, D. C. He joined the organization in 1930 and in 1937 became district sales engineer for its Southern Region. During the war, he served ten months with the fourth fleet of the United States Navy, installing and servicing radio, radar, and underwater sound equipment, as well as training personnel in the use of such equipment, and was closely associated with the Bureau of Ships.

Albert E. Hayes, Jr.

The appointment of Albert E. Hayes, Jr., (A'42–M'46) to the full-time post of national emergency co-ordinator to promote and supervise amateur preparedness in supplying disaster communication, has been announced by Francis E. Handy (A'26), communications manager of the American Radio Relay League. Under Mr. Hayes' supervision, selected radio amateurs in each community will call local meetings to establish common operating procedures and drill periods when the hams' personal stations may be mobilized under simulated emergency conditions.

Mr. Hayes was formerly an engineer with the Bendix Radio Corporation. He is a graduate of M.I.T. and has been active professionally in the electric patent field.

Thomas E. Stewart, Jr.

Thomas E. Stewart, Jr.

Thomas E. Stewart, Jr. (A'44) has been named chief of the applied electronics branch of the United States Army Engineer Research and Development Laboratories at Fort Belvoir, Va.

A graduate of Pratt Institute, School of Science and Technology, in New York City, Mr. Stewart was formerly with the Sylvania Industrial Corporation of Fredericksburg, Va. He has been employed by the Army since 1942, and was recently presented the Exceptional Civilian Service Award for his development of metallic, nonmetallic, and underwater mine detectors; a radio explosives detonator; and a barrage balloon flight analyzer.

P. R. Kendall

P. R. Kendall

P. R. Kendall (A'41–M'45) has been appointed regional sales manager for the communications division of Motorola, Inc., for their New York territory.

Mr. Kendall is 33 years old, a graduate of Case School of Applied Science and holds a B.S. degree in electrical engineering. For ten years he operated the Kendall Radio Company in Cleveland, and he is the designer of the Kendall hearing aid for churches. During the war, Mr. Kendall was employed as a designing and testing engineer on airborne and landing-craft radio-communication equipment. Before joining Motorola, Mr. Kendall worked as sales and field engineering manager of Belmont Radio in Chicago. He is a member of the American Institute of Electrical Engineers.

Chester L. Davis

After a residence of twenty years in Washington, D. C., Chester L. Davis (M'24–M'28–SM'43), has taken up the general practice of law at Perry, in his native State of Missouri.

Mr. Davis received the degrees of LL.B., M.P.L., and LL.M. from National University, Washington, D. C., and is a member of the Bar of the District of Columbia, of the State of Missouri, and of the United States Supreme Court. For fifteen years, he was manager of the Patent Department of the Washington Office of the Radio Corporation of America. In 1922, he was associated with the installation of the first water-cooled transmitter for the Signal Corps at Ft. Leavenworth, Kansas, and in 1925 constructed broadcasting station WJAF at Ferndale, Mich. From 1926 to 1927 he was radio instructor at the School of Engineering, Milwaukee, Wis. During his stay in Washington from 1927 to date he was at various times chairman of the Washington Section of The Institute of Radio Engineers, and of the Committee on Legislation of the Patent Section of the American Association, of which he is still a council member. He is a member of the American Institute of Electrical Engineers, the American Patent Law Association, and the American Bar Association.

Books

Ultrahigh Frequency Transmission and Radiation, by Nathan Marchand

Published (1947) by John Wiley and Sons, Inc., 440 Fourth Avenue, New York 16, N. Y. 312 pages+10-page index+x pages. 140 figures. 5¾×9 inches. Price, $4.50.

This is a new book on the theory of transmission lines, antennas, and wave guides. After an initial chapter on the steady-state theory of transmission lines, there is a chapter developing vector analysis. This is followed by a chapter on Maxwell's equations, after which there are chapters dealing with plane waves, radiation, antenna arrays, and wave guides. The book concludes with an interesting elementary chapter on the grounding, matching, and transformation conditions in transmission lines in some important cases.

The book is suitable as a fourth-year college text on the subjects covered. For more advanced students or research workers, the book will be disappointing, since it does not have much material which is not already covered and clearly explained in such books as Ramo and Whinnery's "Fields and Waves in Modern Radio." On the other hand, the point of view is somewhat more practical than that of previous authors on the same subject, and the book should therefore appeal to those students and engineers who found the earlier works too advanced or theoretical.

STANFORD GOLDMAN
Massachusetts Institute of Technology
Cambridge 39, Mass.

Theory and Application of Mathieu Functions, by N. W. McLachlan

Published (1947) by Oxford University Press, 114 Fifth Avenue, New York 11, N. Y. 394 pages+6-page index+ix pages. 49 figures. 9½×6½ inches. Price, $12.50.

This new book, written by a well-known British engineer and author, provides the physicist and engineer with a comprehensive reference on a useful subject.

In the year 1868 the French mathematician Émile Mathieu published an analysis of the vibration of elliptical membranes in which he introduced the linear variable-coefficient differential equation and certain of its solutions which now bear his name. Since that time the Mathieu functions have arisen in the analysis of a variety of problems, among which are the propagation of electromagnetic energy in elliptical wave guides, the diffraction of sound and of electromagnetic radiation by elliptical cylinders, eddy currents in cores of elliptical cross section, certain types of amplitude distortion in dynamic loudspeakers, problems in frequency modulation, and certain types of dynamical systems which are capable of producing subharmonic oscillations. The Mathieu functions are sometimes called "the functions associated with the elliptical cylinder."

The author has written this book for the engineer and physicist, and has, therefore, devoted considerable attention to applications and to worked numerical examples. The first chapter is historical, and is followed by 257 pages of theory and by 100 pages of applications. The applications can be understood after a perusal of only a portion of the sections devoted to theory. In order to make the text continuous, a considerable amount of new material is included. As the Mathieu functions are perhaps one degree greater in complication than Bessel functions, the book is recommended only for those who have a taste for mathematics and who have some acquaintance with advanced calculus, including Bessel functions.

Engineers and applied scientists will undoubtedly find this volume to be a reference work of enduring value.

WALTER C. JOHNSON
Princeton University
Princeton, N. J.

Mathematics for Radio Engineers, by Leonard Mautner

Published (1947) by Pitman Publishing Company, 2 W. 45 St., New York, N. Y. 319 pages+7-page index+vii pages. 138 figures. 6×8¾ inches. Price, $5.00.

This book is designed to review the mathematical concepts which are useful to the radio engineer. It is not uncommon for a practicing engineer to lose facility for handling mathematics which appear in his field and which are included in much of the current literature. The author attempts to collect in a book of moderate size those problems which the reader might encounter. The author, himself an engineer, has succeeded admirably in his choice of material and in his manner of presentation.

Such topics as logarithms, decibel notation, trigonometric functions, complex algebra, calculus, determinants, power series, differential equations, and Fourier series are covered. In all cases the material is well illustrated with its applications to problems in radio engineering. In addition to these illustrative problems, there are numerous problems for the reader to solve, and the correct answers are given in the back of the book. It is thus possible for the reader to drill himself in the fundamentals described in the text.

To the engineer who needs "brushing up" in the mathematical fundamentals of his profession and to those working in the field of radio who may not have had sufficient formal training in mathematics, this well-written book is recommended.

JOHN R. RAGAZZINI
Columbia University
New York 27, N. Y.

Principles of Electrical Engineering, by T. F. Wall

Published (1947) by Chemical Publishing Co., 26 Court Street, Brooklyn, N. Y. 554 pages+8-page index+xi pages. 497 figures. 5½×8¾ inches. Price, $8.50.

The purpose of the author in writing this book is "to present as comprehensively, and in as limited space as may be possible, an account of the basic principles of the science of electrical engineering, a leading idea throughout the book being to place emphasis on the identity of the principles relating to both heavy-current and light-current engineering practice."

No attempt has been made to write exhaustively of such applications as electrical power machinery, communication systems, or electrical measurements, although material in these fields is included. The primary purpose of presenting fundamental principles is closely adhered to, and the author has covered a surprising amount of ground. Chapters are devoted to such a wide range of subjects as electrical units, atomic structure, the electrical field, currents in networks, magnetic materials, electromagnetism, alternating currents, oscillating systems, harmonic analysis, skin effect, transmission lines, and Maxwell's equations.

The material is up-to-date and, for the most part, clearly treated. The emphasis is distinctly on the mathematical, rather than on the descriptive side, and although frequent illustrative examples are introduced, the companion volume, "Electrical Engineering Problems and Their Solutions," by the same author, is helpful in rounding out the text. A good deal of attention is devoted to electrical transients, and this material seems unnecessarily scattered in its placing.

This is evidently not intended as a first course on electrical engineering principles. References are frequently made to later portions of the book: an elementary knowledge of the subject is apparently assumed. The calculus is freely used, and in some chapters a mathematical facility is assumed more in keeping with the abilities of the graduate rather than the undergraduate student.

The book should be valuable as a reference text for the graduate engineer. Material from a wide range of sources is included. Taken together with its companion volume, it should serve as an interesting source book for the teacher of electrical subjects.

FREDERICK W. GROVER
Union College
Schenectady, N. Y.

lectrical Engineering Problems and Their Solution, by T. F. Wall

Published (1947) by Chemical Publishing Company, Inc., 26 Court Street, Brooklyn, N. Y. 307 pages+4-page index+viii pages. 19 figures. 5½×9 inches. Price, $5.00.

This book, which is a companion volume to "Principles of Electrical Engineering" by the same author, gives the solutions of problems suggested in the "test papers" which accompany that volume.

However, the solution of each problem is here given at length with a development of the pertinent theory and with numerical examples. The selected problems illustrate a wide range of engineering applications of great practical importance, especially in the field of transmission.

Although so closely connected, each of the two volumes is independently useful. The style of presentation is the same in each, with emphasis placed on methods of the mathematical analysis.

A reader well grounded in elementary electrical theory and possessing a knowledge of the calculus and differential equations will find much interesting and useful material in these books. The teacher will find in the volume of problems, illustrative material for a course based on "The Principles of Electrical Engineering."

Frederick W. Grover
Union College
Schenectady, N. Y.

Television Primer of Production and Direction, by Louis A. Sposa

Published (1947) by McGraw-Hill Book Company, Inc., 330 W. 42 St., New York 18, N. Y. 195 pages+11-page index+4-page glossary+3-page appendix+1-page bibliography+x pages. 108 figures. 5½×8 inches. Price, $3.50.

The general public thirsts for more information about the "magic" of television. Probably the "magic" of those thousands of jobs for the inexperienced has something to do with it. What they want to know is not about the dry, technical knowledge that the television engineer must possess in order to put good pictures on the air, but the glamorous business of producing television shows, a business that embraces the intriguing domains of stage, movie lot, and sound-broadcast studio.

Few have the opportunity to visit a busy television studio and still fewer have the chance to learn the steps necessary for a successful television production, from writing the script to the final "fade-out." But from the pages of "The Television Primer of Production and Direction" the reader can get the impression of just how this is done, and in simple, easily understood terms.

Some of the subjects covered are: Lighting, Scenic Design, Titles, Costuming, Make-up, Microphones and Sound, Motion Picture Film, Scripts, Commercials, Production, Directing, Programming. The author has treated each of these important subjects in a very satisfactory manner. Naturally some limitations will be noticed, such as rather brief treatment of some difficult subjects due to space limitations; the author describes principally his experiences obtained at only one station, WABD; and in this new art it must not be forgotten that methods and techniques can change overnight.

The opening chapters, dealing with the technical portion of the television system and the camera, are not up to the standard of the remainder of the book.

The author is at his best when he writes of the field in which he works, television directing. In spite of the modest name of "Primer," don't think this book will not be read by all the professionals. Why? Because it contains ideas, well-organized ideas which Mr. Sposa has found by trial and error really work and produce results. And in television producing, what is more valuable than ideas?

Albert F. Murray
Consulting Television Engineer
Washington, D. C.

Getting a Job in Television, by John Southwell

Published (1947) by McGraw-Hill Book Co., 330 W. 42 Street, New York 18, N. Y. 113 pages+5-page index+ii pages. 6 illustrations. 5½×8 inches. Price, $2.00.

This concise little book attempts to give the answers to the questions that are asked repeatedly of everyone remotely connected with television, and insofar as those questions are answerable it succeeds in its attempt.

Mr. Southwell lists the jobs that television makes available; from director to stagehand, from consulting engineer to technician. He gives the basic skills, education, and training that are necessary in each job, the maximum and minimum salaries that each commands (without too much emphasis on the maximum), the chances and lines of advancement in each. He lists the guilds and unions that claim jurisdiction in each category; the television stations operating, under construction, or applied for; the advertising agencies handling television shows, with names and addresses.

The author is conservative in his treatment, and gives little encouragement to the starry-eyed believer that television offers a royal road to fame and fortune. He emphasizes the amount of work and knowledge necessary to fill even the humbler positions.

Seekers for sinecures may not welcome the book for that very reason, but a young man planning a career should find it useful, and those who are constantly asked about television jobs should find it a godsend.

Donald K. Lippincott
Patent Attorney
San Francisco, Calif.

Electronic Engineering Master Index, 1925–1945, Part II, and Electronic Engineering Master Index, 1946, edited by Frank A. Petraglia

Part II. 1935–1945. Published (1946) by The Macmillan Company, 60 Fifth Avenue, New York, N. Y. 202 pages+7-page index+viii pages, 7×10½ inches. Price, $6.00.

1945–1946. Published (1947) by Electronics Research Publishing Company, 2 W. 46 St. New York 19, N. Y. 162 pages+10-page index+30-page bibliographies of engineering texts and trade literature+x pages. No figures. 6½×9½ inches. Price, $14.50.

The Electronic Engineering Master Index is a bibliography of references to periodical literature, arranged under an alphabetical subject classification extending from Acoustics, Adjacent Channel Interference, Aerials, and Aeronautical Radio, to Wide-Band Amplifiers, X-rays, Yagi Array and Zirconium. Each item is two or three lines long and gives the title and specific citation of the published paper. All of the technical articles in the leading electronic periodicals (including the Proceedings of the I.R.E.) are listed, as well as selected articles from about forty other periodicals in aeronautical, chemical, electrical, and general industrial fields.

For the years 1925–1945, the Index has been issued in two volumes. Part II, for 1935–1945, referred to above, contains approximately 10,000 entries.

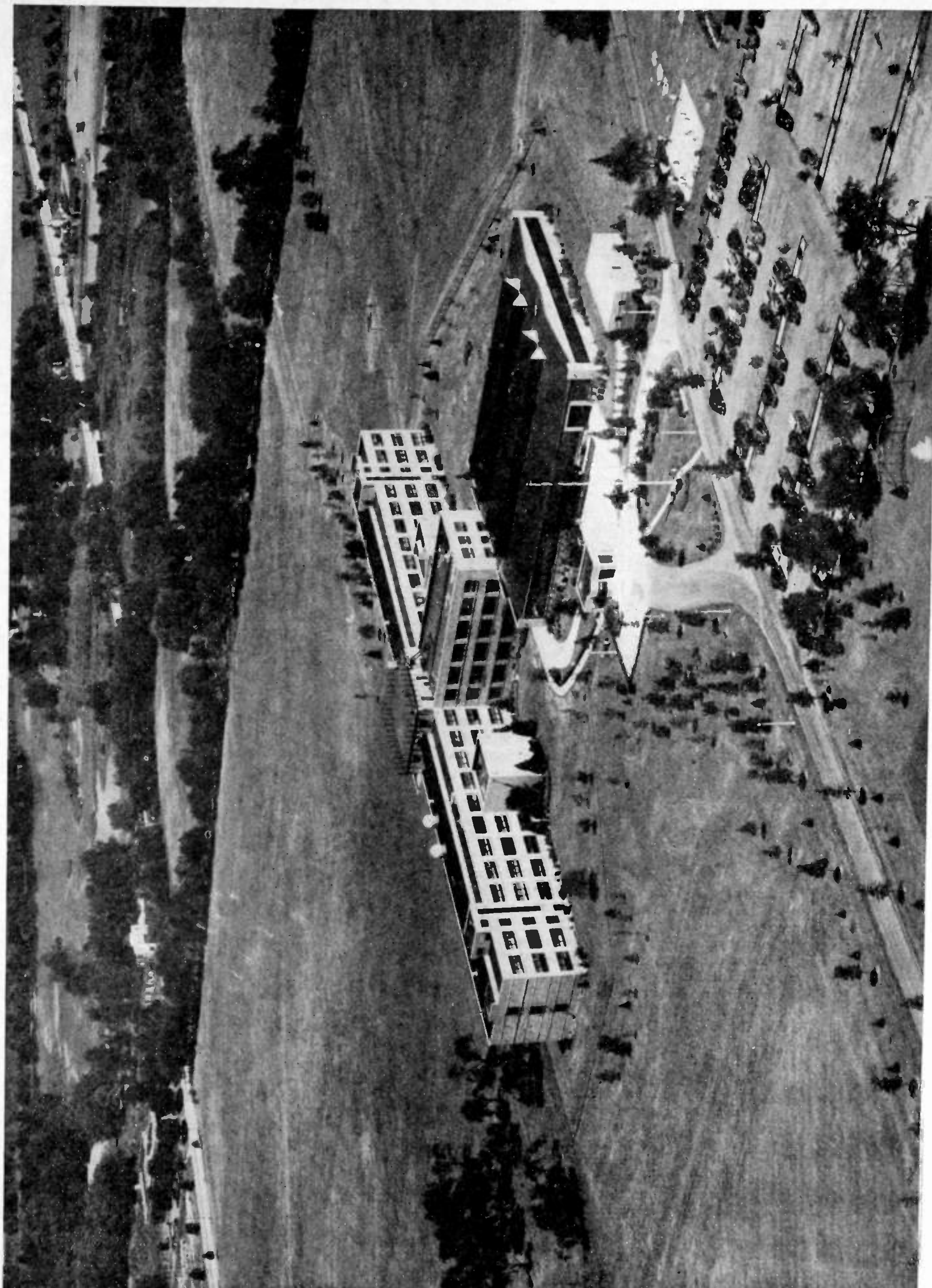
The supplement, covering the period from July, 1945, through December, 1946, contains about 7500 new entries. It includes two new sections, one giving a "Bibliography of Engineering Texts," and the other a survey of trade literature issued during the latter half of 1946.

The comprehensive way in which the field is covered would seem to make this Index very useful to one who wishes to have at hand references to literature on electronic subjects which are published in periodicals not usually included in the bibliographies normally appearing in the Proceedings of the I.R.E.

Laurens E. Whittemore
American Telephone and Telegraph Co.
New York 7, N. Y.

Books for Finland

The Institute of Technology, Helsinki, Finland, will welcome gifts of scientific and technical books and periodicals to take the place of those destroyed and thus to reduce a very serious handicap of Finnish scholars. Any such gifts should be addressed to the Institute of Technology, Helsinki, and sent to the Legation of Finland, 2144 Wyoming Avenue, N. W., Washington, D. C. Their shipment to Finland will be arranged by the Finnish Minister.



RCA LABORATORIES, RESEARCH HEADQUARTERS

Airplane view of the modern RCA research laboratory in the university town of Princeton, New Jersey.

Officers: Dallas-Ft. Worth Section

J. G. Rountree (A'39–M'44) was born in Bee County, Texas, on January 7, 1914. He received the B.A. degree with honors from the University of Texas in 1937, having majored in physics. During his senior year, he was employed by broadcast station KNOW, Austin, Texas, and on graduation, he entered the employ of KTSA,

J. G. ROUNTREE
CHAIRMAN

San Antonio, Texas. In 1939, he was employed by WBAP, Fort Worth, and in September, 1941, he joined the field disivion of the engineering department of the Federal Communications Commission as radio inspector.

During 1942 and 1943, he was attached to Headquarters New Orleans Air Defense Region as a civilian liaison officer. During the summer of 1945, he was in charge of the monitoring station established in Montgomery, Alabama, for the purpose of making and analyzing v.h.f. field strength recordings as a part of the v.h.f. field intensity survey. Since April, 1946, Mr. Rountree has been associated with the consulting engineering firm of A. Earl Cullum, Jr. Mr. Rountree has been active in amateur radio circles since 1932, holding a license for amateur station W5CLP.

Robert A. Broding (S'39–A'40–M'44–SM'47) was born November 1, 1916, at Foley, Minn. He was graduated from the University of Minnesota with a B.E.E. degree in 1939, and operated the radio station at the University, then WLB. In December, 1939, Mr. Broding joined the geophysical department of the Magnolia Petroleum Company and spent the following year with a seismic crew prospecting for oil in northeast Ohio. He was then transferred to the Geophysical Laboratories in Dallas.

During the war, leave of absence was obtained and he worked as a civilian employee at the Naval Ordnance Laboratory in Washington, D. C. Here development work was done on electronic firing devices for magnetic mines. In 1943, Mr. Broding returned to the Field Research Laboratories of the Magnolia Petroleum Company in Dallas. As a senior research physicist, he has since been employed in research and development on electronic control instruments and electrical methods for oil prospecting.

ROBERT A. BRODING
VICE-CHAIRMAN

He has served on several committees of The Institute of Radio Engineers, including Section Vice-Chairman and Chairman of the Meetings and Papers Committee in 1946, becoming Section Chairman in 1947.

New Television Field-Pickup Equipment Employing the Image Orthicon*

JOHN H. ROE†, ASSOCIATE, I.R.E.

Summary—A brief review of the characteristics of the more widely used types of electronic television pickup tubes traces the trend toward greater sensitivity, culminating in the image orthicon. This development results in the present-day ability to televise an unlimited variety of subject matter. Former restrictions imposed by requirements for large amounts of illumination have been almost entirely removed. An important by-product of higher sensitivity is the possible increase in depth of focus of the optical system.

Field or portable equipment has been designed to take advantage of the improved characteristics of the image orthicon. It is a design which lends itself to a maximum of flexibility for various types of operation, including use in studios and in mobile units. Most of the units are shaped like a medium-sized suitcase. The camera includes a four-position lens turret and an electronic view finder. Camera cables may be as much as 1000 feet long. Electrical interconnections are simple and few in number. Each of the major units is described in some detail, along with its function in the system. Discussion of some of the unusual circuits is included in the Appendix.

Introduction

IN EVERY ART, advances occur at intervals which serve as distinct milestones in the progress of that art. They are steps which overcome major limitations, and thus open up new fields which men have only dreamed about before. Such an advance has recently occurred in the art of television in the development of the image-orthicon pickup tube.

Television has made much progress in the past two decades in such things as higher definition, greater picture brilliance and size, greater immunity to interference in transmission, improved techniques in propagation, and the introduction of color on a laboratory scale. However, the requirement for intense illumination of the televised scene has dogged the industry from its inception up to the very recent past. This requirement has limited outdoor pickups to daylight hours with bright sunlight, and indoor pickups either to motion-picture film or to studios where enormous amounts of lighting on the order of 1000 to 1500 foot-candles could be provided.

The lighting equipment for such studios not only represents a large capital investment, but it entails excessive operating expense. Costly air-conditioning systems only partially alleviate the discomfort of performers, who literally have to "sweat it out" in scenes that cannot be retaken if things do not go right the first time. From the producer's point of view, such intense lighting produces flat, shadowless, uninteresting effects which greatly limit the artistic possibilities of the medium.

These conditions are always attendant on operation with the iconoscope as a pickup tube. The iconoscope itself is one of television's milestones because it introduced the storage principle to the art, made the system all-electronic, and thus brought television into a form which has commercial possibilities. It represented a big stride in sensitivity over previous nonstorage devices. However, its lack of sufficient sensitivity to operate satisfactorily outdoors in cloudy weather or in late-afternoon dusk, or indoors under moderate lighting, has been, and still is, its principal limitation.

The next step in the direction of greater sensitivity was the introduction in 1939 of low-velocity scanning in the RCA-1840 orthicon-type of pickup tube. It retained the storage principle and added a great improvement in efficiency with a corresponding improvement in sensitivity of the order of five times. This meant the possibility of reducing incident illumination to about 200 or 300 foot-candles.

Wartime development of military television equipment[1] accelerated work on a pickup tube which had its beginnings before the war started. The result of this work we know today as the image orthicon, a pickup tube which embodies the old principles of storage and low-velocity scanning, and, in addition, the principles of image-electron multiplication and signal-electron multiplication. The tube and the theories underlying its operation and incorporation into television cameras have been described in detail in recent literature.[2]

The image orthicon has as its most outstanding characteristic very great sensitivity, of the order of 100 times greater than that of the iconoscope. One of the most obvious and useful results of the high sensitivity of the tube is that, under medium or high illumination, the lens opening may be stopped down to a very small size, thus giving an enormous depth of focus. Even under relatively low illumination, the depth of focus of the image orthicon is much greater than that obtainable with less-sensitive tubes.

In contrast with the simple orthicon, the image orthicon has another outstanding characteristic; namely, its ability to accommodate a tremendous light range without serious loss of contrast. The scene illumination may be changed from dark shadows to bright sunlight and back again without losing essential picture information.

* Decimal classification: R583.6. Original manuscript received by the Institute, August 25, 1947.

† Radio Corporation of America, RCA Victor Division, Camden, N. J.

[1] A series of papers on military television developments appeared in *RCA Rev.*, vol. 7, September and December, 1946.

[2] A. Rose, P. K. Weimer, and H. B. Law, "The image orthicon—a sensitive television pickup tube," Proc. I.R.E., vol. 34, pp. 424–432; July, 1936.

Other important characteristics are: (a) small target size, (b) small over-all tube size, and (c) high output signal level.

The small target area makes it possible to use relatively small lenses which lend themselves to a reasonable turret design. Lenses for such a field are readily available in a variety of focal lengths and apertures. The small size of the image orthicon is a factor of great importance in making the camera itself as compact and light as possible.

All previous types of standard pickup tubes have such low signal outputs that very high-gain amplifiers are required where shot noise in the first stage limits the signal-to-noise ratio. The image orthicon, in contrast to these, produces a high signal output, so that a comparatively low-gain amplifier may be used. Hence, shot noise in the amplifier is very low, compared with noise in the beam.

These characteristics have opened up a wide field of opportunities in television programming, such as night games under standard incandescent lighting, daytime athletic and other events lasting into late-afternoon shadows, and all sorts of special events at any time of day or night, as well as studio and theatrical shows with standard stage lighting, and a host of industrial and military applications.

Field-Pickup Equipment

The first and most obvious application for the image orthicon is in field or remote-pickup equipment.[3] This type of equipment must be so designed that it can be transported quickly and easily and set up almost anywhere for operation with little more than a moment's notice. Usually, under such conditions, it is impossible to control the amount of illumination on the scene; hence, if it is to be truly useful, the pickup device must have sufficient sensitivity and range to function with the amount of light available at any time or place. The new field-pickup equipment being produced by the Radio Corporation of America has been designed to meet this need.

In the design, consideration has been given to the possible needs for using the field equipment under three different types of conditions. These are:

1. In temporary locations, inaccessible to vehicles, to which the equipment must be carried by hand.
2. In temporary locations accessible to vehicles where all of the equipment except the cameras may remain in a suitable mobile unit which serves as a control center.
3. In permanent locations where the equipment may be used for studio productions.

One of the first two of these conditions is encountered in every operation in the field. The third condition may exist in the case of a small broadcaster who wishes to begin studio operations with a minimum of capital investment. He may wish to use the same equipment for both field and studio work in case he is operating on a limited schedule which permits the necessary breaks for transporting the equipment. This third condition may also apply to the ambitious broadcaster who, like many in these times, is unable to obtain any other type of equipment immediately, and who, in spite of this, wishes to get the training of technical and program personnel under way for more extensive operations in the future.

These conditions, together with electrical considerations, dictate in large measure how the equipment should be divided into units. Each unit should be small and light enough to be carried by one man. On the other hand, the number of units must be kept to a reasonable minimum in order to facilitate assembling and disassembling in the field. The shape of the units must permit easy handling, and also permit setting them side by side on a bench or table so that the assembly of units has the general appearance and utility of a console. Simple and rapid means of electrical interconnection are a further requirement. To meet these requirements, most of the major units of the field equipment have been housed in cases resembling a medium-sized suitcase in both shape and dimensions. Cameras, view finders, and master monitors have special requirements which necessitate deviations from this standard shape.

The block diagram of Fig. 1 shows the arrangement of major units required to make up a system of field-pickup equipment consisting of two or more cameras,

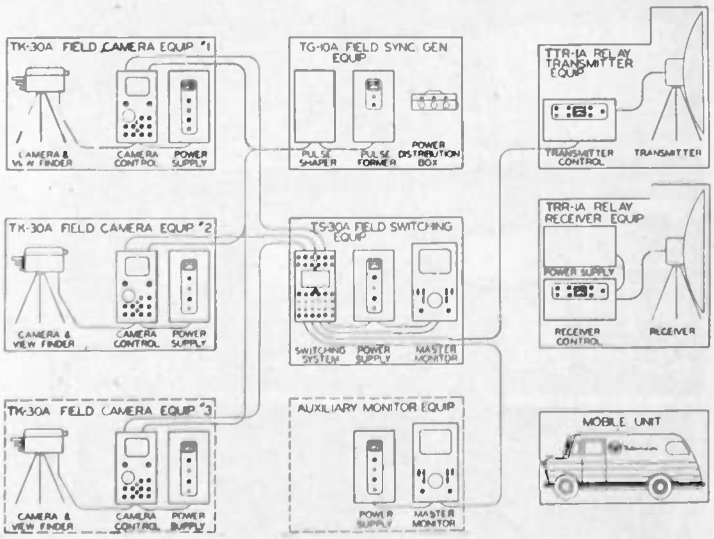


Fig. 1—Block diagram of field-pickup equipment.

with necessary switching facilities, radio relaying, and a mobile unit. It includes also a simplified schematic diagram of the interconnections. The two large upper-left-hand blocks show the actual camera equipment required for a standard two-camera system. The third block be-

[3] R. E. Shelby and H. P. See, "Field television," *RCA Rev.*, vol. 7, pp. 77–93; March, 1946.

low (in dotted lines) illustrates how additional cameras, up to a total of four, may be included in the system. The blocks (in solid lines) in the center and right-hand side of the diagram show equipment which is common to the entire system, whether it be composed of two, three, or

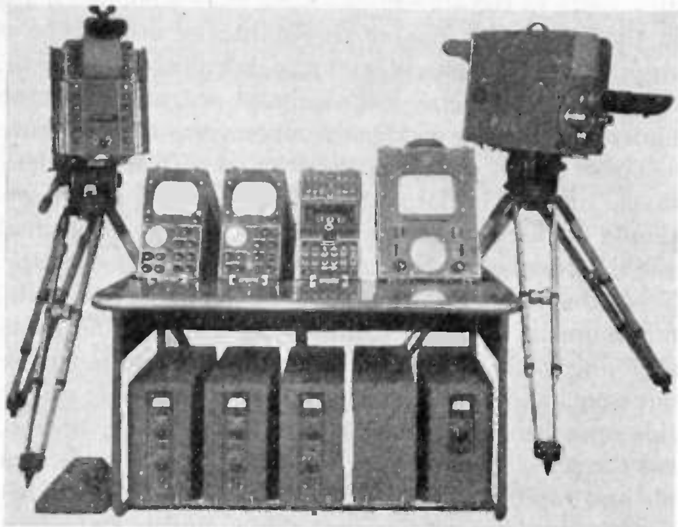


Fig. 2—Two-camera system including desk.

four cameras, and which need not be duplicated when cameras are added to the system. The dotted block in the lower center of the diagram shows additional monitoring equipment which may be added to provide a sec-

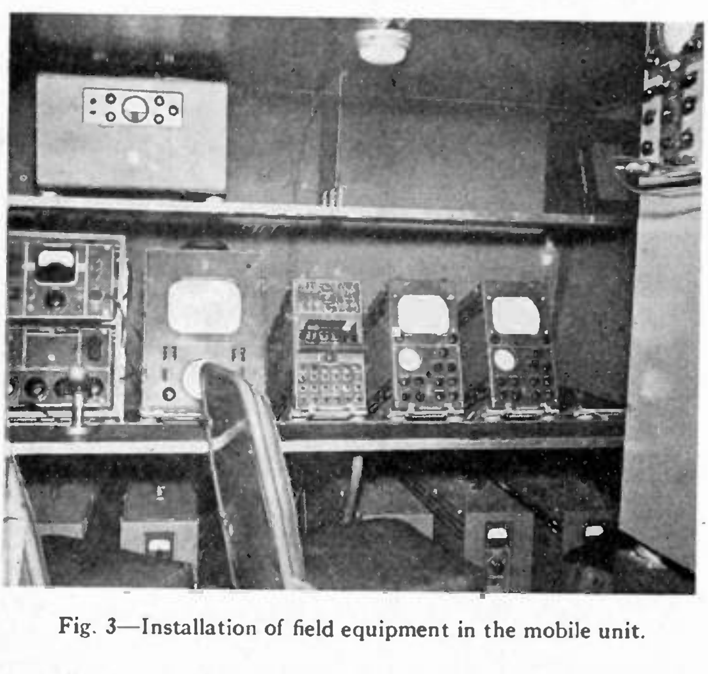


Fig. 3—Installation of field equipment in the mobile unit.

ond viewing position for an announcer, for visitors, or for other special purposes. In the case of single-camera operation, the switching equipment and auxiliary monitoring equipment are omitted.

The system illustrated provides a maximum of flexibility with a minimum number of separate units. As a system it provides many features which make for ease in operation and fine performance.

Fig. 2 illustrates the equipment required for a two-camera setup, mounted on a desk such as may be used for studio operation. The units on top of the desk include two camera controls, a master monitor, and a switching system. These units contain all the controls normally required by the operators during the program. The other units under the desk are those which normally require little or no attention during program time. These units are the synchronizing generator and the power supplies.

Fig. 4—Mobile unit in operation, with camera and relay transmitter on the roof.

Fig. 3 shows the same equipment mounted in a similar manner in a mobile unit. Fig. 4 is an external view of the mobile unit, showing how access to the roof is provided through a hatch, and how a camera may be set up for

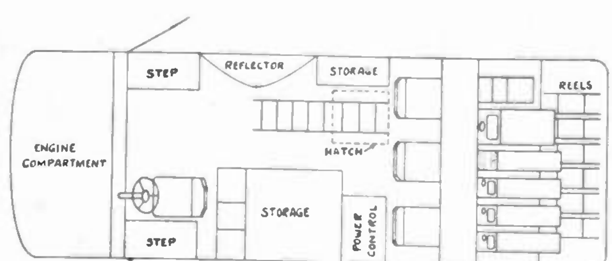


Fig. 5—Plan diagram of the mobile unit.

operation on the roof. Sufficient space is also available on the roof for setting up a microwave relay transmitter. Storage space for a maximum of 1200 feet of camera cable is provided on reels with swing-out brackets at the rear of the mobile unit. The general plan of the mobile unit showing operating positions and storage space for cameras, tripods, view finders, relay transmitter, sound-pickup equipment, and miscellaneous accessories, is illustrated in Fig. 5.

Camera

Full advantage has been taken of the relatively small size of the image-orthicon tube in designing a compact camera. The dimensions of the case, including the cover, but without lenses or view finder, are $20 \times 10\frac{1}{2} \times 11\frac{1}{4}$ inches, and the weight is 65 pounds (see Figs. 2 and 9).

The principal features of the camera are as follows:

1. Image-orthicon pickup tube.
2. Completely self-contained deflection circuits.
3. A four-position lens turret with rear control for quick change of lenses.
4. Miniature tubes in picture preamplifier.
5. Small, flexible camera cable.
6. Operation over a long cable (up to 1000 feet).
7. Forced-air ventilation.
8. Accessibility for servicing.
9. Rugged mechanical construction.

Though the use of lens turrets is well known on photographic cameras, their application to television cameras has not been attempted before, mainly because the lenses required for iconoscope and orthicon cameras are too large and heavy for a suitable turret mechanism. Furthermore, the use of optical viewfinders on many such cameras, requiring matched pairs of lenses, at least doubles the difficulties of turret design.

The useful photocathode area of the image orthicon is a rectangle 0.96 inch in height by 1.28 inches in width. Since this is approximately the same size as the frame of many miniature photographic cameras which use 35-mm. film, it is possible to use lenses designed for such cameras. The Kodak Ektar lenses for the Ektra camera provide a useful series of focal lengths which have been applied to the image-orthicon camera. Available lenses include 50-, 90-, and 135-mm. focal lengths. These lenses are light in weight and are excellent for turret operation. Special lightweight lenses up to 25 inches in focal length and with $f/5$ apertures have been constructed using achromats in black bakelite barrels with quick-change slotted mountings. These weigh only 2 to 3 pounds and may be attached to the turret (see Fig. 2).

The four-position turret is mounted on a hollow shaft which extends through the camera to a control handle and indexing mechanism in the rear at the operator's position. Releasing the indexing detent automatically cuts off the picture signal while the turret is being rotated to another position.

Optical focusing is accomplished in a novel manner by moving the pickup tube, along with its focus and deflection-coil assembly, instead of by motion of the lens. The mechanism is self-locking in any position of the camera. The greatest advantage in this system is the obvious simplification of the turret. A second important advantage is the increased range of focus obtainable when lenses with individual focusing mounts (such as the Ektar lenses) are used. The total available relative motion between lens and target is then the sum of the individual motions. A further advantage of the individual focusing mounts is that lenses of different focal lengths may be preset to focus on the same scene, thus eliminating the need for adjusting optical focus after rotation of the turret.

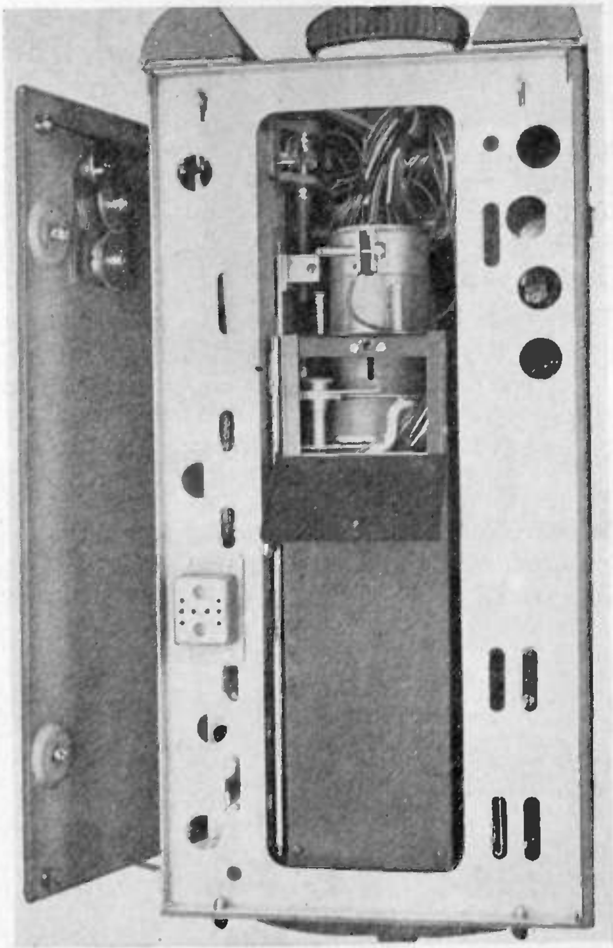


Fig. 6—Top view of camera, showing coil assembly.

Fig. 6 shows a top view of the camera in which the coil assembly and magnetic shield are exposed. The coil assembly is supported on a steel plate which moves on three rollers. At the rear of the compartment may be seen the focusing drive screw and the wiring to the base of the image orthicon. A small trap door at the rear end of the magnetic shield box exposes the cross field or alignment coil and the gear drive used for rotating this coil.

Fig. 7 shows the focus coil alone. This is a simple, random-wound solenoid long enough to enclose both the deflecting coils and the image section of the image orthicon tube with an overhang of about one-half inch at the front and one inch at the rear. The deflecting coil assembly, which is mounted within the focusing coil, is illustrated in Fig. 8.

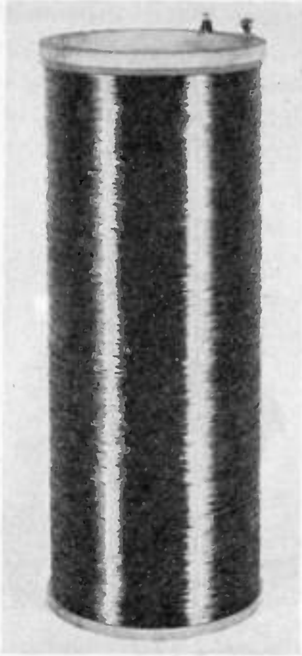


Fig. 7—Focus field coil for the image orthicon.

The deflection circuits are included in the camera in order to reduce the number of major units in the field equipment. To make the camera capable of operating over a long cable, it is necessary to locate the deflection generators either in the camera itself or in an auxiliary unit adjacent to the camera. Locating the deflection circuits and part of the picture preamplifier in an auxiliary unit makes it possible to keep the size and weight of the camera to a minimum. Such an arrangement, however, complicates the system by increasing the number of units, and hence the number of connecting cables and the time and effort required for setting up, dismantling, and transporting the equipment. A further objection is that, in some field operations, an auxiliary unit is a serious nuisance, especially when the camera has to be set up on a small stage or platform where space is restricted. In the case of the image-orthicon camera, it is possible to include all of these circuits in the one unit without making the camera unreasonably large or heavy. With this arrangement, it is necessary to transmit over the cable only the timing information in the form of driving pulses. The transmission lines used for this purpose are easily terminated with resistors, and the pulses, which are not unduly critical as to wave form, are then easily amplified to usable levels.

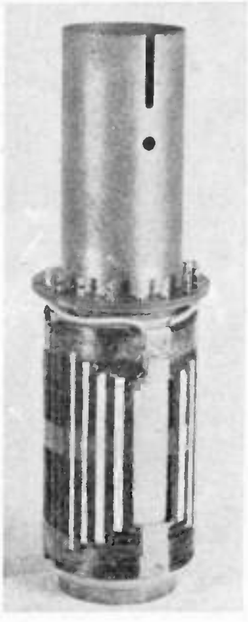


Fig. 8—Deflecting-coil assembly for the image orthicon (outer tube removed).

The horizontal-deflection circuit, in common with similar circuits in other parts of the system, employs two new types of tubes, the 6BG6G and 6AS7G. The 6BG6G is similar to the 807, but has special characteristics for deflection output service. The 6AS7G is a twin triode, having very low plate resistance and large power capabilities. It is used as a damper or reversed-current output tube.

The horizontal retrace period is made about 10 per cent of the total horizontal scanning period, in order to avoid the necessity for artificial compensation for delay in long camera cables. The difference between the minimum kinescope blanking width (16 per cent) and this retrace is 6 per cent, or 3.8 microseconds. This is just slightly in excess of the time required for a round trip (2000 feet) in a 1000-foot cable.

The high voltage required for operating the image-orthicon tube totals about 2000 volts, -500 volts required in the image section and $+1500$ volts in the signal multiplier. This is generated by amplifying and rectifying the pulse signal that appears across the horizontal deflecting coils. Negative pulses are partially integrated and fed to the grid of a 6V6GT amplifier with its plate coupled to the primary of a special step-up transformer. The screen and cathode circuits of this amplifier are made degenerative in such a way as to compound the plate current. As a result, the peak plate current at the beginning of each retrace period is constant over a two-to-one range of pulse input to the grid. Thus the voltage fed to the rectifier is nearly independent of the horizontal scanning amplitude (width). The high-voltage transformer includes a small heater winding for the filament of a type 1B3/8016 rectifier. Suitable voltages for the various electrodes in the image orthicon are obtained from a filtered bleeder.

Negative feedback is employed in the vertical-deflection circuit by deriving a voltage from the drop across a small resistor in series with the deflecting coils and, after amplification, injecting the feedback signal into the plate circuit of the first sawtooth-amplifier stage. This feedback does two important things. It eliminates almost entirely the effect of iron saturation in the transformer core and nonlinearities in the amplifiers. It also minimizes the effect of varying tube characteristics, and makes the vertical scanning linearity largely independent of amplitude.

Blanking signal for the target in the image orthicon is derived from the horizontal and vertical driving signals by mixing.

Controls associated with the scanning circuits are all located in the camera. These include height, width, centering, and linearity controls. Other controls also located in the camera are preamplifier gain, image accelerator, orthicon decelerator, and horizontal shading. None of these controls requires attention during actual operation, and hence the camera man is left free to aim the camera and focus the optical system.

The picture signal is amplified in a five-stage preamplifier built into the camera. The preamplifier employs miniature tubes and circuits compensated to give uniform output up to approximately 8 Mc. The cathode follower in the final stage serves to feed the signal over a coaxial transmission line to the camera control, and also to provide signal for operation of an electronic view finder which may be used with the camera.

Components in all parts of the camera are accessible for servicing, and can be removed easily in case replacement becomes necessary.

A single camera cable contains all the electrical connections to the camera. It includes three 50-ohm coaxial transmission lines and 21 other conductors used for power, control, and communication. The cable is unusually small in diameter (0.84 inch) and light in weight.

View Finder

Television cameras have been equipped in the past with a wide variety of view finders, ranging from two screw heads used as rifle sights, through wire frames and double-lens systems, to electronic finders in which the scene is reproduced on small kinescopes mounted beside or above the cameras. Each type has advantages, but no one type has all the desired characteristics. In the cases of iconoscope and orthicon cameras, the optical view finder employing a second lens identical with the camera lens has enjoyed the greatest popularity because it not only serves to indicate focus, but is capable of including portions of the scene outside of those actually being televised. This has been considered important because the camera man can see and avoid unwanted objects before they intrude themselves in the picture.

In the case of the image-orthicon camera, the double-lens type of optical finder becomes completely useless when the equipment is used under limiting low-light conditions. This is true because the image orthicon can operate with such low illumination that the image on a ground-glass screen is nearly invisible. Thus the electronic view finder is the only remaining type capable of indicating both focus and the outline of the scene. It has two distinct advantages over the optical system. It is entirely free of parallax errors, and it provides an erect image where a single-lens direct optical finder provides an inverted image. The electronic view finder has a disadvantage in that it cannot include anything outside of the televised scene.

The view finder designed to be used with the image-orthicon camera employs a flat-faced 5-inch kinescope tube (type 5FP4) with about 7000 volts on the second anode. This arrangement provides a picture with sufficient brilliance to be seen readily under bright ambient light. The view finder is constructed as a separate unit to be mounted on top of the camera. The two units are styled to appear as a single unit when thus assembled.

The physical arrangement is such that the kinescope faces the operator at the rear of the camera. The face of the tube may be shaded with either of two types of viewing hoods. One of these includes two mirrors in a periscopic arrangement which may be reversed so that the operator's eye level is either above or below the kinescope, depending on the height of the camera. The other hood provides a direct view of the kinescope. A single cover opens on a hinge at the front, exposing the entire internal assembly (see Fig. 9).

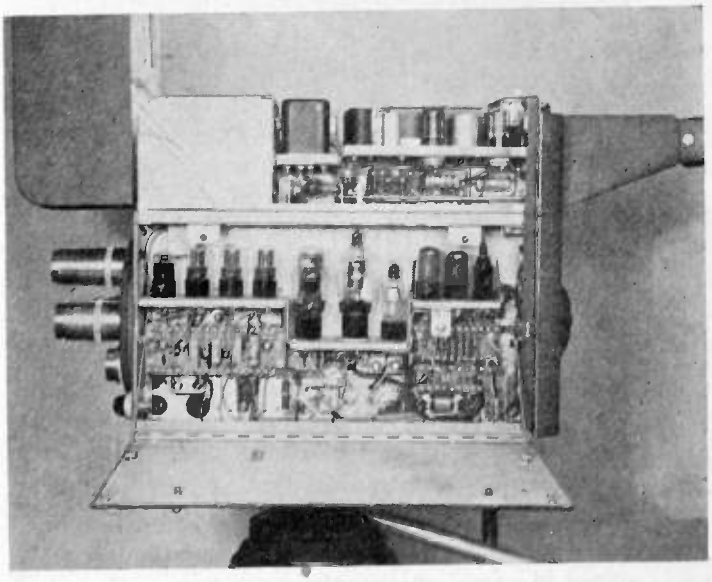


Fig. 9—Deflection-amplifier side of the camera and view finder (internal view).

The circuits include the picture and blanking amplifiers required to drive the kinescope, and also the deflection generators and high-voltage supply. The latter is a pulse type of supply associated with the horizontal-deflection circuit. Necessary controls are accessible at the rear in line with the operating controls on the camera. All electrical connections are made through a multicontact plug and receptacle (see Fig. 6).

An auxiliary view finder in the form of a polaroid ring sight may be mounted on top of the periscope viewing hood (Fig. 2), or, in the absence of the electronic view finder, on the camera itself. This ring sight produces a series of concentric spectral interference rings which appear to be at a considerable distance in front of the sight. Because they appear at a distance, the eye can observe the rings and the scene simultaneously with a minimum of strain. This device is useful in following action which moves too rapidly and too far to be followed readily on the kinescope. Its usefulness is limited, however, because it does not indicate either correct focus adjustment or the boundaries of the scene. It is simply an aiming device.

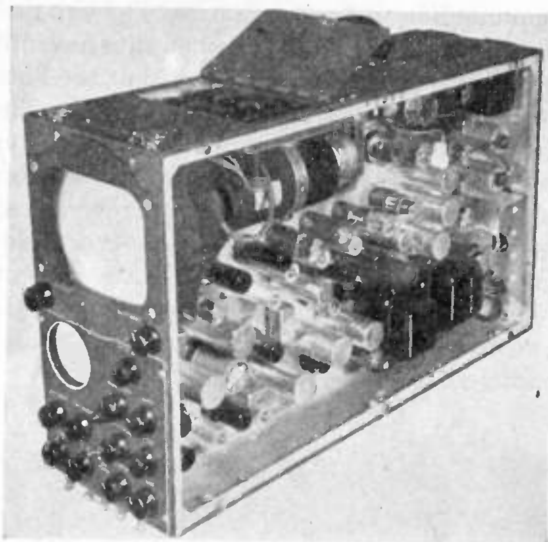


Fig. 10—Field-camera control.

Camera Control

The camera control (Fig. 1) is a unit which performs all of the functions not already performed in the camera itself that are necessary to the production of a complete composite picture signal. These functions include:

1. Amplification of the picture signal to the standard level required for feeding outgoing lines.
2. Addition of kinescope blanking signal.
3. Establishment and maintenance of the peaks of the blanking pulses at true "black" level.
4. Addition of the receiver synchronizing (sync.) signal in cases where only a single camera is in use.
5. Monitoring of the finished picture signal to check the accuracy of optical and electrical focus in the camera and the general quality of performance of the camera chain by means of the following:
   (a) A picture monitor tube (kinescope) which reproduces the scene being televised.
   (b) A wave-form monitor tube (cathode-ray oscilloscope) which shows the wave form of the picture signal and measures the amplitude of this signal.
6. Controlling electrical focus and other parameters involved in operation of the image-orthicon tube in the camera.

From consideration of these six functions it is apparent that the camera control is necessarily a complex unit, for it includes all the circuits and components found in that part of a television receiver which follows the second detector, also those required for a wide-band

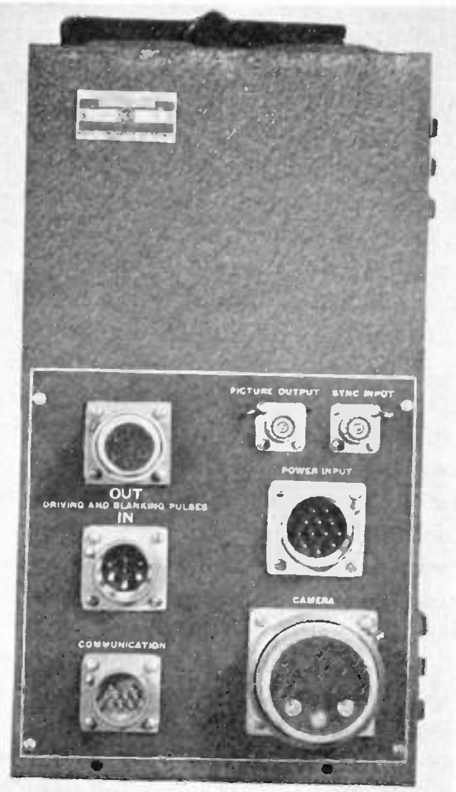


Fig. 11—Rear of the camera control.

cathode-ray oscilloscope, and, in addition, amplifiers, special circuits and controls, and cable connectors required directly for operation of the camera.

As indicated previously, the shape of the camera control is that of a medium-size suitcase, the dimensions being approximately $8 \times 15 \times 24$ inches (Figs. 2, 10, and 11), and the weight about 65 pounds. The chassis and case are spot-welded into a rigid, durable assembly. The kinescope (type 7CP4), the c.r.o. tube (type 3KP1), and the most important controls are mounted on the front end of the case. All small tubes, capacitors, and transformers are mounted on one side of the chassis, with wiring on the opposite side. Controls of secondary importance are mounted under a trap door in the top of the case. Past experience and a good deal of thought have produced a chassis layout which provides a maximum of accessibility for servicing, and at the same time a system for rigid, vibration-proof mounting of components which contributes much to trouble-free operation. A removable metal cover protects the cathode-ray tubes and controls during transportation. The two side panels or covers are easily removed by releasing three cowl fasteners at the top of each, and lifting them from three spring retainers at the bottom. All external electrical connections are made through plugs and receptacles on the rear of the case (Fig. 11). This same general construction is followed in the other suitcase units described hereinafter.

The circuits in the camera control include:

1. The picture amplifier, with stages for mixing kinescope blanking and synchronizing pulses.
2. A picture amplifier for the monitor kinescope.
3. A picture amplifier for the c.r.o. tube (for vertical deflection).
4. Deflection circuits for both c.r. tubes.
5. Distribution amplifiers for feeding driving pulses to the camera.
6. A filament transformer.
7. A high-voltage transformer, rectifier, and filter for the c.r. tubes.
8. Camera circuit controls.
9. "On-the-air" tally and intercommunication system.
10. Remote power control.

The picture amplifier consists of several stages of types 6AC7 and 6AG7 tubes in conventional frequency-compensated circuits. One stage in this amplifier performs the very important function of establishing the peaks of blanking at "black level." To do this, the control grid is clamped at the end of each scanning line to an arbitrary reference potential. Because the target in the image orthicon is blanked during the scanning retrace (i.e., made sufficiently negative to repel the scanning beam) the picture signal from the camera during this retrace period is fixed with respect to black level, though it may vary continuously with respect to an arbitrary fixed reference because of the addition of hum, power-supply surges, or other spurious signals. The clamping action serves to set up a fixed relationship between the actual black level in the retrace periods of the picture signal and the arbitrary reference by connecting the control grid mentioned above to the reference potential through a very low impedance. At all times, except during the retrace periods, the grid is disconnected from the reference, and thus is free to follow the normal potential variations in the picture signal.

An important by-product of this clamping action is the elimination of the low-frequency components of any spurious signals, provided they do not have sufficient magnitude to cause amplitude modulation in any preceding stage. Hence, the clamp circuit removes power-supply surges and low-frequency hum, and minimizes microphonics. In fact, it limits the amplitude of any spurious additive signal to the amount which occurs in the period of one scanning line. (For a more detailed description of clamping, see the Appendix.)

Kinescope blanking is mixed with the camera signal just ahead of the clamper. It provides undistorted, noise-free blanking intervals by the addition of independent, carefully controlled pulses. Since this added blanking is constant in amplitude, it does not affect the clamping action in any way except to shift the constant relationship between black level and the reference to a different constant value. After clamping, the combined camera and blanking signal is clipped near black level, thus producing a final signal in which the peaks of blanking bear a definite relationship to black level. The clipper makes use of a diode as a switch in series with the picture signal circuit. It depends for its accuracy in maintaining black level on the clamping which precedes it. This clipper is somewhat more complicated than the usual plate-current-cutoff type of clipper, but is justified because it cuts off very abruptly and is almost perfectly linear in the neighborhood of cutoff. (See Appendix.) A manual control (BLANKING) adjusts the clipping level to any desired point near black level, and thereafter the circuit maintains clipping at that level. Usually the clipper is adjusted so that the peaks of the blanking pulses are slightly "blacker than black," thus assuring complete removal of the retrace lines in receiver kinescopes.

D.c. restoring circuits maintain black level (or sync. peaks when sync. is present in the output) on the grid of the kinescope and on the grids of several stages where it is important to minimize distortion.

Deflection circuits for the kinescope are of the driven type. These circuits are of the same general kind as those used in the camera described previously, the only differences being in the deflecting-coil design and matching transformers.

Seven electrical controls, grouped on the front panel, provide for maintenance of proper operating conditions in the camera and associated picture-amplifier circuits in the camera control during the program. These are: (1) GAIN, (2) BLANKING, (3) BEAM CURRENT (ORTH.), (4) ORTHICON FOCUS, (5) IMAGE FOCUS, (6) TARGET POTENTIAL, and (7) MULTIPLIER FOCUS.

Only the first two of these require frequent checking during operation. However, the others are easily available to the operator at the camera control without the need of distracting the attention of the camera man, who is occupied with his normal duties. Location of these controls in the camera control is particularly useful in the process of making adjustments when a new image orthicon tube is installed in the camera, because the number of adjustments to be made in the camera itself is reduced to a minimum. Controls of secondary importance, such as size and centering for the kinescope and c.r.o., are located under a small trap door on top and near the front of the unit, easily accessible to the operator.

Plate current for all of the amplifier tubes is obtained from a regulated power supply entirely separate from the camera control. A power switch on the front panel of the camera control actuates a relay in the power supply which, in turn, opens or closes the power-input circuit for the entire camera chain.

Power Supply

The problem of providing the large amount of highly stabilized d.c. required for the large number of amplifier tubes in a camera chain has been solved in a unique way in the field power supply. The problem resolves itself into one of finding means to reduce the weight of the unit to a point where one person can carry it. The difficulty may be understood when it is pointed out that the total plate-current drain in a single camera chain is approximately 1 ampere at 285 volts, regulated within limits of less than ± 0.5 volt. The regulation does not constitute the major part of the problem, but simply adds to it by increasing the total voltage required from the rectifier.

The general attack on this problem was developed several years ago in the design of portable television equipment for the type-1840 orthicon.[4] A very lightweight transformer with the core divided into sections, with large openings in the end turns of the windings and with only a small fraction of the usual amount of iron and copper, was designed to be used with a continuous blast of air through the openings (Fig. 12(a)). This transformer, together with the blower and motor, weighed less than 20 pounds, and the entire power supply, including case, transformer, blower, tubes, and other components, weighed only 58 pounds. This design achieved the required objective, and gave reasonably good service in field use for several years.

In the field power supply for the image-orthicon equipment, a new and much improved transformer has been developed by making use of silicone enamel on the wire and glass fabric impregnated with silicone varnish for insulation between layers of the windings. The core is not sectionalized, and the windings are tight, as in conventional transformers (Fig. 12(b)). The running temperature may be as high as 180° C. without danger of deterioration. As a result of this design, the over-all weight of the field power supply has been kept the same as in the earlier model, and the reliability has been increased.

New regulator tubes have made possible an improvement in efficiency. The type 6AS7G, a heavy-duty twin triode with estremely low plate resistance and the ability to dissipate 25 watts, is used for series regulation. This is the same tube that is used as a damper in the horizontal-deflection circuits. These tubes have appreciably less voltage drop than other types previously used in such service, and hence are more efficient. They also have very high transconductance, and therefore provide improved regulation control.

The rectifier is connected to a two-stage choke-input filter using electrolytic capacitors, through a thermal time-delay relay which prevents application of the high d.c. voltage until all tube heaters have attained operating temperature.

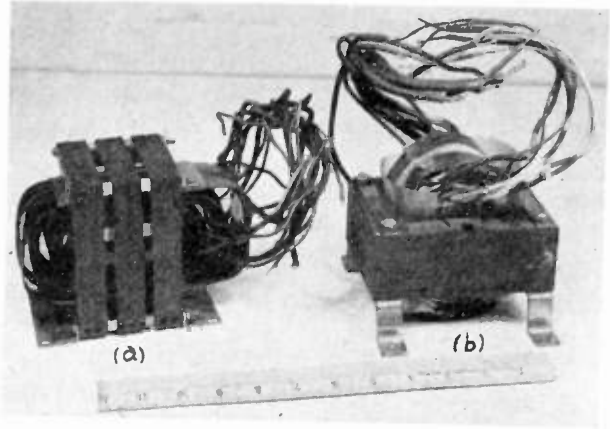


Fig. 12—(a) Prewar design of a forced-air-cooled power transformer. (b) New design.

A 6SL7GT tube functions as a two-stage control amplifier, and two OD3/VR150 tubes serve as voltage references.

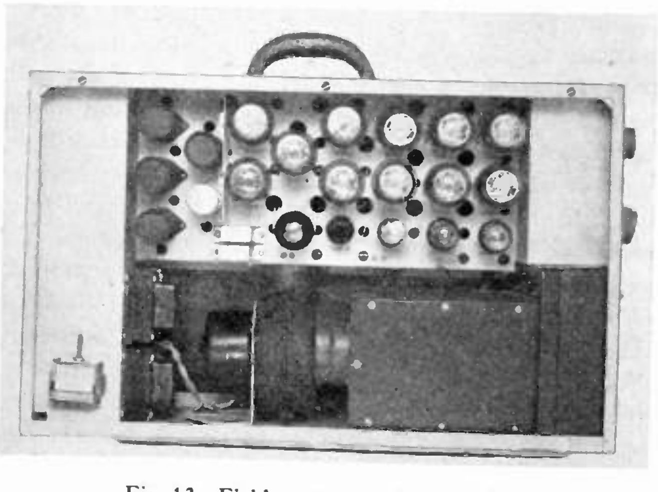


Fig. 13—Field power supply, tube side.

The field power supply is capable of delivering 950 ma. at 285 volts continuously to the main load, and, in addition, 75 ma. to the focusing field coil in the camera. This latter supply is current-stabilized so that changes in the resistance of the coil do not change the current. Fig. 13 is a side view of the field power supply, showing the transformer housing, blower, and tubes.

The primary power circuit includes means for switching and metering of taps, so that a wide range of sup-

[4] M. A. Trainer, "Orthicon portable television equipment," Proc. I.R.E., vol. 30, pp. 15–19; January, 1942.

ply voltage may be accommodated. Provision is also made for metering currents and voltages in parts of the output system.

Synchronizing Generator

The new field synchronizing generator, which is part of the image-orthicon equipment, is designed on the same basic principles as earlier models, but improvements and new features have been added which make its performance the equal in every respect of that of the studio type of generator. Equality of performance is obviously necessary, especially in view of the increasing importance of field operations in television programming.

The field synchronizing generator comprises two suitcase units having the same size and shape as the field camera control. They are called the field pulseformer and field pulse shaper, respectively. These two units generate four distinct signals required for operation of the entire television system, including the receivers. All four signals, though different in wave shape, are accurately synchronized with each other by being derived from a single primary frequency source. Two of these signals appear directly in the composite picture signal which modulates the r.f. carrier. They are "kinescope blanking" and "synchronizing" (or "sync."), respectively. The wave shapes of these two signals are specified completely in standards recommended by the Radio Manufacturers Association.[5] The remaining two signals, "horizontal driving" and "vertical driving," respectively, are simple pulse signals used locally in the pickup equipment for triggering camera and monitor scanning circuits, and for target blanking and clamp-circuit keying.

The principles underlying the operation of this generator have been described fully in a previous publication.[6] No basic changes have been made in the arrangement of circuits, but refinements have been included to increase the stability of the primary frequency source and also to improve the steepness of wave fronts in the outputs. Among these, specifically, are a crystal oscillator which may be used in locations where the power supply frequency is unstable, an improved a.f.c. circuit for lock-in with a 60-cycle power supply, an additional counter to reduce the maximum number of steps in any given counter, and a cathode-ray-tube indicator to provide a means of quickly checking the operation of the counters.

One of the two units includes a built-in regulated power supply, thus making the synchronizing generator completely self-contained in the two units. Separation of the circuits occurs at a point where only three signals require connections between units. A single multiconductor cable connects the pulse former to the pulse shaper. The only input to the pulse former is a.c. power. Output from the pulse shaper is split in two cables, one a single coaxial line for synchronizing signals and the other a multiple coaxial cable for the other three signals. The two suitcases appearing in the lower right hand corner of Fig. 2 are the pulse shaper and pulse former.

Switching System

One of the most important operations in television programming is that of switching from one camera to another. Switching must be accomplished smoothly without either interrupting or disturbing the receiver synchronizing, even momentarily. If precautions are not taken to avoid surges in switching, it is possible that the sync. may be clipped later in the system during the period of the surge. Some receivers are very sensitive to such interruptions. Cases have been known in the past where switching surges have been so large as to overload the transmitter and throw it off the air. It is not possible to experience such difficulties in properly designed television systems today because means are used to maintain constant black level at all points where surges are harmful. Since switching is likely to produce surges, it is desirable to eliminate them at this point. A successful means for accomplishing this is the clamp circuit which was described previously in the section on the camera control. This circuit restores the picture signal to some arbitrary reference level at the end of each scanning line; i.e., during the retrace or blanking period. It is independent of anything that takes place in the signal. Thus no surge can exist longer than the period of one line.

The field switching system is a suitcase unit of the same shape and size as the other units described previously (Fig. 14). On the front panel are located two sets of push-button switches, the lower one of which provides for switching among four cameras and two auxiliary picture circuits. Each of these buttons has an associated tally light which operates in conjunction with tallies on the respective camera and camera control selected by it. These six switches connect six coaxial 75-ohm lines, one at a time, to the input of the picture amplifier contained in the unit.

The picture amplifier consists of three stages, the last one being a cathode follower which feeds the picture line to the relay transmitter, or a line directly to the main studio (75-ohm coaxial). A blocking capacitor separates the line from the cathode, so that no direct current flows in the line. The grid of this cathode follower is subjected to the action of the clamp circuit. Hence, no surges appear on the outgoing line.

Two other coaxial lines also provide signal to other parts of the system. One of these is connected to a line

[5] "Synchronizing Generator Waveforms," a drawing compiled by the Subcommittee on Studio Facilities of the RMA (revised, October 9, 1946).

[6] A. V. Bedford and J. P. Smith, "A precision television synchronizing signal generator," *RCA Rev.*, vol. 5, pp. 51–69; July, 1940.

monitor, or field master monitor. It is fed through a separate unity-gain amplifier contained in the switching system. The input of this amplifier may be switched with the upper set of push buttons to any of several points in the pickup equipment. The second line may be used to feed an additional monitor for the use of spectators or an announcer, or it may feed a stand-by relay transmitter. All three output lines carry identical signals.

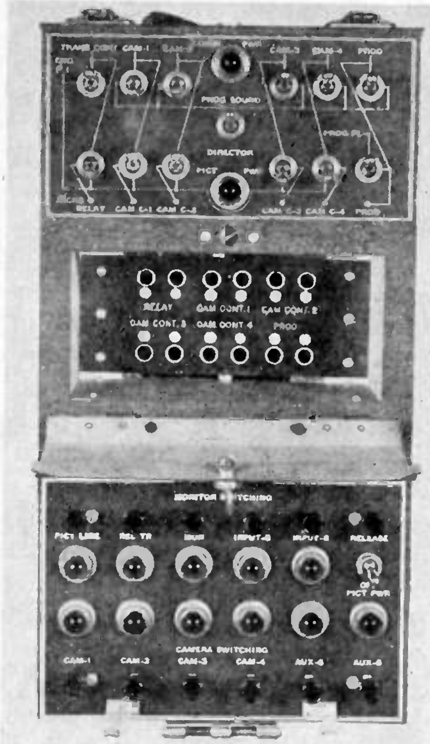


Fig. 14—Field switching system, front side.

The synchronizing signal is mixed with the camera signal in the switching system to form the final composite picture signal. The synchronizing signal is supplied to the switching system directly from the pulse shaper, and is coupled to the picture output line through a two-stage amplifier. Thus, the synchronizing pulses are always transmitted independently of the camera switching. In cases where picture signal already including the synchronizing pulses is being received over one of the auxiliary input circuits, the local synchronizing signal may be disconnected by turning a special switch on the front panel.

Keying signal for the clamp circuit is derived from the sync. signal. In cases where the incoming signal includes sync., the sync. is separated, as in a receiver, and delayed so that keying is done on the "back porch," i.e., on the peaks of blanking just following the sync. pulses (see Appendix). In the usual case where the picture signal is received from a local camera chain, sync. is not present at the clamped grid; hence it is not necessary to delay the keying in order to clamp at black level. In either case, clamping is done automatically at black level.

Circuits are included in the switching system for communication between the various technicians operating the equipment. Two sets of telephone jacks are mounted in each camera, one for the cameraman and the other for a program assistant stationed at the camera. Connections for these telephone sets are included in the camera cable. These intercommunication circuits all terminate in the field switching system where the technicians operating the camera controls and switching system, and the program director may connect their telephone sets. Private telephone lines to the main studio also terminate here, and may be connected to the local circuits. A variety of communication network combinations may be secured with the set of toggle switches on the upper front panel.

Each telephone set consists of a carbon-button microphone and two earpieces. The microphone and one earpiece are used for the intercommunication circuits. The second earpiece is connected to a separate circuit which carries program sound. Thus, each operator can hear the program sound at all times, and get useful cueing information from it.

Power for operation of the telephone circuits is obtained through a selenium-disk rectifier from the power lines, and is entirely independent of the power for the picture-amplifier circuits.

Field Master Monitor

A high-quality picture monitor primarily designed for studio applications has been adapted for use with the field equipment by the design of a special carrying case. (See Fig. 2.) This monitor contains a 10-inch, nearly flat-faced kinescope with aluminum backing (type 1816P4), and a 5-inch oscilloscope tube. The kinescope, which operates at about 9000 volts, provides an exceptionally bright picture suitable for program monitoring in high levels of ambient illumination. The oscilloscope provides a large, clear trace of the outgoing picture signal, and associated circuits include means for accurate calibration of signal level. As indicated in Fig. 1, the monitor is operated on the same power supply as the field switching system.

The shape of the master monitor case is somewhat different from the shape of the other suitcases because of the size of the tubes used. However, the dimensions and weight are of the same order.

Appendix

Two unusual circuits which contribute to the satisfactory performance of the field equipment are described in some detail in the following paragraphs. One of them has been discussed in a previously published article,[7] but will be reviewed here in the light of its direct applications in this equipment.

Clamp Circuit

As applied in television, a clamp circuit is a device for establishing an arbitrary reference potential at fixed and regular intervals on some chosen circuit element in the picture amplifier. The operation of establishing such an arbitrary reference is very useful wherever it is desirable to restore the d.c. component of the picture

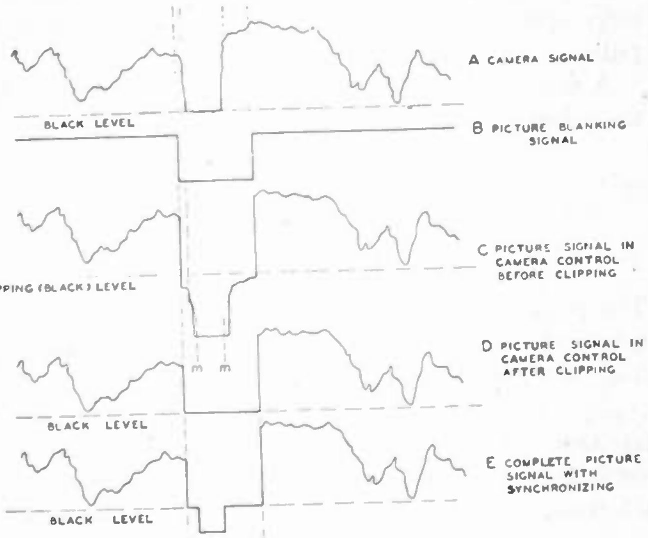


Fig. 15—Television signal waveforms.

signal, or, in other words, to make the actual "black level" in the signal coincide with an arbitrary reference. The clamp circuit is capable of doing this with an unusual degree of accuracy. It is a pulse-driven switching circuit, and is applicable only in cases where there are pulse intervals present in the amplifier signal during which the switching operation can take place. Since a television picture signal (having blanking pulses to suppress the scanning beam during the retrace periods) is of this type, the clamp circuit may be used successfully.

The orthicon and image-orthicon tubes both employ low-velocity scanning. Because of this, they generate picture signals which contain accurate black-level information during the blanked retrace periods. However, because these signals are generated at low level, the blanking pulses, which contain the "black-level" information, may include noise and other spurious components which make the pulses unsuitable for blanking in the receiver. This condition is illustrated in Fig. 15(a). To provide clean-cut blanking pulses in the final signal, it is customary to add another blanking signal (Fig. 15(b)) at a high-level point in the amplifier, giving the result shown in Fig. 15(c). Then this composite signal is clipped at black level to give the signal shown in Fig. 15 (d). To insure proper operation of the receiver, the clipping must be done accurately at black level. Here the clamp circuit is an indispensable tool. It is used to bring about a firm correlation between the black level existing in the negative peaks of the camera blanking pulses and the grid bias on the clipper stage of the amplifier. It should be noted that the addition of a constant signal (such as the blanking signal shown in Fig. 15(b)) does not affect the accuracy of the correlation between black level and clipper bias, but simply shifts the bias to a new value.

A simplified diagram of an amplifier controlled by a clamp circuit is shown in Fig. 16. It consists of two amplifier tubes, V_1 and V_2, with a clamp circuit, consisting of the switch S in series with a small resistance R, connected to the control grid of V_2 on which it operates. Whenever the switch S closes, the grid of V_2 is established at the potential of terminal P (which is the arbitrary reference potential), provided that S is closed for a time interval that is long compared to the time constant $(R+R_L)C_1$. This latter condition is necessary for proper operation of the clamp circuit.

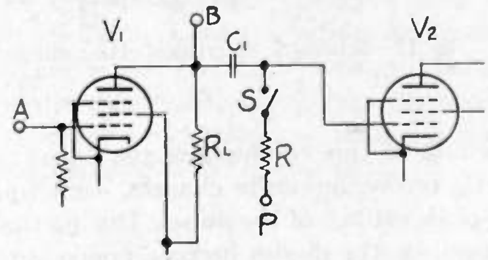


Fig. 16—Simplified schematic diagram of clamp circuit.

Assume that the camera signal of Fig. 15(a) has been introduced at terminal A in Fig. 16, and the blanking of Fig. 15(b) at terminal B, but with polarities such that the resultant mixed signal as shown in Fig. 15(c) appears on the plate of V_1. Now let the switch be closed for intervals such as m–m included within the peak of each camera blanking pulse, and let it be open the rest of the time. Thus the grid of V_2 is established firmly at the potential P at each peak of the camera blanking. The tube, V_2, is made part of a clipper or limiter, hence, when P is set at the proper value with respect to the cutoff potential of the clipper grid, the clipping will take place at black level.

In actual practice, the switch is a pair of diodes (contained in the twin diode, V_3) which are keyed on and off by equal pulse signals of opposite polarity, as shown in Fig. 17. These two pulse signals are coincident with each other and also with the time interval m-m in Fig. 15(c). Thus, both diodes conduct simultaneously and provide a low-impedance path for current flow to change the charge on C_1. In this case, the critical time constant is

$$\left(R_L + \frac{R_D}{2} + R'\right)C_1$$

where R' is the effective resistance of the signal source

[7] C. L. Townsend, "The clamp circuit," *Broadcast Eng. Jour.*, February and March, 1945.

which generates the keying pulses, and R_D is the effective resistance of one diode. In most cases, C_1 is made about 500 $\mu\mu$fd. and the total resistance about 2500 ohms. Hence, the time constant of the circuit is about 1.25 microseconds. Since the total keying interval is usually about 6 microseconds, there is time for the charge on C_1 to approach equilibrium.

In Fig. 17, the reference potential is that which exists at the midpoint of R_1. This may be deduced as follows. During the keying-pulse intervals, both diodes conduct, and hence both terminals of R_1 are at the same potential. Because of this conduction, the equal capacitors, C_2 and C_3, receive opposite charges, each equal to the peak-to-peak voltage of the pulses. During the intervals between pulses, the diodes become nonconducting, and the charges placed on C_2 and C_3 cause a current to flow in R_1, producing a voltage drop equal to the sum of the pulse voltages on the two capacitors. The polarity of the voltage is shown in Fig. 17. Since both the circuit and keying signals are balanced, it is then apparent that the diodes always arrive at a single potential during the pulse intervals which is the same as the potential existing at the midpoint of R_1 during the intervals between pulses. The time constant $C_2R_1 = C_3R_1$ is made very large compared to the period of the pulses, so that the current in R_1 is small; hence the charges on C_2 and C_3 remain essentially constant.

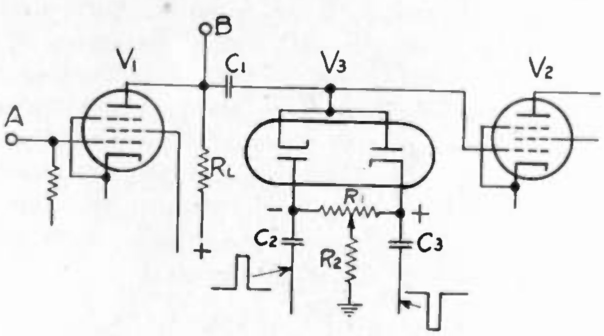


Fig. 17—Schematic diagram of clamp circuit.

If R_1 is connected as a potentiometer, as shown in Fig. 17, the reference potential (at the midpoint) may be shifted with respect to ground. For example, if the control is moved to the left, the midpoint becomes positive with respect to ground. This control is an effective and useful means of adjusting the bias on the grid of the amplifier tube, V_2. Whenever the control is moved away from the midpoint, the circuit becomes unbalanced, and difficulty may be experienced in maintaining pulse shape, especially when the control is near one end of R_1. To minimize this effect, a resistor, R_2, may be inserted in the ground connection. Since the average current in such a resistor is always zero, it may have a very large resistance. Use of this control in no way disturbs the accuracy of the clamping action in establishing the grid of V_2 at the reference potential (midpoint of R_1).

The only path for charging current from the capacitor C_1 (in the absence of grid current in V_2), is through the clamp circuit. During the open-circuit intervals in the clamp circuit, it is therefore impossible for the charge on C_1 to change. Hence the low-frequency response of the coupling circuit between V_1 and V_2 is not attenuated even though the capacitance of C_1 be made very small.

It is important that the keying interval *m-m* shall come to an end *before* the end of the blanking pulse, so that the charge which is left on C_1 will always correspond to black level in the picture signal, and not to some other level existing in the signal after the blanking pulse.

A further consideration is necessary in determining the proper value for the time constant

$$\left(R_L + \frac{R_D}{2} + R'\right)C_1.$$

The peaks of the camera blanking pulses usually contain some high-frequency noise signal originating in the low-level parts of the signal system. The response of the charge on C_1 to the clamping action must be slow compared to the period of the noise signal, in order to avoid variations in the correlation between black level and the reference potential. Black level may be considered as the average of the noise signal. Hence the clamp must be slow enough to average out the noise. Since the resistive elements are usually determined by the requirements of the keying circuits, the value of C_1 is used to control the time constant. Values cited previously have been found to work well in most cases, though where the noise signal contains low-frequency components it may be necessary to use a larger value for C_1.

The chief advantage in the clamp circuit, as compared to other types of leveling or d.c. restoring circuits, is the fact that its action is entirely independent of the picture signal in the amplifier. It depends only on the keying pulses. These may be controlled at will with respect to amplitude and timing. It should be noted that the amplitude of the keying pulses is usually made about twice the amplitude of the picture signal in the amplifier at the point where the clamp operates.

The ability to control the time at which the clamping is done is sometimes of great advantage. For example, in the field switching system, it is necessary to clamp the signal after the camera switching. At the point where the clamping is done, it is necessary to accommodate two types of picture signal: (a) from the cameras, in which case the sync. pulses are not present, since they are added subsequently, and (b) from an outside source which provides complete composite signal including the sync. pulses. Clamping must be done on the same level in both types of signal. Obviously, clamping cannot be done on the peaks of sync. in case (a); hence, it must be done on blanking peaks, or at true black level, which is present in both cases.

In case (b), the only available space on which the clamp may operate at black level is that portion of the blanking pulse immediately following the sync. pulse, commonly called the "back porch" (see Fig. 18). Keying

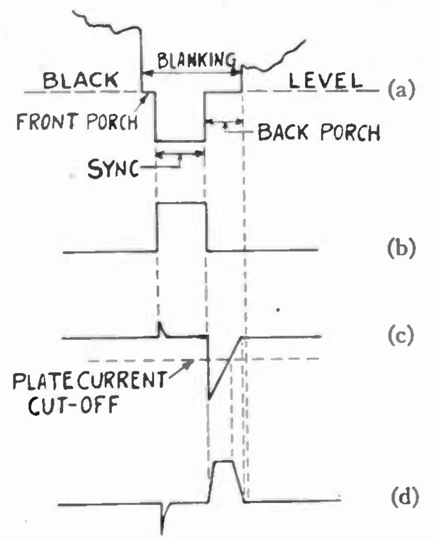


Fig. 18—Derivation of keying pulses for back-porch clamping.

pulses are derived from the sync. by separation as in a receiver, and subsequent forming. A sync. pulse after separation is shown in Fig. 18(b). This separated signal

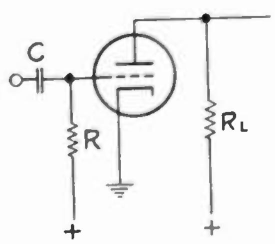


Fig. 19—Schematic diagram of differentiating circuit.

is fed to the grid of a triode as in Fig. 19. The capacitor C is made very small, about 20 $\mu\mu$fd., and the grid leak R is connected to a positive voltage source. Before the leading edge of the pulse, it may be assumed that the grid is drawing current, hence is approximately at cathode potential. The positive excursion of the leading edge causes very little change in the grid potential because of the low impedance from grid to cathode. The resulting sharp exponential pulse is shown in Fig. 18(c). After returning to its original potential, the grid is excited by the trailing edge of the pulse in the negative direction. This excursion stops the flow of grid current and also swings far enough to go beyond plate-current cut-off, as illustrated. The signal voltage causes no further change, but the positive voltage on the grid leak immediately starts to recharge the capacitor C, and thus produces the positive slope shown in the diagram. The rise in grid voltage stops abruptly as soon as grid current starts to flow. The steepness of this slope is proportional to the positive bias voltage, and inversely proportional to the value of R. Hence the duration of this negative sawtooth may be adjusted by changing either the positive bias or the resistance of R.

The pulse of voltage appearing on the plate of the triode is shown in Fig. 18(d). The leading edge of this pulse may be sloped a little to acquire some delay, by making R_L large and thus allowing the stray capacitance on the plate circuit to integrate the slope. Further clipping of this signal eliminates the negative "pip" caused by the leading edge of the original pulse, and makes it suitable for a keying signal in a clamp circuit to operate on the "back porch."

Examination of the functioning of this circuit during the serrated vertical pulse in the standard RMA sync. signal shows the formation of keying pulses which are timed to coincide with the slots in the vertical pulse. Thus keying of the clamp circuit at black level is carried on with no interruption throughout the vertical sync. pulse.

Use of this type of clamp circuit in the switching system eliminates surges introduced by switching, and also any surges and low-frequency additive cross talk from other sources which may have been introduced ahead of the clamping point. It further insures constant sync. output under all conditions of varying picture signal by confining the sync. pulses to a fixed portion of the $e_g - i_p$ characteristic of the output stage.

Clipper Circuit

The clipping operation required in the process of correlating the peaks of blanking pulses with black level, illustrated in Fig. 15, must be performed with rigid accuracy. In order to maintain the clipping level with the necessary accuracy, it is imperative that the critical electrode in the clipper stage be controlled by a clamp circuit, as described in the preceding paragraphs. When, as is usually the case, a screen-grid tube is used as a clipper, it is further necessary to supply current to the screen grid from a regulated source so that its potential does not change with variations in the average brightness of the scene. With these two precautions, any of the usual types of clipper will maintain the correct black level in the blanking pulses.

The clipper circuit used in the field equipment has unusual characteristics which merit description. It is a circuit which was developed originally by K. R. Wendt[8] to overcome the inherent curvature near cut-off in the plate-current-cutoff type of clipper. Such curvature increases the gamma of the system. Since the orthicon type of camera tube has unity gamma, and the average kinescope has gamma in the neighborhood of 2, the resultant system gamma is higher than is ordinarily desired. Therefore, it is desirable to avoid increasing the gamma at any other point in the system.

[8] RCA Laboratories, Princeton, N. J.

The basic circuit (Fig. 20) includes a pentode amplifier, V_2, which has for its principal plate load a resistor R_2 in series with a diode, V_4. An additional load, R_1, is connected in parallel with R_2 and V_4. R_1 is much larger than R_2. The diode acts as a peak limiter, preventing the flow of current in R_2 whenever its cathode rises above the potential $E_b/2$. The tube, V_2, is the same as

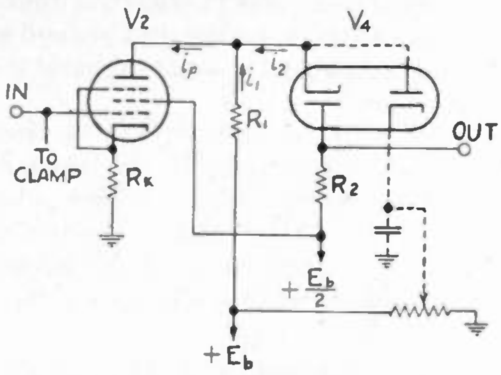


Fig. 20—Schematic diagram of linear clipper.

V_2 in the previous Figs. 16 and 17, with its control grid connected to a clamp circuit. Both of the supply voltages shown, E_b and $E_b/2$, are closely regulated and have approximately the 2-to-1 relationship indicated. Under these conditions, black level will correspond to some definite value of plate current in V_2. By adjusting the clamp reference potential (equivalent to adjusting the bias on V_2) to the proper value, black level may be made to coincide with point B in Fig. 21.

The curve A–B–D represents the normal e_g–i_p curve of V_2. From A to B, no current flows in R_2, and hence $i_p = i_1$. However, between B and D, current flows in V_4 and R_2, and hence i_p divides itself between R_1 and R_2. Therefore, $i_p = i_1 + i_2$ in this region of the curve. By proper selection of the value of R_1, it is possible to place point B (cutoff point of the diode, V_4) so that B-D covers the linear portion of A-B-D. Then at all times the useful part of the picture signal swings only over this linear part of the tube characteristic. The curved portion of the characteristic from A to B is not used, and hence does not affect the gamma of the system.

It should be noted that the cutoff of this circuit is extremely abrupt. This arises from the fact that V_2 is a screen-grid tube having a very high plate resistance. In other words, the flow of plate current is not influenced appreciably when the external load resistance is changed by the opening of the diode. Therefore, as i_2 approaches zero, the ratio i_1/i_2 rises very rapidly, and, since R_1 is large (about 20 to 30 times R_2), the potential of the cathode of V_4 also rises very rapidly, carrying the diode abruptly through its cutoff region.

The diode limiter, V_4, has one serious fault; namely, its plate-cathode capacitance permits feed-through of unwanted parts of the signal—particularly, steep wavefronts involving high frequencies. This trouble may be largely nullified by the simple device of connecting a second diode (available in any of the twin diodes) across the plate circuit of V_2, as shown by the dotted lines in Fig. 20. By proper adjustment of the bias control on this diode, it may be made to start conducting at a potential just above the cutoff potential of the limiter diode, thus causing a low-resistance shunt to appear across the signal source which effectively "squelches" the signal and prevents feed-through.

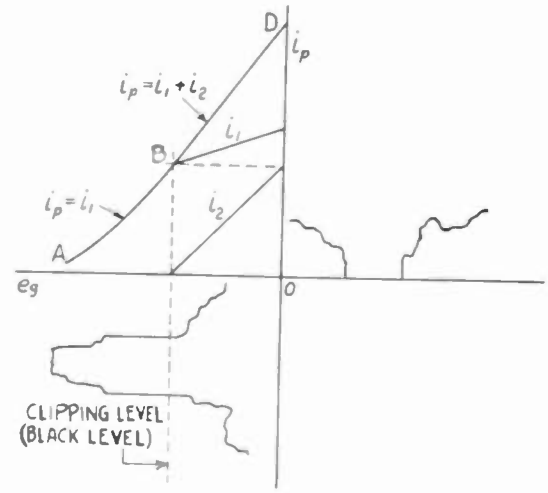


Fig. 21—Characteristic curve of linear clipper.

Acknowledgment

The equipment described in this paper is the result of the combined efforts of a large number of engineers with whom the writer has been associated during the past decade. Development of a workable system embracing so many complex circuits cannot be ascribed to the abilities and judgment of one or two, or even a few, persons. It is necessarily the product of the thinking of many individuals working collectively on the various problems. Among these are M. A. Trainer, W. J. Poch, H. N. Kozanowski, G. L. Beers, N. S. Bean, J. M. Brumbaugh, H. M. Potter, and F. E. Cone, of the RCA Victor Division, Radio Corporation of America, Camden, N. J., and R. D. Kell, A. V. Bedford, J. P. Smith, K. R. Wendt, and A. C. Schroeder of the RCA Laboratories Division, Radio Corporation of America, Princeton, N. J.

New C.B.S. Program Transmission Standards*

HOWARD A. CHINN†, FELLOW, I.R.E., AND PHILIP EISENBERG†

Summary—Over a period of years, broadcast listeners have complained that the musical portions of radio programs are sometimes unpleasantly loud—that is, music is too loud compared with speech. Two surveys conducted by the Columbia Broadcasting System in 1940 and 1944 found this to be true of all broadcast stations, and established the validity of the complaints. This led to more definitive studies.

Two related listener studies in 1945 were undertaken, to (1) discover proper (pleasing-to-listener) relative levels at which music and speech should be transmitted, and (2) determine the range within which the *peak* levels of a program should fall in order to please the largest number of listeners.

A total of 224 persons, representing a cross section of the radio audience, took part in individual listener tests. In both studies the listeners, one at a time, heard a series of passages from radio programs. They were asked to adjust the volume of each passage to the most pleasant listening level. Every variable which could be anticipated was provided for, including such matters as introducing controls to account for differences in room noise levels.

The major findings of the studies were as follows:

(1) Listeners prefer to hear broadcast music and speech at about the same peak levels (as read on a standard volume indicator).

(2) The limit of the range of *peak* volume levels tolerated by the largest number of listeners is approximately 8 db (4 db above or below the average peak-volume level of the passage.)

(3) Within this range (8 db) volume-level changes are less annoying when made gradually, in two or more steps.

The 8-db limit mentioned above refers to the range of *peak* or maximum volume levels, *not* to the range of minimum and maximum sound intensities or "dynamic range." It is important that this range of *peak* levels not be confused with "dynamic range."

In addition to the three principal findings, the studies uncovered data on a number of related points which obtained, irrespective of the sex, age, education, or musical taste of the individual listeners. For example:

(1) Listeners like to hear broadcast music and speech at the same *relative* levels, regardless of the *absolute* sound level that is preferred.

(2) Listeners prefer an even level, regardless of whether they are hearing variety, drama, narrative, or music.[1]

(3) The peak sound level that the average listener prefers ranges from 65 to 70 db above the acoustical reference level of 10^{-16} watts/cm^2.

This study has led to the adoption by CBS of a new set of program transmission standards, in order to make broadcasting more pleasing to the listener. The old transmission standards provided for maximum peak levels or "ceilings." The new standards retain "ceilings" (but different from past ones) and, in addition, provide minimum peak levels, or "floors," below which the level of the *main program peaks* should not fall.

I. Introduction

IT IS AXIOMATIC that broadcasting programs should be written and produced in a manner that makes for pleasant listening. Aside from the content of a given program, a broadcast may be easy or difficult to listen to, depending upon the method of presentation. Tonal range, sound reproduction level, and the range of peak volume levels are three of several factors that have a bearing on ease of listening. A former study[2] has shown that both tonal range and sound intensity influence listeners' preferences, and that listeners can distinguish more readily between different sound intensities than between different tonal ranges.

Broadcast receivers can be readily designed to provide the listener with means for selecting the tonal range and the sound-reproduction level he finds most pleasant. However, the listener is not in a position to materially alter the *relative* sound levels between different parts of a program.

Radio listeners have complained that, while listening to a program, they have to make volume adjustments because some parts of the program are broadcast too loud, and others not loud enough. The complaint has been registered for many years and to essentially all radio stations. As the Columbia Broadcasting System studied these complaints, it became clear that not all programs, nor even all programs of a certain type, were cited by the listeners. Certain programs were singularly free from such criticism. However, as many programs were monitored in an effort to isolate the causes of the complaints, it appeared that even the "non-offenders" occasionally offended.

The program transmission standards which have been in general use result in the broadcasting transmitters being modulated an equal amount, by both speech and music, when the prescribed peak levels are maintained. This practice insures the maximum use of the available power. From a purely technical standpoint, the procedure is sound. However, it does not take into consideration the listener's preferences.

From an analysis of listener complaints, it became clear that a searching study of the matter was needed. It was evidently necessary to determine accurately (1) what practices in broadcasting studios and control rooms were causing complaints of annoying sound level changes, and (2) what steps could be taken to remove the causes.

II. A Preliminary Study

Radio listeners most frequently complained that music was transmitted too loud as compared with speech. A preliminary study was conducted, therefore, to determine the relative levels at which listeners prefer to hear music and speech. The study was also designed

* Decimal classification: R550×R020. Original manuscript received by the Institute, December 31, 1946; revised manuscript received, April 7, 1947.

† Columbia Broadcasting System, Inc., New York, N. Y.

[1] These findings do not necessarily apply to symphonic music, which was not analyzed in these studies.

[2] H. A. Chinn and Philip Eisenberg, "Tonal-range and sound-intensity preferences of broadcast listeners," PROC. I.R.E., vol. 33, pp. 571–581; September, 1945.

so as to discover the effect on the preferred relative levels of (a) the room noise level, (b) the listeners' sex, age, education, and musical preference, and (c) the sound level at which they listened to the program.

Method

One hundred and thirty-two participants, one at a time, listened to eight passages taken from broadcasts. The passages ranged from 14 to 35 seconds in length. Half of the passages were music, and half were speech. Each subject was instructed to manipulate a control knob to adjust the volume of each passage to the level he preferred. The final setting for each passage was noted.

Half of the subjects participated in these tests with a room-noise level of 29 db (above the acoustical reference level of 10^{-16} watts/sq. cm.). This is representative of a very quiet suburban residence. The other half listened at room-noise level of 43 db, this being representative of the average residence.[3] The noise was introduced by a concealed electric fan whose presence was not noticed by any participant. The environment and equipment were essentially the same as for the main experiment which is described later.

Results

The results of this preliminary experiment were as follows:

(1) Listeners set the music passages at about the same peak level as the speech passages, indicating that they prefer to hear music and speech at approximately the same level.

(2) At the average room-noise level (43 db), listeners set the acoustical level from 5 to 7 db higher than in a relatively quite room (29 db). However, they preferred about the same relative levels for music and for speech for both room-noise conditions.

(3) Individual differences in sex, age, education, and musical taste (preference for popular or serious music) did not affect the desire for a relatively even level between music and speech.

(4) Individuals varied in their absolute sound-level preferences from 47 to 78 db, with an average of 64.5 db, in the "quiet" room, and from 58 to 83 db, with an average of 70.3 db, for the "average-room" noise conditions (see Table I). But no matter how high or how low individuals adjusted the absolute sound level, they still preferred music and speech at approximately equal levels. Thus, the problem of transmission standards is greatly simplified by the fact that individual loudness-level preferences do not have to be taken into account.

(5) The design of the experiment also permitted a determination as to whether the level at which a passage was introduced had any influence upon the participants' reactions. This was done by introducing one-half of the test passages at a relatively high level and one-half at a relatively low level. It was found that, when passages are introduced at a high level, listeners tend to set the volume higher than when passages are introduced at a low level. The difference, however, corresponded to the minimum change in level that the average person would notice when listening to program material at the sound levels involved. Thus, listeners adjust to the sound level they prefer, within their ability to discriminate. The level at which a passage is introduced does not influence the listener's desire for a fairly constant level between music and speech.

[3] D. F. Seacord, "Room noise at subscriber's telephone locations," *Jour. Acous. Soc. Amer.*, vol. 12, pp. 183–187; July, 1940.

TABLE I

INDIVIDUAL SOUND-LEVEL PREFERENCES

	At 29-db noise level	At 43-db noise level
Number of listeners	68	64
Index of sound level preference*		
80 to 84 db	0 per cent	3 per cent
75 to 79	7	16
70 to 74	10	44
65 to 69	34	29
60 to 64	27	6
55 to 59	16	2
50 to 54	3	0
45 to 49	3	0
	100 per cent	100 per cent

* An index of sound-level preference was computed for each individual. The index consisted of the median of 16 volume-level settings made by each during the course of the experiment.

Home Listening Experiences

An additional check was made on the complaints received in writing by radio stations by means of a questionnaire which was filled out by the participants in these tests. The results of this survey were as follows:

When listening to a given radio station at home, how often do you find that changes in volume occur which annoy you?

Regularly	5 per cent
Fairly often	22
Once in a while	48
Never	25
	100 per cent

What kind of volume changes during broadcasts annoy you most? (For those who complained of annoying changes.)

Music too loud	65 per cent
Commercials too loud	16
Applause too loud	14
Continual changes within the program	8
Introductory music too loud	7
Speech too loud	6
No data	1
	117 per cent[4]

[4] Totals more than 100 per cent because some listeners indicated more than one type of annoying volume change.

On what type of program do you find that such volume changes most often occur? (For those who complained of annoying changes.)

Variety	48 per cent
Drama	13
All programs	10
Serial drama	9
Miscellaneous	13
No data	7
	100 per cent

As many as 75 per cent of the respondents complained of the occurrence of annoying volume changes. The chief offenders were generally listed as music that was too loud and variety programs that had a wide range in levels. Evidently, the letters of complaint received by radio stations are representative of the feeling of most listeners that annoying volume changes occur in radio programs.

III. Plan of the Main Experiment

The preliminary experiment verified the letters of complaint and eliminated such factors as room noise and individual differences as experimental variables. It was therefore possible to design a study which would answer more specific problems related to broadcast program transmission standards. The main study was designed to discover:

(1) The range of peak levels acceptable to the listener;
(2) The rate of change tolerable within this range of peak levels;
(3) The relative peak levels at which various types of program material (speech, music, sound effects, applause, laughter) should be transmitted in order to please the largest number of listeners.

Method

In this experiment 92 participants, one at a time, listened to passages from radio programs. Perhaps the most concise way to outline the experimental procedure is simply to reproduce the instruction sheet given to each participant. This read as follows:

> We have invited you to come here to tell us at what volume levels you would like to hear your radio programs. We are asking you, as a member of a representative radio audience, to tell us what *you* like best. There are no right or wrongs answers. What you like is right.
>
> We are going to ask you to do something different from what you do at home when you listen to a radio program. At home you are usually seated away from the radio. If the program gets too loud or too soft, sometimes you do not change the volume because it is inconvenient.
>
> But here, the volume-control dial is right before you. It works just like the dial on your radio. We are going to play parts of radio programs, and we would like you to keep your hand on the dial while you are listening. Whenever the program gets too louds or too soft for you, even if it is only for a few seconds, change the dial to the level you like best.

After the participant indicated that the instructions were understood, a practice test was given. For this purpose, a portion of a radio program *not* used in the final test was reproduced. When it was established that the subject understood the instructions—to make an adjustment whenever the sound level became too high or too low for him—the test series was begun.

Each participant heard five samples, varying in length from $2\frac{1}{2}$ to $8\frac{1}{2}$ minutes. The samples were chosen from regular broadcasts and ranged from program excerpts with relatively small volume-level changes to those in which the volume varied considerably. A continuous, synchronized chart was automatically drawn by a moving-tape recorder of the volume control settings made by the participant.

Program Material

All test selections were recorded from broadcast programs, and were chosen to provide a wide variety. Material from five kinds of programs was used, namely, popular music, comedy quiz, comedy variety, narrative drama, and crime drama (see Table II).

TABLE II
PROGRAM PASSAGES

Passage Number	Type	Content	Duration
1	Popular music	Theme music, announcements (2 male, 1 female); "Just a Prayer Away," by Tobias (male and orchestra); "My Dreams Are Getting Better," by Curtis (female and orchestra)	4 minutes, 59 seconds
2	Comedy quiz	Comedy quiz (3 males, 1 female) Commercial (male announcer) Comedy music (orchestra)	7 minutes, 5 seconds
3	Drama	Opening Announcement (male) Theme music Conversation (2 males)	2 minutes, 17 seconds
3	Comedy-variety	Opening Announcement (male) Theme music Comedy (2 males) Comedy (3 males) Commercial (1 male) Comedy (2 males)	8 minutes, 31 seconds
5	Crime drama	Conversation (3 men in a car) A murder Conversation (2 men) Conversation (several men)	2 minutes, 34 seconds

Environment

The tests were conducted in a small room furnished as far as feasible like an average living room. The room measured 23 feet long, 14 feet wide, and 10 feet high. The average noise level in the room, without the loudspeaker operating, was slightly less than 29 db above the acoustical reference level.

The loudspeaker was located at one end of the room at a height approximating ear level of the listeners. The person being tested sat in an armchair, directly facing the loudspeaker and ten feet away from it. In front of the participant was a small table, on which there was a small box equipped with a dial for controlling the volume level of the selections heard.

Equipment

Uniformity in the program material presented to each listener was assured by the use of especially recorded "masters" cut on cellulose-nitrate coated disks. The recordings were made by the Columbia Recording Corporation and employed the standard electrical-transcription recording characteristics. Original master recordings were used because of the uniform response characteristic, the low nonlinear distortion, and the very low surface-noise level that this type of recording affords. As soon as any distortion or noise became detectable under the conditions of the tests, a new recording was used.

In order to simulate home listening conditions, a band-pass filter was placed in the reproducing channel. The actual over-all response of the system, including the recording and reproducing process, but excluding the loudspeaker, is shown in Fig. 1. The loudspeaker

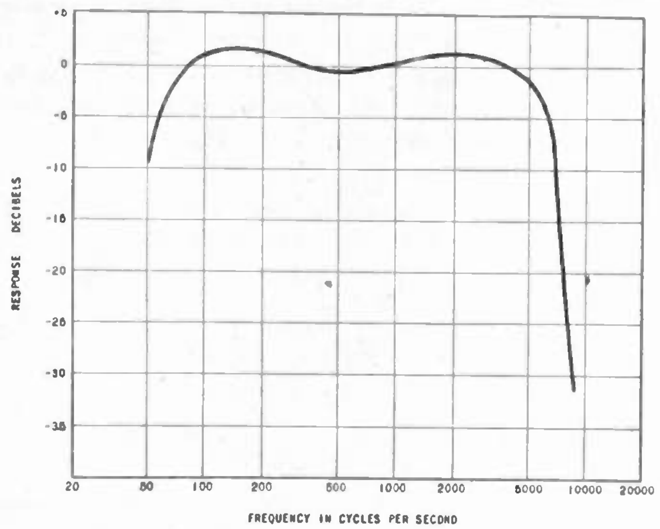


Fig. 1—Average home sound-reproduction conditions were simulated for this study by employing a band-pass filter in the test channel. The frequency-response characteristic of the channel (exclusive of the loudspeaker) is shown in this chart.

was a dual unit, employing a folded horn for the low frequencies and a multicellular horn for the high frequencies. Facilities were not available for a free-space calibration of the loudspeaker, but the manufacturers' measurements, which are believed to be reliable, indicate the system is uniform over a much wider range than was employed for this experiment.

TABLE III

COMPOSITION OF THE GROUP

Number of persons = 92	
Sex	
Male	55 per cent
Female	45
	100 per cent
Age	
Under 26 years	47 per cent
26 to 40 years	34
Over 40 years	19
	100 per cent
Education	
Grammar school	12 per cent
High school	61
College	27
	100 per cent

Although probably of secondary importance for the study undertaken, the harmonic distortion of the system was as low as it is possible to obtain with the best present-day equipment.

Subjects

The group of participants were representative of a cross section of the radio audience. They were secured by means of spot announcements over the CBS key station, WCBS, located in New York City. The exact composition of the 92 subjects, all adults, taking part in the individual listener tests are detailed in Table III.

IV. RESULTS OF MAIN EXPERIMENT

Method of Analysis

The data were analyzed from two points of view: (1) the actual volume changes within the passages as broadcast, and (2) the listeners' reactions to these volume changes.

A unit of analysis was established by dividing each of the five passages into a number of *intervals*. A new interval started whenever any clear change in volume occurred, or when the program content changed. Intervals varied in length from 3 to 20 seconds, although most intervals were of 5 to 10 seconds duration. In all, the five program passages were divided into 390 intervals.

The average peak level of each of the 390 *intervals* was determined, using a standard volume indicator,[5] by averaging the readings of two trained observers who, incidentally, agreed very closely. These readings were expressed as differences (in db) above or below the average level of the *entire passage*. Listeners' reactions were then analyzed in terms of *direction* and *amount* made by the listener with each interval.

Direction of Change

Although listeners could turn the volume up, turn it down, or make no adjustment at all, any changes they made were significant only in relation to the actual volume changes occurring in the passage itself. Therefore, the direction of the change in the passage had to be related to the direction of the listeners' adjustments.

The adjustments made by the listeners were therefore classified as follows:

(1) Counter-adjustments, which offset the change in volume of a passage.
(2) Pro-adjustments, which accentuated the amount of change in the volume of the passage.
(3) No adjustment.

In other words, listeners made counter-adjustments when variations in volume were too great for pleasant

[5] H. A. Chinn, D. K. Gannett, and R. B. Morris, "The new standard volume indicator and reference level," PROC. I.R.E., vol. 28, pp. 1–17; January, 1940. Also, *Bell Sys. Tech. Jour.*, vol. 19, pp. 1–44; January, 1940.

listening—that is, they turned the volume up or down to keep it within the range they preferred. They made pro-adjustments when they turned the volume up or down in the same direction as the volume changes occurring in the program material.

Amount of Change

Listener adjustments were also measured by amount of change. It was possible to measure the amount of change in db for each listener and for each interval. These changes were then averaged.

It is worth noting that, since a measurement was taken for each of the 92 test participants for each of the 390 intervals, a total of 35,880 individual reactions were analyzed.

Preferred Range of Peak Levels

Two typical examples of the way listeners reacted to changes in the peak levels during a program passage are shown in Fig. 2. In this figure, the line graphs show the

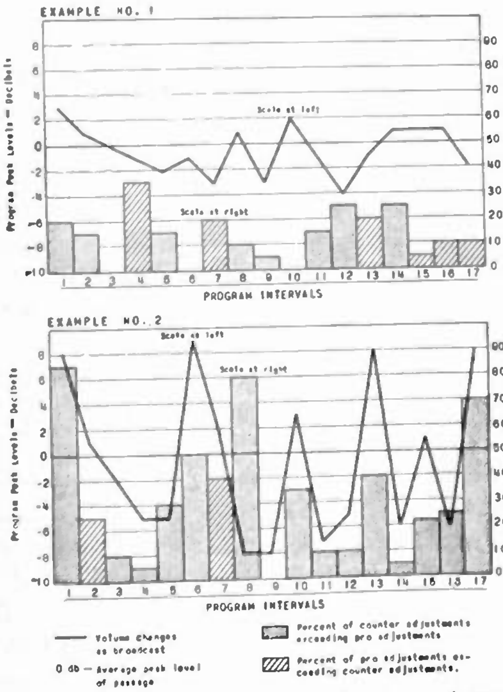


Fig. 2—The variations in peak volume levels, as shown by the line graphs in this chart, are typical of the program excerpts used in this study. In example No. 1 the range of peak levels is not very great, whereas in example No. 2 the range is rather large.

The vertical bars show the extent of the reaction of the listeners to these volume changes. The bars show the difference between the proportion of listeners who made counter-adjustments and those who made pro-adjustments.

variations in average peak levels, as broadcast, of two of the program passages listed in Table II. In the first example, the maximum variation in peak levels was only 7 db; 3 db above to 4 db below the average for the passage. In the second example, on the other hand, the variation was 17 db; 9 db above to 8 db below the average.

The vertical bars in Fig. 2 show the reaction of the listeners to these volume changes. The bars show the difference between the proportion of listeners who made counter-adjustments and those who made pro-adjustments.

In the first example the volume variations were slight, and few of the listeners found it necessary to make volume adjustments. On the other hand, the second example, containing marked volume changes, caused large numbers of listeners (in some instances more than 50 per cent) to make adjustments that offset these changes.

TABLE IV

Distribution of Program Intervals

Deviation of interval from the average level of the passage	Number of intervals
9+ db	15
8	20
7	17
6	31
5	26
4	43
3	51
2	74
1	79
0	34
Total	390

Fig. 2 presents the data for only a few typical intervals. In all the test passages, the intervals totaled 390. The distribution of these intervals, broken down

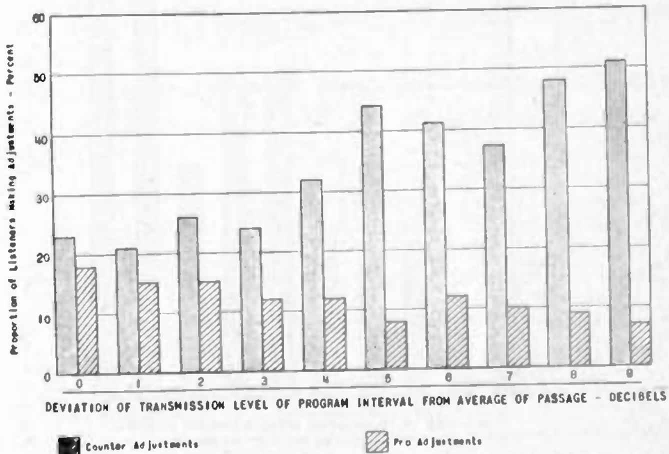


Fig. 3—This chart shows the proportion of listeners making volume adjustments as a function of the deviation of the transmission level of the program interval from the average of the complete passage. The limit beyond which listeners appear to object to large volume variations seems to occur at 4 db above or below the average of the passage.

according to the differences (in db) from the average levels of the passages in which they occurred, is shown in Table IV.

The proportion of listeners who made counter-adjustments and the proportion who made pro-adjustments for each of these groups of intervals is shown in Fig. 3. It is seen that when the intervals were transmitted at the average volume level of the passages (0 db), more than half of the listeners made no adjustment at all. Those who made adjustments did so to offset or to accentuate the change in roughly equal proportion (23 per cent made counter-adjustments; 18 per cent made pro-adjustments). In other words, these changes tended to cancel out.

Fig. 3 shows that the farther the average peak level of an *interval* lies from the average of its *passage*, the greater is the listeners' tendency to offset it—that is, to narrow the volume range down to the limit of pleasant listening. The limit beyond which listeners appear to object to large volume variations seems to occur at 4 db above or below the average of a passage. (At this point, more than 30 per cent of the listeners made counter-adjustments.)

Since listeners evidently will tolerate a variation up to 4 db above or below the average of a passage, *the range of listener tolerance for peak levels appears to be no more than 8 db.*

The foregoing analysis was made by *direction* of adjustment. It remains to analyze the *degree* of adjustment made by the listeners. These data are presented in Fig. 4 which shows by how many db the average listener who made adjustments moved the volume-control dial.

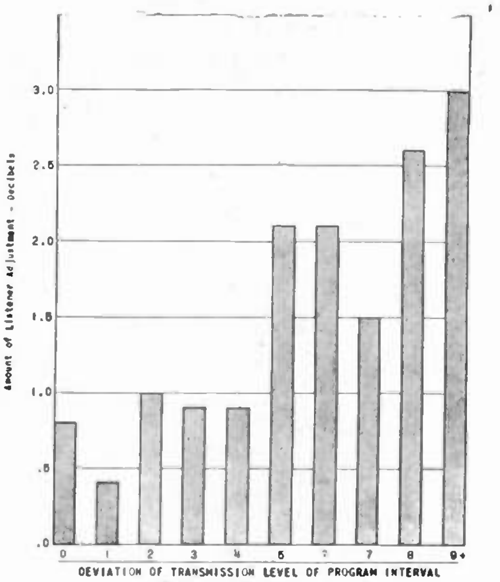


Fig. 4—The extent of the volume-level adjustments made by those listeners who made a change is shown in this chart as a function of the deviation of the transmission level of the program interval from the average of the complete passage. Again it is seen that beyond the 4-db point there is an abrupt break, and for greater deviations listeners tend to counter-adjust by larger and larger amounts.

It is seen that where the deviation of the peak level of an interval is 4 db or less from the average for the passage, the amount of adjustment made by the listeners was less than 1 db. Beyond the 4-db point, however, there is an abrupt break, and for greater deviations the listeners tend to counter-adjust by 2 or more db. It is also evident that the farther a peak level lies from the average peak levels of a passage, the greater the adjustment.

Tolerable Rate of Change in Peak Levels

Thus far, the analysis has been limited to listener preferences in terms of peak sound intensity, in relation to the average peak level of passages. This does not however, take into account the complexity of volume changes as they occur in radio programs. At least two other factors are significant in influencing listener reactions: (a) The *amount* of volume change from interval to interval, and (b) the *direction* of the volume change from interval to interval.

An illustration of the effect of these factors can be seen in Fig. 2(b). Interval 10, while it is only 3 db above the average of the passage, is 11 db above the peak level of the preceding passage. A relatively large proportion of listeners reacted to this change (the difference between the percentage of listeners making counter-adjustments and the percentage making pro-adjustments was 35 per cent). This illustrates the importance of the amount of volume change from interval to interval.

Furthermore, the direction of the volume change is of some consequence. In the same example, interval 8, which is 8 db below the average, has moved 17 db, in two steps, from a preceding interval. From a peak at interval 6, the volume dropped in the *same* direction to interval 8.

On the other hand, intervals 13 through 17 illustrate the situation when volume changes zigzag back and forth; listeners' adjustments are in contrary direction. In order to unravel these complexities, it was necessary to analyze listener adjustments of intervals in terms of the two factors mentioned above.

The amount of volume change from interval to interval was divided into three categories: (1) large = more than 6 db; (2) moderate = 3 through 6 db; and (3) small = 2 or fewer db.

For *direction* of volume change, intervals were divided into those which continued in the same direction as the preceding interval and those which changed direction.

However, the analysis would not be complete unless the *distance* of peak levels from the average were also taken into account. That is, some intervals were either very high or low in volume, as compared with the average for the passage, while others were close to the average of the passage. Therefore, intervals were further subdivided into those which were a large distance from the average of a passage (4 or more db) and those which were a small distance (3 or few db).

The great mass of data resulting from the analysis of all participants' reactions to the 390 intervals involved can best be presented in chart form. Fig. 5 presents the data in terms of the proportion of listeners making adjustments. This figure shows the proportion of listeners making counter-adjustments and those making pro-adjustments where the volume change between intervals is large (more than 6 db), moderate (3 to 6 db), and small (less than 3 db). The left-hand column of bars is for those intervals where the volume level of the interval changed in the same direction as in the preceding interval. The right side of the chart is for those intervals where the volume level changed in a contrary direction to that of the preceding interval.

From these data it is seen that:

(1) The direction of volume change influences listener adjustments. A greater proportion of listeners made counter-adjustments and few make pro-adjustments when volume changes occur in the same direction as the preceding interval.

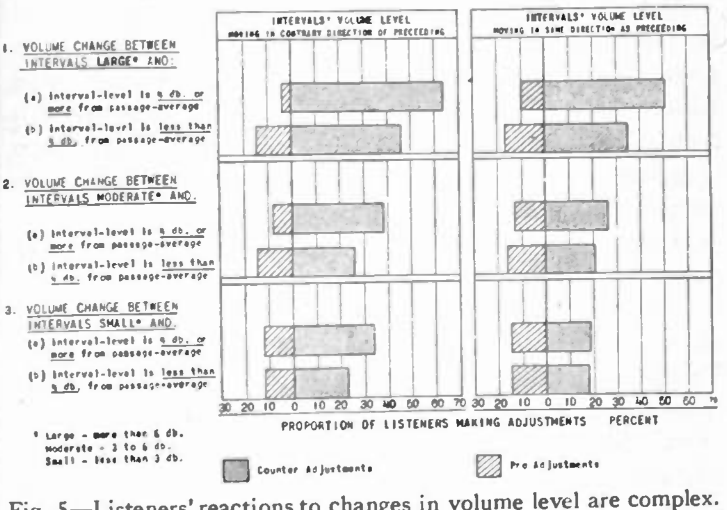


Fig. 5—Listeners' reactions to changes in volume level are complex. As this chart shows, they are influenced by (a) the direction of the volume change, (b) the degree of the volume change, and (c) the relative volume level with respect to the average of the passage.

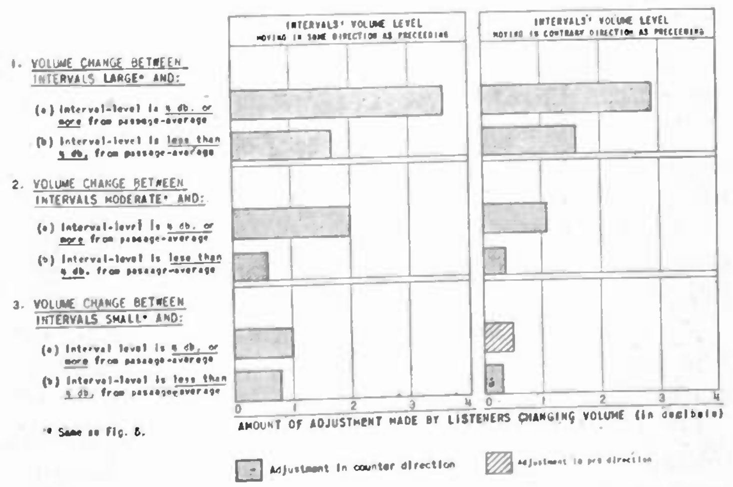


Fig. 6—From the *amount* of the volume adjustment made by listeners, it is seen that not only is a fairly narrow range of *peak* levels preferred, but, in addition, within that range large changes are not liked. Changes in the peak volume level must be made gradually for easy listening.

(2) The degree of volume change from interval to interval also influences listener adjustment. The larger the change, the more counter-adjustments.

(3) Distance from the average peak level of a passage influences adjustments. The greater the distance, the more counter-adjustments.

The intervals in which most listeners make counter-adjustments are those which move in the same direction as the preceding interval and which, at the same time, represent a large volume change from the preceding interval. These intervals are also farthest removed from the average level of a passage. The fewest changes are made in intervals with reverse characteristics.

The data, in terms of average db change, for those listeners who make adjustments, are given in Fig. 6. It is seen that direction of volume change, *amount* of volume change, and *distance from average level* also influence the amount of this compensation made by the listener. These data show that the listener not only prefers a fairly narrow range of peak levels (not more than 8 db) but, in addition, within that range does not like large changes. In other words, the whole range of 8 db cannot be jumped in one step. *Volume changes must be made gradually.*

Influence of Program Content

Thus far, listener preferences have been covered only in terms of relative volume levels without regard to the program content. In general, radio programs can be broken down into three types of content, (1) speech, (2) music, and (3) laughter, applause, and sound effects (grouped together as a "special" type of content.) The test passages were therefore segregated into these three categories and the peak levels analyzed, both as originally broadcast and as adjusted by listeners during this study.

The average difference from the average peak-levels of the passages as originally broadcast was as shown in Table V.

TABLE V

Type of Program	Average Difference from Average Level of Passage
Speech	−0.6 db
Music	+0.7 db
Laughter, applause, and sound effects	+3.9 db

A survey[6] made on a wider basis scrutinized the practices of a variety of programs as broadcast by various networks, and verified that the program samples used in these experiments were typical.

The average peak level of all music intervals in these experiments was 1.3 db higher than the average of all speech intervals. This difference corresponds closely with the former standard practice, wherein music was peaked at 100 on the volume indicator and speech at 80. This corresponds to a difference of 1.8 db.

In order to relate the volume differences (as broadcast) to the adjustments which the listeners made, their

[6] Unpublished report, CBS General Engineering Department, April, 1945.

reactions to each type of program content were analyzed in terms of the distance of each interval from the average of the passage. This analysis shows listeners' adjustments to music and speech intervals of the same loudness.

Fig. 7 shows the proportion of listeners making counter- and pro-adjustments for intervals of various differences from the average of all passages.

Listeners turn the volume up for below-average musical intervals in about the same proportion as they do for speech. The reaction to intervals containing

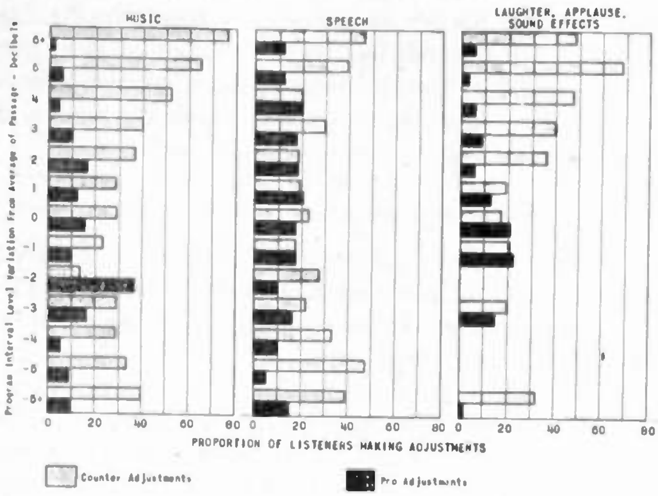


Fig. 7—The reactions of listeners to different types of program material is shown by this chart. Listeners' reactions to musical passages, speech and laughter, applause and sound effects follow the same general pattern.

laughter, applause, and sound effects is similar to the reaction to music.

The evidence of this study supports this conclusion: For radio program material (exclusive of symphonic music, which was not studied) listeners prefer a fairly even level for all types of program content.

V. Discussion

The listener's preference for a limited range of peak levels differs sharply with the ideas of some radio program directors, conductors, and performers (and some scientists). They have sometimes felt that a restricted range of peak levels unduly limits the dramatic and musical effects that can be achieved. This feeling on the part of producers, conductors, and performers may be a carry-over from the stage and concert halls, in which, for reasons to be mentioned, a wide range in peak intensities probably does not irritate the listener.

Broadcasting, however, is an entirely different medium. It differs from the stage and the concert hall in at least two important respects: (1) It is intimate. The listener is in a small room and cannot tolerate shouts and loud orchestral crescendos. (The listener, moreover, frequently considers his neighbor's feelings, as well as his own.) (2) Listening to the radio is a monaural process. That is, although the listener actually hears with both ears, the sound comes from a single source and all sense of direction, and almost all sense of perspective, is lost. The listener to "real-life" sound, on the other hand, enjoys binaural listening. He can pay attention to what he *wants* to hear and discriminate against undesirable sounds. The radio listener, however, must listen to all the sounds that come from the loudspeaker.

Although the reason is not yet known, monaural hearing is apparently more pleasant when the range of peak intensities is somewhat restricted. The manner in which sound is used in motion pictures supports this fact. Motion-picture sound, also monaural, is generally reproduced at a narrower range than that which is transmitted by radio.

It is essential that this monaural peculiarity of radio be taken into account. Although it calls for a restricted range of peak intensities, the desired dramatic effect may be gained in another dimension—*timbre*. Varying the distances of speaker or musical instrument from the microphone will effect the apparent loudness, even while maintaining an even level as measured on the volume indicator, because of the change in liveness of the pickup. The opportunity for such effects should be exploited to the fullest degree.

VI. Conclusions

The conclusions resulting from the two studies reported are:

(1) Listeners prefer to hear music and speech at about the same peak levels (as read on a standard volume indicator).

(2) Listeners like to hear music and speech at the same relative levels, regardless of the absolute sound-intensity level that they prefer.

(3) Listeners prefer an even level regardless of whether they are hearing variety, drama, narrative, or music.[7]

(4) The limit of the range of *peak* volume levels tolerated by the largest number of listeners is approximately 8 db (4 db above or below the *average volume* level of the passage).

(5) Within this range (8 db) it appears that volume-level changes are less annoying when made gradually—in two or more steps.

The 8-db limit mentioned above refers to the range of *peak* or maximum volume levels, *not to* the range of minimum and maximum sound intensities or "dynamic range." It is important that this range of *peak* levels not be confused with "dynamic range."

[7] These findings do not necessarily apply to symphonic music, which was not analyzed in these studies.

As a result of this study, CBS has adopted the following program transmission standards:[8]

CBS Program Transmission Standards

(1) *Speech and Music*	*VI Peaks*
Normal passages	Peaks of 100
Low-level passages	Not less than 40
(2) *Theme under station breaks*	Peaks of 40
(3) *Applause and audience reaction*	Maximum peaks of 70

(4) *Transition*

The transition from a low-level passage to a normal-level passage (or vice versa) must be in steps of not more than 4 db, preferably less (i.e., peaks of 40, then 60 and finally 100, or vice versa). Similarly, two succeeding passages (voice, then music, or voice, then a sound effect, etc.) must not differ in level by more than 4 db, preferably less, even when a contrast is intentional.

(5) *Peaking Practice*

Peaking program material according to the prescribed standards means "gaining" in such a manner that the maximum VI peaks reach the specified values as frequently as possible without being inconsistent with the program content. It is understood that occasional peaks beyond the prescribed values are unavoidable, but these must be kept to a minimum.

VII. Practical Test

Theoretically, the adoption of new program transmission standards based upon the foregoing conclusions should remove the cause of listener's complaints arising from former practices. To check this, a practical before-and-after test was made. Three pairs of program excerpts were used in these tests. In each pair the excerpt was produced in two ways—the old way, with wide range in *peak* volume levels, and the new way, in accordance with the findings of the study. Thus, listeners were enabled to express their direct preference for one type of transmission or the other. For the test the selections used were as follows:

(1) A passage containing a loud, but typical orchestral bridge.
(2) A loud but typical opening followed by conversation in low tones.
(3) A passage containing a loud scream.

[8] It is hardly necessary to state that the standards contained in this report may not remain static indefinitely, since they are based upon current broadcast-pickup techniques. Future operations and experience may indicate that permanent or temporary departures from them are desirable, and actual practices may, from time to time, vary from these standards. They represent, however, the conclusions which have been reached as a result of the investigation which is reported upon in this paper.

The preferences for the old and the new transmission standards are shown in Fig. 8. It can be seen that in two of the three cases there was an overwhelming preference for the new standards.

The third test selection emphasized still another condition which plainly contributes to the sources of complaint mentioned earlier. In this test, production rather than engineering played the major part. The director realized that some of the life-like quality of the scream would be lost if the technician held the transmission level within the prescribed limits by "fading-down" the output of the microphone. To avoid this loss of dramatic effect, the producer accepted an annoyingly loud peak in the original production of the program.

The desirable solution, which would have retained the dramatic effect and eliminated the high peak, would have been to move the performer farther from the microphone. Advantage was not taken of this technique in the third practical test; rather, the levels were

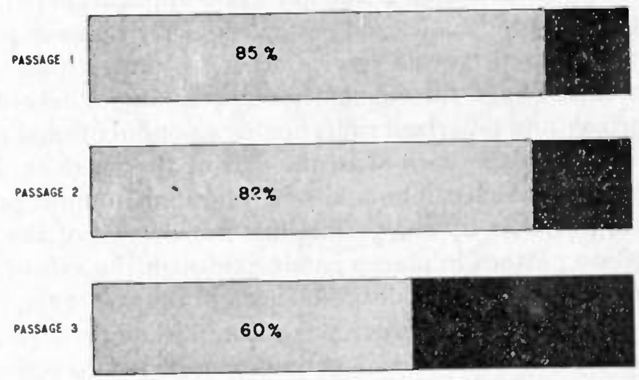


Fig. 8—The proportion of listeners favoring the new standards over the old, in a practical test involving three typical passages, is shown in this chart. In passage 3, changes in production techniques would have been required (as explained in the text) in addition to the new transmission standards in order to make the passage more acceptable.

simply faded-down. Even so, 60 per cent of the listeners preferred the new transmission practices to the old.

An even more practical test has been the introduction of the new transmission standards to programs originated by the Columbia Broadcasting System. This was done during 1946, and the number of listener complaints on this score has markedly decreased. Furthermore, listeners have not felt that there has been diminution of dramatic value in program production. Rather, they have experienced a more even transition between segments of program.

VIII. Acknowledgment

Studies of this type require a great deal of detailed preparation and care to insure reliable data. Without the patience, ingenuity, and skill of A. G. Peck and D. C. Battin, it would have been difficult to have successfully completed this study. The authors gratefully acknowledge their contribution.

"Cloverleaf" Antenna for F.M. Broadcasting*

PHILLIP H. SMITH†, SENIOR MEMBER, I.R.E.

Summary—The radiation requirements and general design considerations for transmitting antennas suitable for f.m. broadcasting are briefly discussed, and an explanation of the design and operation of the arrangement of radiating elements and associated feed system employed in the "cloverleaf" antenna is given. Both calculated and measured data are included, showing field-intensity distribution, gain, impedance-frequency characteristics, etc. Design features which are discussed include a simple coaxial impedance-matching transformer developed initially for microwave application, and the method and facilities provided for the removal of sleet.

Introduction

THE CLOVERLEAF antenna was engineered particularly to meet the demands of a relatively new and rapidly expanding phase of the radio broadcasting industry—f.m. broadcasting. In a broad sense the cloverleaf design is not limited in application to f.m., to radio broadcasting, or to the frequency range at present allocated to this service. It is, in fact, a practical antenna design for efficiently transmitting or receiving horizontally polarized radiation in a nondirectional pattern in a plane normal to the axis of the antenna. The field gain which can be achieved is limited for most practical purposes by the permissible narrowness of the radiation pattern in planes passing through the axis of the antenna, and by the desired length of the structure.

The development work was accomplished through the use of 1/10-scale model antennas designed to operate in the frequency range of 880 to 1080 Mc. The scale models are themselves entirely serviceable antennas for operation at these higher frequencies.

The cloverleaf f.m.-broadcasting antenna is characterized principally by its rugged construction, uniform horizontal-plane radiation characteristics, simplicity due to combining structural and electrical members, and its ability to handle high power. No insulators are required throughout.

Radiation Requirements

One of the principal considerations in designing a transmitting antenna for any specific application is the required radiation-field-intensity distribution. This can usually be represented by two separate radiation patterns, one showing the distribution of field intensity in the horizontal plane and the other in the vertical plane. In general the field-intensity-distribution requirements in the horizontal plane are largely dictated by the location of the antenna with respect to the receiving locations, and in the vertical plane by the propagation characteristics of the frequency employed over the transmission paths and distance involved.

Considering first the horizontal-plane-radiation requirements for f.m.-broadcast transmission in the 88- to 108-Mc. band, it is at once seen that, unlike long-range communication systems, the radiation must be directed to receivers located over a wide area in the general vicinity of the transmitting station. Thus, in this plane the required pattern usually will be circular in shape.

In the vertical plane the energy should, if possible, be beamed parallel to the ground, since propagation at these high frequencies is due primarily to the direct "line-of-sight" transmission path, and therefore energy radiated at angles appreciably above the horizon is wasted. Fortunately, since the wavelengths involved are short in comparison to the practicable length of a radiating structure, it is possible to obtain a fairly high degree of directivity in the vertical plane with relatively small and simple structures.

A third radiation requirement which has been standardized for f.m. broadcasting is that the radiation shall be horizontally polarized.[1]

Antenna Design Considerations

The increase in field intensity in the direction in which an antenna is designed to emit maximum radiation, compared to the field intensity which would be produced by a half-wave antenna supplied with the same power, is defined as its field gain. The power gain is the square of this value. Neglecting losses and assuming that the individual radiating units are (a) nondirectional in azimuth, (b) properly spaced, and (c) properly excited, the power gain of almost any vertical antenna array of radiating units will be the same for the same over-all array length in wavelengths. Furthermore, the gain will be a linear function of the array length. Thus one of the most important criteria for good antenna design becomes largely a measure of the extent of electrical and mechanical simplification which can be incorporated into the design.

One of the simplest forms of antenna which radiates a horizontally polarized electric field of uniform intensity in the horizontal plane is the horizontal loop.[2] Fundamentally, the loop type of antenna comprises a single conducting turn or loop in which r.f. current is induced in some manner to flow continuously and uniformly, both in amplitude and phase, around the loop.

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† Bell Telephone Laboratories, Inc., Whippany, N. J.

[1] "Standards of good engineering practice concerning f.m. broadcast stations," Federal Communications Commission, September 20, 1945.

[2] A. Alford and A. G. Kandoian, "Ultra-high-frequency loop antennas," *Trans. A.I.E.E.* (*Elec. Eng.*, December, 1940), vol. 59, pp. 843–848; December, 1940.

The loop diameter is significant chiefly because at certain critical diameters the radiation in the plane of the loop drops to a null.[3]

The over-all gain of a stacked array of horizontal loops is substantially independent of the individual-loop directivity for loop diameters between 0.2 and 0.6 wavelength, so that within this range it is possible to allow mechanical and other considerations to dictate the diameter of the loop employed. The uniform peripheral current in a horizontal loop antenna can also be altered in certain ways without bringing about impairment of the desired circular horizontal-plane radiation pattern. If, for example, the currents at corresponding points in each of a number of individual elemental radiators constituting a loop are made equal and in-phase, then essentially circular distribution of the radiation in the plane of this equivalent loop will still prevail.

"Cloverleaf" Design

Individual elemental radiators constituting an equivalent loop are most conveniently excited from a common feed line, in parallel, to establish the desired current phase and amplitude relations. An arrangement which fulfills this requirement comprises a cluster of four half-wave curved radiating elements arranged in the pattern of a four-leaf clover as shown in cross section on Fig. 1.

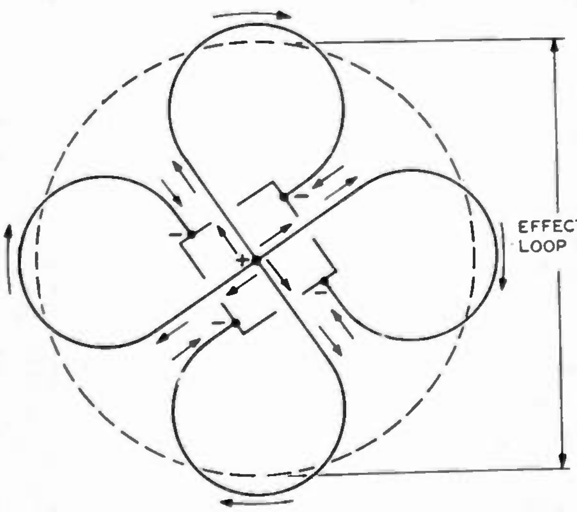

Fig. 1—Arrangement of four radiating elements constituting a radiating unit of the cloverleaf antenna. Arrows indicate assumed instantaneous directions of current.

One end of each of the radiating elements is connected to a common central conductor of a single coaxial line, while the other ends are bolted solidly to each of the four posts of a lattice tower structure which serves as the outer-return conductor. Maximum potential difference along the individual curved radiating elements will exist across their two extreme ends. The potential at the two ends is unbalanced, however, with respect to the "ground" plane, but the current distribution in the radiating elements is approximately sinusoidal. Due to this configuration of the elements, there is coupling between them which has an effect upon their combined impedance-frequency characteristic. This, as well as the unbalanced excitation voltage, requires that each radiating element be approximately 20 per cent short of a half-wavelength to present a purely resistive load, i.e., to be antiresonant.

The assumed instantaneous current directions in the various radiating elements of the cloverleaf assembly are as indicated by the arrows on Fig. 1. It will be observed that radial components of current in adjacent radiating

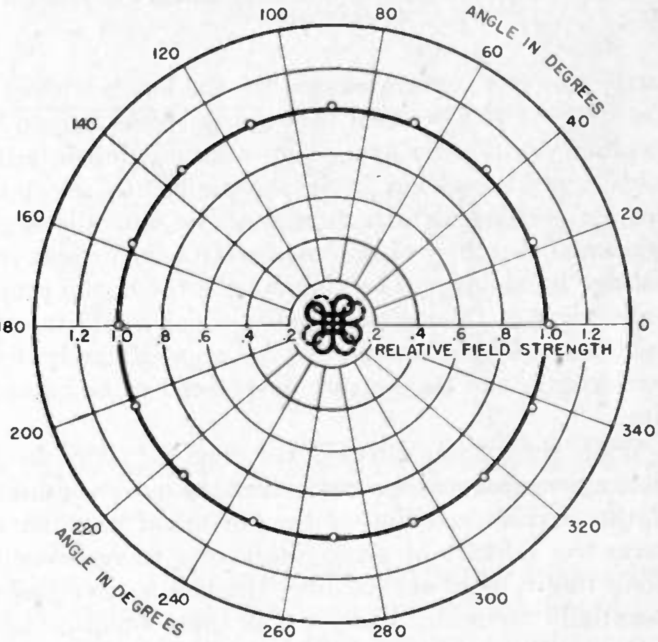

Fig. 2—Azimuth-plane-radiation pattern of a four-unit scale-model cloverleaf antenna. Measured points show variation from a circle to be within the limits of experimental accuracy.

elements flow in opposite directions and radiation therefrom will therefore tend to cancel, whereas peripheral currents are in-phase around the loop. The resultant horizontal-plane radiation pattern has been found to be circular at all frequencies within the allocated f.m. band to within the accuracy of measurement (about 2 per cent), as indicated on Fig. 2. The elevation-plane radiation pattern of a single horizontal radiating unit is, as expected, approximately cosinusoidal in all planes.

Fig. 3 shows a stacked array of radiating "cloverleaf" units all of which are effectively in parallel, inasmuch as they are located electrically $\frac{1}{2}$-wavelength apart along the coaxial feed line. The direction in which the individual radiating elements constituting each unit are curved is reversed for each adjacent radiating unit to compensate for the 180° phase reversal which occurs at half-wavelength intervals along the main transmission line. The radiating units are attached to the tower posts at proper intervals according to the operating frequency of the station. The proper spacing has been found to be 94.5 per cent of the half-wavelength, since the velocity of propagation within the

[3] D. Foster, "Loop antennas with uniform current," Proc. I.R.E., vol. 32, pp. 603–607; October, 1944.

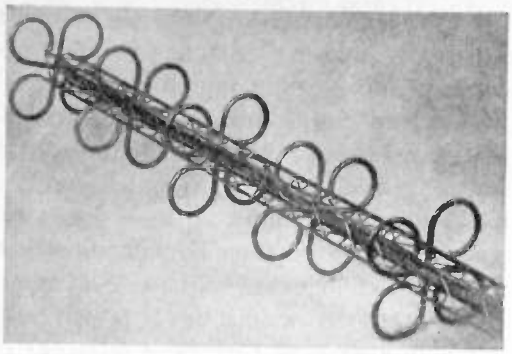

Fig. 3—Stacked array of five radiating units of a scale-model cloverleaf antenna.

lattice-tower structure caused by the loading effect of the struts is 94.5 per cent that of free space. Should the required position for any given radiating unit interfere with a strut position, it is permissible to select the nearest opening for attachment of the unit allowing a minimum clearance of approximately 1 inch to prevent voltage breakdown. The maximum error in the proper positioning of the radiating units which could thereby exist is negligible, amounting to approximately 1/50 wavelength, and such errors never need to be cumulative.

While the loading effect of the diagonal struts in the lattice-tower structure constituting the outer conductor of the coaxial feed line of the cloverleaf antenna reduces the velocity of propagation of a wave traveling along this type of coaxial line, the line is nevertheless essentially "smooth," inasmuch as these small periodic variations in capacity and inductance are only about 1/20 wavelength apart. Spacings of 1/10 wavelength are, for example, considered sufficiently close for the bead insulators in standard RMA coaxial line for these frequencies.

Radiation Resistance and Impedance

The cloverleaf configuration of radiating elements, each of which is about 0.4 wavelength long at a frequency of 98 Mc., results in an effective loop diameter of approximately 0.3 wavelength. The radiation resistance of a hypothetical uniform-current loop 0.3 wavelength in diameter has been calculated to be 130 ohms. This is altered to some extent by the mutual effect of adjacent loops, and the change in the loop radiation resistance is a function of the loop spacing.

Fig. 4 is a plot of the introduced resistance in a hypothetical uniform-current loop 0.3 wavelength in diameter due to the presence of adjacent loops, as computed by W. H. Wise in an unpublished work. The net effective radiation resistance of each loop is its initial value of 130 ohms modified by adding algebraically the contribution from each of the other loops.

One requirement for maximum antenna gain is that the current in all the radiating loops shall be equal. Although the radiation resistance of each equivalent "loop" or radiating "unit" of the cloverleaf antenna

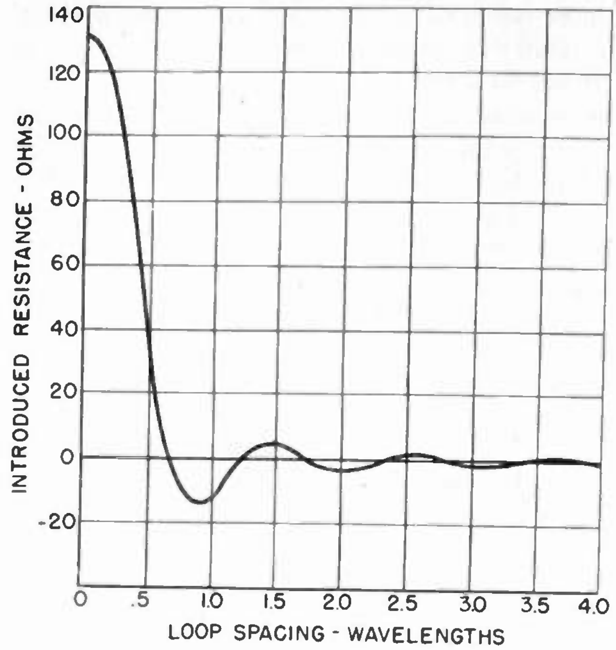

Fig. 4—Introduced resistance in a hypothetical uniform-current loop 0.3 wavelength in diameter, due to the presence of a similar adjacent loop.

will vary somewhat because of mutual impedance between the units, the voltage across the common junction of the four elemental radiators constituting each radiating unit will be the same at each unit position along the feed line, since the units are electrically $\frac{1}{2}$-wavelength apart. The effective or equivalent loop current, however, will vary in accordance with the way in which the radiation resistance is related to this driving voltage. This is a complicated relationship involving distributed circuit constants which depend on the exact geometrical configuration of the elements, and is difficult to evaluate mathematically. It may be seen, however, that there is a tendency for the effective loop current to be independent of the radiation resistance and directly a function of the driving voltage only.

An approximation of the problem is indicated in a series of four steps shown in Fig. 5, proceeding from left to right.

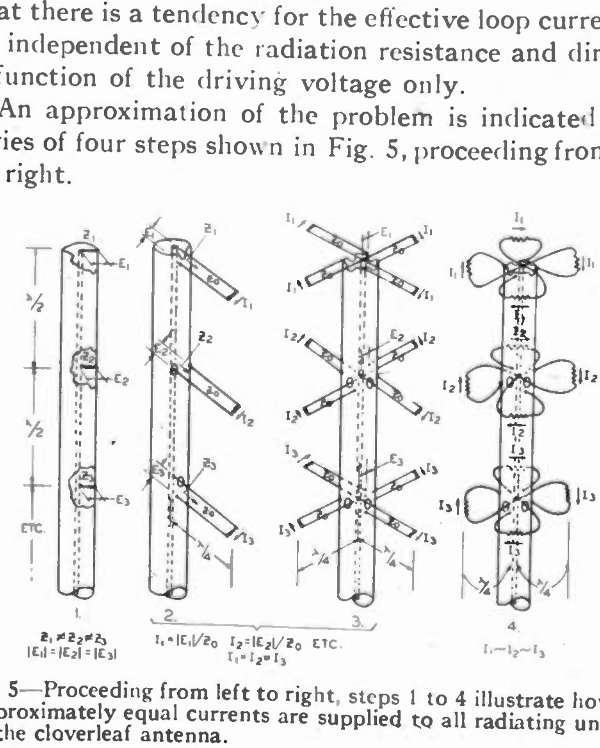

Fig. 5—Proceeding from left to right, steps 1 to 4 illustrate how approximately equal currents are supplied to all radiating units of the cloverleaf antenna.

1. As has been indicated, if a lossless transmission line is loaded with shunt impedances at intervals of any integral number of $\frac{1}{2}$ wavelength, the voltage across these impedances will be identical and independent of the magnitude or phase angle of the impedances.

2. If in place of these impedances $\frac{1}{4}$-wavelength lines are substituted, which are terminated in load impedances having any magnitude or phase angle, then as a result of a well-known property of $\frac{1}{4}$-wavelength lines the currents in these load impedances are equal in magnitude and independent of the magnitude or phase angle of the various load impedances.[4]

3. If any number of additional $\frac{1}{4}$-wavelength lines are likewise bridged across the main line in parallel with the foregoing $\frac{1}{4}$-wavelength lines, the currents in the load impedances terminating these additional lines are likewise identical in magnitude and independent of the magnitude or phase angle of any of the load impedances.

4. Since the radiation from radial components of the cloverleaf-antenna configuration is partially canceled, the largest portion of the total radiation takes place near the maximum-current points on the periphery of the loop. The radiation resistance may therefore be assumed to be lumped at these points, which are approximately $\frac{1}{4}$ wavelength from the driving ends of the radiating elements, and accordingly the conditions may be expected to approach the previously described conditions of step 3 wherein equal currents are known to exist.

The correlation in the amplitude of the first minor lobes between measured and computed radiation patterns, assuming equal current distribution in the loops, indicates that a close approximation to the condition of equal current distribution, and consequently optimum gain, is actually realized (see Fig. 7).

The antiresonant impedance of each radiating unit comprising a cluster of four radiating elements is found to be of the order of 400 to 650 ohms, depending upon the position of the unit in the array. The variation of its impedance with frequency is indicated on Fig. 6 by the curve labeled "one unit." Inasmuch as the phase angle of the impedance is relatively unimportant, a single radiating-element length has been found suitable for the entire f.m. band.

The variation of the resultant impedance with frequency, as measured across the terminals of the lowest radiating unit for two, five, and eight radiating units in which the spacing is optimum for 98 Mc., is as shown on Fig. 6 by the correspondingly labeled curves. The data as plotted is "normalized" with respect to the 100-ohm characteristic impedance of the lattice-tower-coaxial feed line. These curves indicate the actual operating impedance across the lowest radiating unit only at a frequency of 98 Mc. and its sideband frequencies, since in practice the spacing of the radiating units and the length of the overhanging suppressor rods are set at optimum dimensions for each operating frequency. They are, however, representative of the frequency-selectivity characteristics of the antenna.

[4] These principles were utilized in a "turnstile" antenna designed by J. F. Morrison, Bell Telephone Laboratories, Inc., U. S. Patent No. 2,350,916, June 6, 1944.

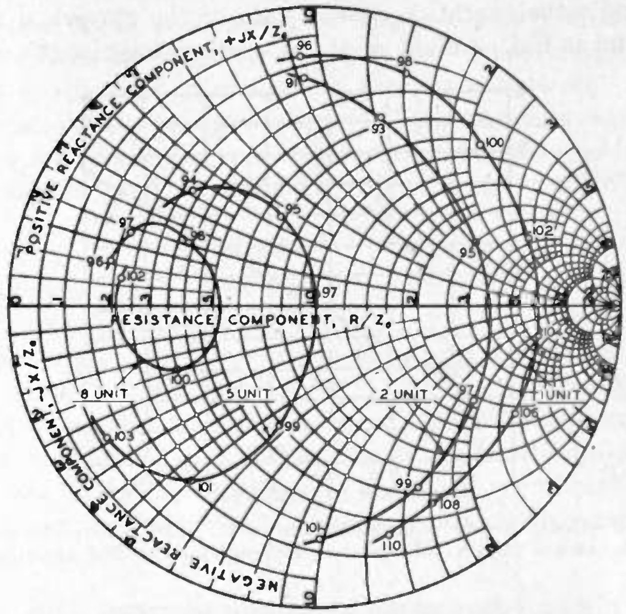

Fig. 6—Measured variation of normalized impedance with frequency across the terminals of the lowest radiating unit for one, two, five, and eight radiating-unit cloverleaf antennas when unit separation is optimum for 98 Mc.

A two-slug tuner is used to match the antenna impedance across the lowest radiating unit to that of the main coaxial feed line. Its operation and adjustment are described later.

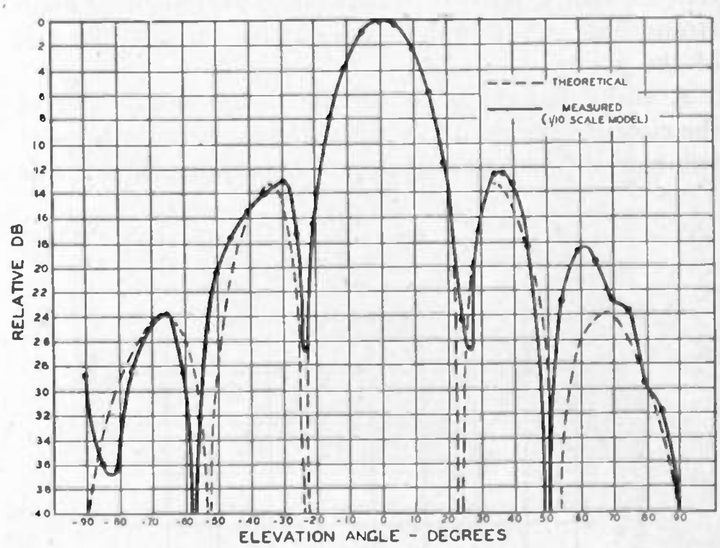

Fig. 7—Calculated and measured free-space vertical-plane field-strength pattern of a five-unit cloverleaf antenna.

Antenna Gain and Radiation Patterns

The power gain G of the cloverleaf antenna is given by the following empirical relationship, in which the error with respect to theoretical gain is less than $\frac{1}{2}$ of 1 per cent, if n is greater than 1:

$$G = 0.565n + 0.18. \tag{1}$$

This is plotted on Fig. 8. The gain of a single radiating unit is 0.88 with respect to a dipole, or 1.43 with respect to an isotropic radiator. Since the radiating units are always spaced at a constant-fractional part of the operating wavelength, the power gain of the cloverleaf antenna is independent of the operating wavelength and

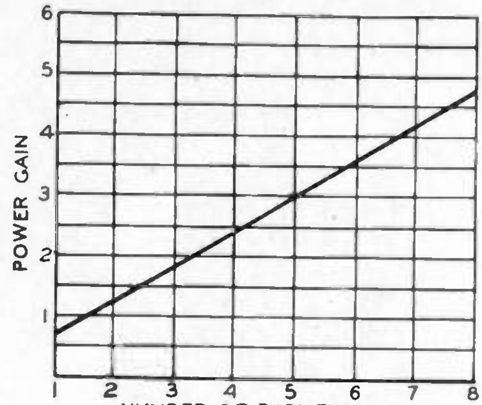

Fig. 8—Computed free-space-antenna power gain, referred to a dipole, versus number of radiating units of a cloverleaf antenna.

is a function *only* of the number of radiating units. To provide maximum gain the maximum number of radiating units which a given over-all antenna length will accommodate at the specified spacing is generally used. Gain measurements on scale-model antennas have confirmed the theoretical gain to within 0.2 db.

The vertical-plane radiation pattern may be calculated in the conventional manner employed for calculating the pattern of a linear in-phase array of equal current elements, each of which has a cosinusoidal field-strength distribution in planes passing through the axis of the array.

A simplified equation for the beam width (Φ) of the cloverleaf antenna, as measured between half-power points as a function of the number of radiating units, is

$$\Phi = \frac{107}{n}. \tag{2}$$

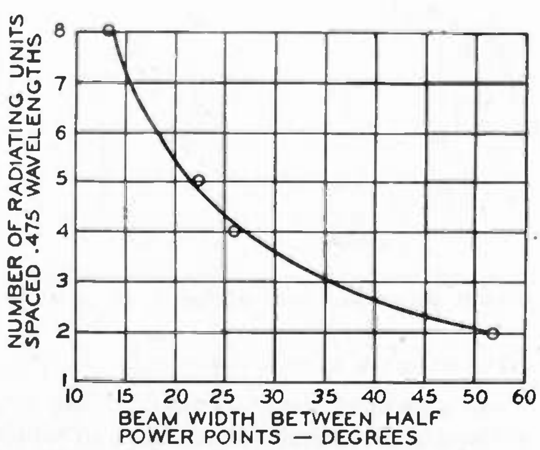

Fig. 9—Calculated and measured vertical-plane beam width between half-power points versus number of radiating units of a cloverleaf antenna.

The maximum error in beam width as obtained from (2) with respect to the theoretical beam width as obtained from a calculated field-intensity plot is negligible if n is greater than 1. The theoretical beam width, as a function of the number of radiating units, is plotted on Fig. 9. The points indicated thereon are from scale-model measurements.

Suppression of Spurious Radiation

Unless precautions are taken, longitudinal currents will be induced in the outer surface of the lattice-tower structure by the unbalanced potential at the two ends of the curved radiating elements. This induced longitudinal current gives rise to undesired vertically polarized radiation, as is the case, for example, when a dipole antenna is excited at its center by a coaxial transmission line.

Referring to Fig. 10, it will be observed that the instantaneous direction of the induced longitudinal current flowing on the outer surface of the coaxial lattice-tower structure is opposite in direction to the feed-line

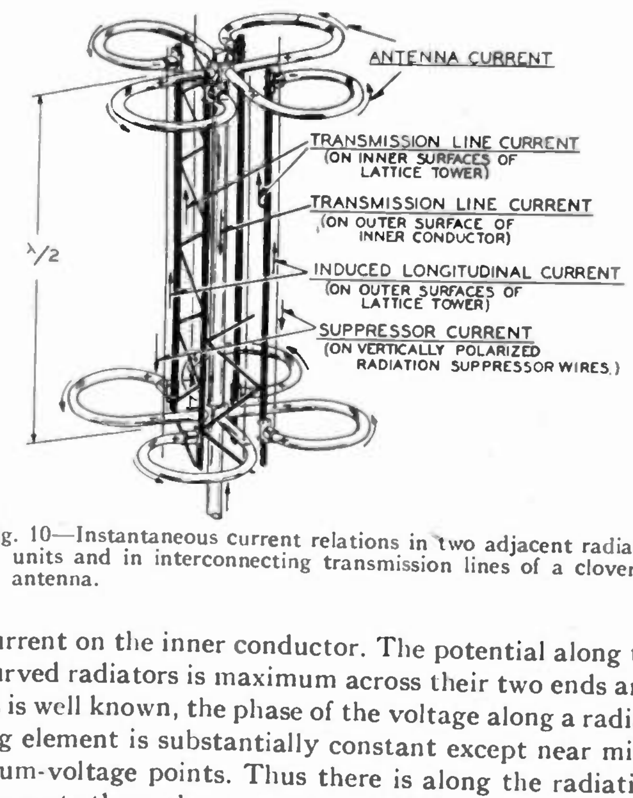

Fig. 10—Instantaneous current relations in two adjacent radiating units and in interconnecting transmission lines of a cloverleaf antenna.

current on the inner conductor. The potential along the curved radiators is maximum across their two ends and, as is well known, the phase of the voltage along a radiating element is substantially constant except near minimum-voltage points. Thus there is along the radiating elements themselves a source of voltage external to the tower structure of the proper phase and of a variable amplitude which may be used to drive the necessary neutralizing current through external conductors paralleling the lattice-tower structure to cancel the vertically polarized spurious radiation caused by induced longitudinal currents.

The four "suppressor wires," one paralleling each face of the tower, are adjusted to points along the radiators where the voltage is of the proper magnitude to provide substantial cancellation of this unwanted radiation. Maximum suppressor current is obtained at a tap-off point close to the tower structure, and maximum current is obtained in each wire as it is moved out along the radiating element to a voltage-null point.

To obtain maximum radiation efficiency, precautions must also be taken to prevent the flow of induced longitudinal current in those portions of the tower above the top and below the bottom radiating units. For this purpose a similar means has been employed, which in effect comprises an extension of the suppressor wires previously described for $\frac{1}{4}$ wavelength above the top and below the bottom radiating units. These extensions are terminated on the four outer faces of the lattice structure, as shown on Fig. 11. The current in these quarter-wave extensions is 180 degrees out of phase with the

Fig. 11—Quarter-wave rods used to suppress induced longitudinal currents in the outer surface of a lattice-tower structure.

induced longitudinal current in the overhanging ends of the lattice structure, and its action is such as to cancel unwanted vertically polarized radiation from these parts of the lattice structure. The optimum diameter of the quarter-wave extensions has been found to be greater than that of the wires, but their point of attachment to the radiating elements can be made the same.

In practice it has been found that the diameter and spacing of the suppressor wires is not critical. The conductor actually used between radiating units is $\frac{1}{4}$-inch stranded galvanized cable, and this is positioned $3\frac{1}{8}$ inches away from the tower face. The quarter-wave extensions are $\frac{5}{8}$-inch-diameter galvanized-iron rod.

A measured loss in antenna gain of from 1 to 3 db is observed when the suppressor wires are omitted. The horizontally polarized radiation pattern is, however, unaffected. This loss varies with the length of the antenna and with its particular mounting arrangement. It is generally greater for two- to four-unit antennas than for five- to eight-unit antennas. Measurements of the radiation pattern of a five-unit antenna without the wires shows that the vertically polarized field pattern has two maximums which correspond closely to the pattern about an unterminated wire 2 wavelengths long (a distance equivalent to the space between the top and bottom radiating unit). When the suppressor wires are added, the loss in gain is recovered to within the accuracy of measurement (approximately 0.1 db).

Impedance Matching

A two-section coaxial "slug tuner," located in the base section of the antenna shown on Fig. 12, is used to eliminate standing waves on the main coaxial feed line coming from the transmitter. The "slugs" are enlarged-diameter sections supported on the 3-inch inner conductor. The slugs are adjustable (a) in position along the line, and (b) in separation, the proper combination resulting in the desired impedance match. No other adjustments are required.

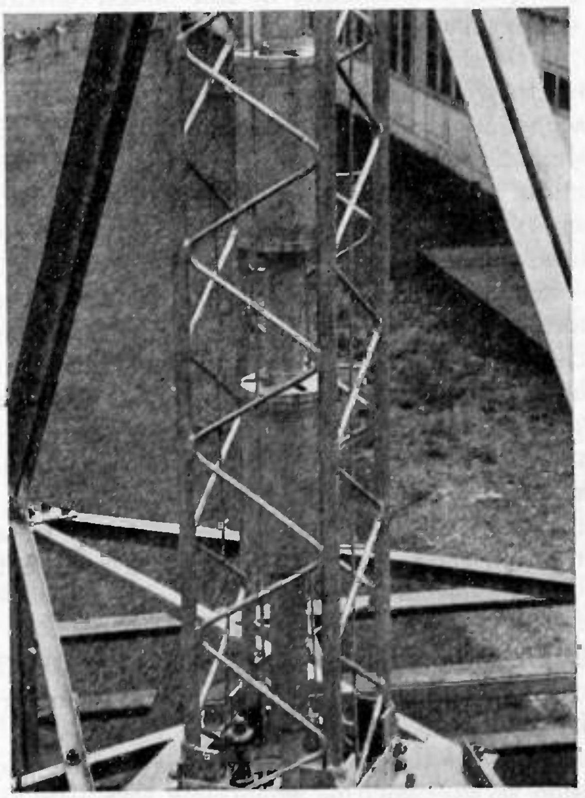

Fig. 12—Two-slug coaxial-line tuner located in the base section of a lattice-tower structure.

The impedance-matching capability of this type of transformer increases as the length of the slug is increased from zero to $\frac{1}{4}$ wavelength, and also increases as its characteristic impedance decreases. For any fixed slug length and slug characteristic impedance, its range of adjustment permits the elimination of standing waves of any position along a line and of any amplitude ratio between unity and a particular maximum value. The dimensions selected permit elimination of standing waves along the main coaxial feed line from a two- to eight-unit cloverleaf antenna at any operating frequency within the f.m. band. These slugs are $\frac{1}{8}$ wavelength long at the middle of the band and have a characteristic impedance of about 40 ohms.

Mechanical Features

The lower end of the 3-inch-diameter center conductor is restrained from vibration in high winds by a $\frac{1}{4}$-wavelength-long coaxial "metallic-insulator" support stub comprising a $1\frac{3}{4}$-inch-diameter section of galvanized-iron pipe which passes coaxially through the center of the base plate of the antenna up through the inside of the 3-inch-diameter inner conductor for approximately $\frac{1}{4}$ wavelength, at which point it mushrooms out to the inner diameter of the 3-inch conductor and is solidly attached thereto. The $1\frac{3}{4}$-inch pipe is clamped in the center of the base plate. This support stub also serves effectively as an even-order harmonic shunt across the main coaxial transmission line.

Steel is used throughout the antenna structure, since all parts may conveniently be zinc-plated (galvanized) to a depth substantially exceeding the depth of penetration of the high-frequency currents. The skin depth (depth to which the current penetration is $1/\epsilon$ of its surface value) is, for zinc, approximately 0.0006 inch at 98 Mc., and hot-dip galvanizing commonly applied to structural steel is approximately 0.003-inch thick. The resistivity of zinc is approximately equivalent to that of brass, and, although this is about three or four times that of copper, the large surfaces which may be used for all antenna conductors reduce the current density sufficiently to keep I^2R losses to a negligible value.

The tower structure is made in sections 9 feet, 8 inches long, which can be bolted together as required upon installation, and the inner conductor is likewise in sections which can be bolted together to make a complete assembly. A 300-millimeter code beacon may be mounted on the top of the structure, if required.

Sleet-Melting Facilities

The prevention or elimination of ice from all parts of the cloverleaf antenna has been found to be important, particularly when heavy icing is experienced. The construction of the antenna, however, makes the application of sleet melting facilities a relatively simple matter. De-icing of the radiating elements themselves is accomplished by means of electrically operated Calrod heating elements which are inserted into the "grounded" ends of the loops (see Fig. 13), the power connections being brought up through a conduit which is clamped at short intervals to the inside corners of the tower structure.

For de-icing the remainder of the antenna a method similar to that used by the power companies for thawing frozen water pipes is used. The antenna is connected in

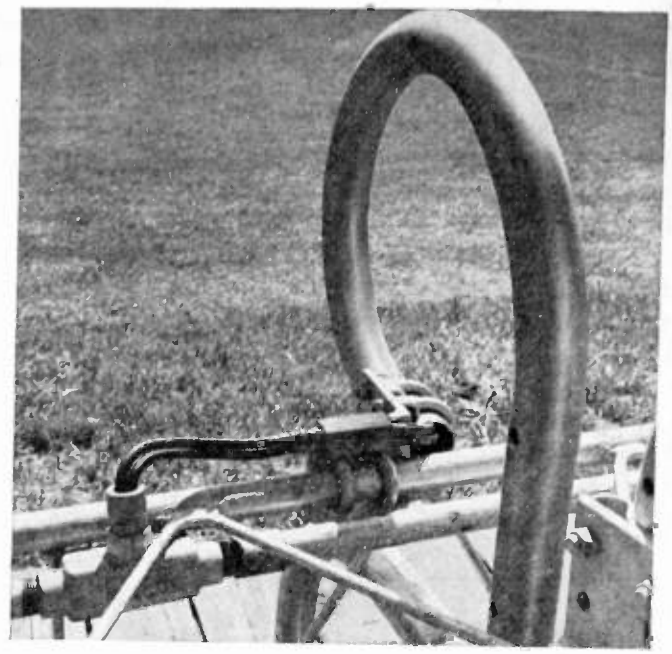

Fig. 13—Single radiating element of a cloverleaf antenna, showing a calrod heater element for sleet removal.

a series circuit for 60-cycle current by inserting insulating gasket material at appropriate loop connection points. A low-voltage current transformer is then connected between the center-conductor quarter-wave support stub and the base of the antenna. The gasketed connections provide an effective r.f. by-pass capacitor and accordingly permit simultaneous high-frequency operation of the antenna. The reactance of these "by-

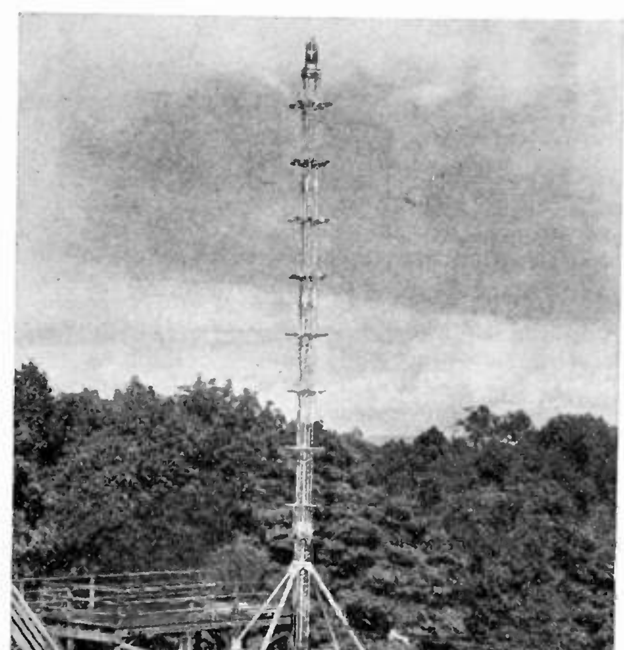

Fig. 14—A completely assembled prototype of an eight-unit cloverleaf antenna erected on a special stand for testing.

pass capacitors" is very low at 98 to 108 Mc., and consequently their power factor is of no importance.

The heat-dissipation requirements to prevent the formation of ice on the antenna have been investigated. The United States Weather Bureau was consulted regarding temperature and wind conditions under which sleet forms throughout the country. Their records show that, with the exception of mountain-top locations such as Mount Washington, sleet practically never forms when the temperature is below 10° F. Furthermore, a wind velocity of 20 m.p.h. is rarely exceeded when sleet is forming. Weather Bureau records also show that the average interval during which sleet actually forms is short, seldom exceeding about two hours. However, the greater time required to melt ice after it has once formed makes it desirable to have de-icing equipment in operation in advance of a possible storm.

A dissipation of approximately $\frac{2}{3}$ watt per square inch was determined, experimentally, to be sufficient to prevent the formation of ice under the above temperature and wind conditions. This was also found to be close to the dissipation required to just remove ice which had already formed under the same conditions. A total de-icing power of 21 kw. for an eight-unit cloverleaf antenna satisfies the above requirements. Proportionately less power is required for shorter antennas.

A photograph of a completely assembled prototype of an eight-unit cloverleaf antenna erected on a special stand for testing is shown in Fig. 14.

Acknowledgment

The writer would like to express his appreciation for the help afforded by all of his associates, and in particular, by E. L. Younker and E. H. Karleen.

Theory and Design of Progressive and Ordinary Universal Windings*

MYRON KANTOR†, ASSOCIATE, I.R.E.

Summary—Using as a basis the previous papers by Simon on the subject of progressive and ordinary universal coils, their theory is extended to improve the accuracy by taking additional factors into account.

The present paper offers a more thorough treatment of the subject by deriving accurate results, and by employing theoretical expressions to replace previously required empirical rules. In addition, due to a certain convenient change of definition, equations are derived for practical use which are considerably simpler and, at the same time, more accurate than those given by Simon. It is shown that the proper number of throws of wire per coil revolution is a function of the coefficient of friction between the surface of the coil form and the insulation of the wire.

To avoid confusion, the symbols used by Simon are also used here, and for the sake of completeness and a minimum of cross reference, the entire analysis is presented, including the derivation of formulas for the rate of progression and the gear ratio. A brief description of the geometry of the winding is included, and, finally, a section is devoted to an outline and discussion of practical design procedure.

I. Introduction

THERE ARE two varieties of the so-called "universal" coil winding now in common use; the progressive universal winding, and the more widely used ordinary or stationary universal winding. These have been treated previously in the literature, the most detailed analysis being that of Simon.[1,2]

Universal coils are wound by using a special machine in which the wire is fed onto a rotating cylindrical coil form by a guide or shuttle which is oscillated parallel to the coil axis, so that the wire lies in a regularly defined zigzag path around the circumference. A familiar example of this style of winding appears in a spool of twine. The rotation of the shaft supporting the coil form is eventually translated into the linear back-and-forth motion of the wire guide by means of a gear train and heart-shaped cam. The principal problem involved in the design of these coils is the determination of the proper gear ratio required to specify the angular rotation of the cam with respect to that of the coil. In the ordinary universal winding the only motions are the rotation of the coil and the displacement of the guide, so that the coil builds up to a sizable height and has a rectangular cross section. In the progressive universal winding there is, in addition, a uniform axial displacement of the coil form occurring simultaneously with the other two motions so that the coil can not become very high, but instead more nearly resembles the familiar solenoid winding. Universal windings are used principally where low distributed capacitance is required in a coil with a large number of turns.

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[1] A. W. Simon, "On the winding of the universal coil," Proc. I.R.E., vol. 33, pp. 35–37; January, 1945.

[2] A. W. Simon, "On the theory of the progressive universal winding," Proc. I.R.E., vol. 33, pp. 868–871; December, 1945.

II. Description of Winding

The developed pattern of a typical progressive universal winding is illustrated in Figs. 1 and 2. The winding surface is shown cut axially along a cylindrical element, and then spread out flat. The wires lie in straight-line paths because the displacement of the wire guide is directly proportional to the angular rotation of the drive shaft. In Fig. 1, the wire starts its motion relative to that of the coil form at point A, arriving at B after the cam has revolved through an angle of π radians corresponding to one throw or guide displacement c.

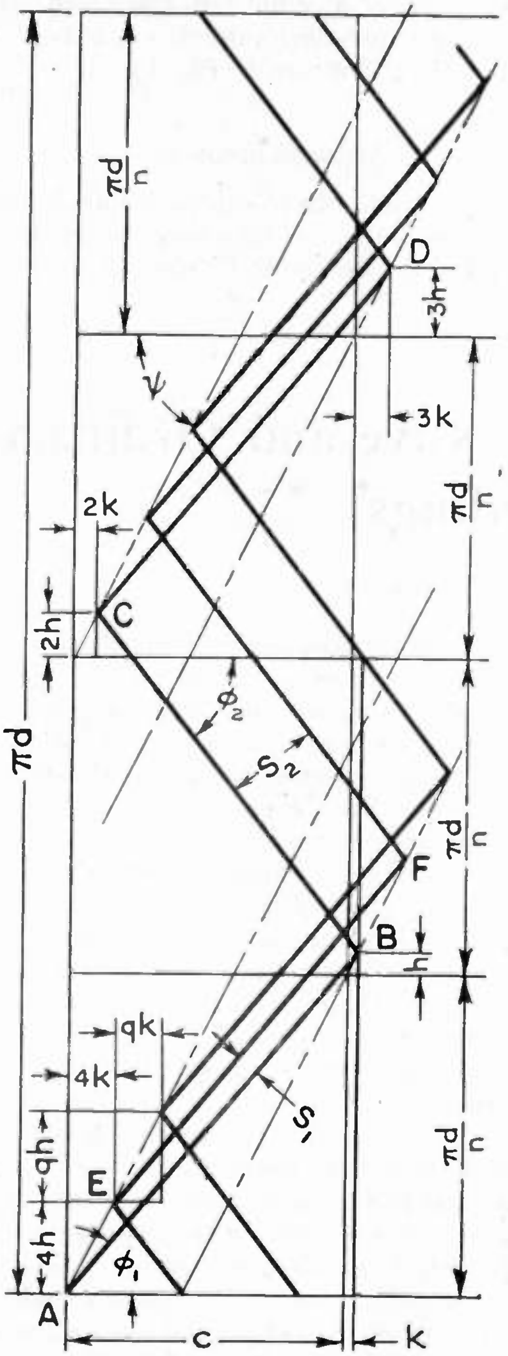


Fig. 1—Developed plane diagram of a progressive universal coil with progressive layering.

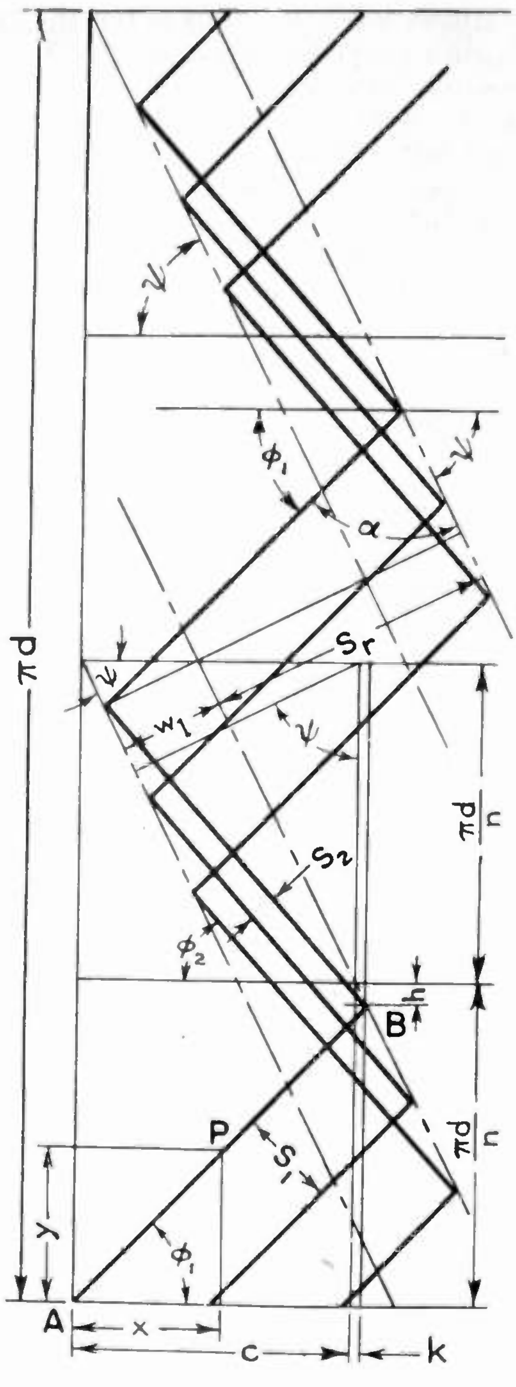


Fig. 2—Developed plane diagram of a progressive universal coil with retrogressive layering.

The circumference of the coil form is divided into n equal parts, where n is the nominal number of throws per coil revolution. In the illustration, n is taken equal to four. The axial displacement of point B from point A is greater than the cam throw c by the small distance k, which is the amount that the coil form moves in an axial direction during the time required to complete one throw. The winding is so arranged that point B does not lie exactly at $\pi d/n$ from the starting line, but is located so that it is either advanced or retarded by the small amount h. This must be done to insure the subsequent overlapping of the wires so that the winding may become firm and self-supporting. In Fig. 2, where B is retarded, the layering is called retrogressive, and in Fig. 1, where B is advanced, the layering is called progressive.[3]

The wire is laid down (Fig. 1) along the path B-C-D-E-F, etc. When the second throw is completed at point C, the circumferential displacement becomes $2h$ and the axial displacement becomes $2k$. Similarly, the displacements from the nominal conditions become $3h$ and $3k$ at the end of the third throw, $4h$ and $4k$ at the end of the fourth throw, etc. The first winding cycle is completed when the wire reaches point E and is about to fall adjacent to the first throw. The second cycle is exactly like the first, except that it originates from point E instead of A. It should be observed that the number of throws completed in a winding cycle is always an even integer. The dashed lines in Figs. 1 and 2 indicate the loci of the bend points of the wire, thus forming the helical ridges around the coil. They are a natural feature of the progressive universal winding.

In previous papers the statement has been made that, in the time of the forward throw AB, the wire guide and the coil form move in the same axial direction, and that during the backward throw BC they move in opposite directions. However, since the time required for each stroke is the same, the axial wire displacements must be proportional to the relative speeds prevailing during the two throws. Because the wire displacement evidently is greater during the forward throw, it follows that the larger relative velocity also must occur for this throw. This can be true only if the two motions during the forward stroke are in opposite directions.

III. Derivation of Gear-Ratio Formulas

The symbols used in the derivation are as follows:

n = nominal number of throws per coil revolution, expressed as a simple fraction, q/v, where q is always chosen as an even integer of the least possible magnitude. (For example, if $n=1.5$ or $3/2$, then $q/v=6/4$.)

q = number of throws per winding cycle

v = number of coil revolutions per winding cycle

d = diameter of the coil form, inches

c = cam throw corresponding to one-half revolution of of the cam, inches

n' = precise number of throws per coil revolution

h = circumferential displacement per throw of a bend point of the wire from its nominal location, inches

k = axial displacement per throw of a bend point of the wire due to the axial motion of the coil, inches.

s_1 = spacing between centers of adjacent turns of wire produced on the forward throw, inches

s_2 = spacing between centers of adjacent turns of wire produced on the backward throw, inches

θ_c = angular rotation of cam and cam gear, radians

θ_d = angular rotation of coil and drive gear, radians

r = ratio of number of cam gear teeth to number of drive gear teeth $=\theta_d/\theta_c$.

R = ratio of number of drive gear teeth to number of cam gear teeth $=\theta_c/\theta_d=1/r$

μ = coefficient of static friction between the surface of the coil form and the insulation of the wire

ϕ_1 = winding angle between an element of the coil form and the direction of a forward throw of wire, radians

ϕ_2 = winding angle between an element of the coil form and the direction of a backward throw of wire, radians

p = rate of axial progression of the coil form in inches per coil revolution

ψ = angle between the axis of the coil and the direction of the helical ridges

σ = ratio of the width of a helical ridge to the mean distance between ridges.

When the wire travels from one point to another, such as from A to P in Fig. 2, its circumferential displacement is $y=\pi d(\theta_d/2\pi)$, and its axial displacement is $x=(c\pm k)(\theta_c/\pi)$ where the positive sign indicates the forward throw and the negative sign the backward throw. Since θ_c/θ_d is equal to the gear ratio R, the quotient y/x, which is the tangent of the winding angle, becomes

$$\tan\phi_1 = \pi d/2R(c+k) \tag{1a}$$

$$\tan\phi_2 = \pi d/2R(c-k). \tag{1b}$$

From Figs. 1 and 2, however,

$$\tan\phi_1 = [(\pi d/n) \pm h]/(c+k) \tag{2a}$$

$$\tan\phi_2 = [(\pi d/n) \pm h]/(c-k) \tag{2b}$$

where the upper sign indicates progressive layering and the lower sign retrogressive layering, *a convention which will be used in all subsequent equations.*[4] The gear ratio is purposely expressed as R, rather than its reciprocal r, because this choice leads to simpler final results.

Comparing (1) and (2), it is evident that

$$\pm h = (\pi d/2R) - (\pi d/n). \tag{3a}$$

By definition, $h>0$. Hence, for retrogressive layering, $2R>n$, and for progressive layering, $2R<n$. Therefore, the expression for h is given as

$$h = \pm \pi d(n-2R)/2Rn. \tag{3b}$$

The progression p in inches per coil revolution is the product of the axial displacement k in inches per throw, and the exact number of throws per coil revolution n'. By definition, $1/n'$ is that fraction of a coil revolution corresponding to the actual circumferential displacement of one throw of wire and is equal to $[(\pi d/n)$

[3] The reader should note that the term "progressive," as applied to the circumferential displacement h, must be distinguished from the name employed to describe the type of winding.

[4] The term "retrogressive layering" refers to the convention of building up the helical ridges in an opposite direction from those of progressive layering. This distinction may be noted by comparing Figs. 1 and 2.

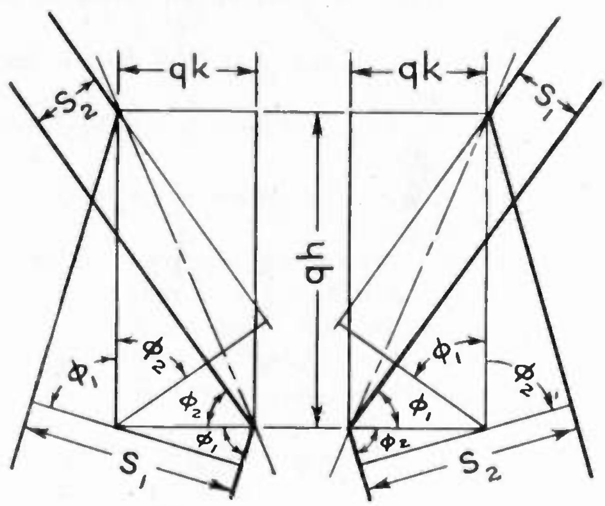
$\pm h]/\pi d$. But, from (3a), the numerator of this expression is $\pi d/2R$, so that $n' = 2R$ and

$$k = p/n' = p/2R. \tag{4}$$

From Fig. 3, which shows the geometry of the spacing of adjacent turns of wire in greater detail, the following equations are obtained:

$$s_1 = qh \cos \phi_1 \mp qk \sin \phi_1 \tag{5a}$$

$$s_2 = qh \cos \phi_2 \pm qk \sin \phi_2. \tag{5b}$$



(a) (b)

Fig. 3—Detailed view of the spacing of adjacent turns of wire. (a) Retrogressive layering. (b) Progressive layering.

In designing a progressive universal coil it is desirable to specify only the smaller of the two wire spacings, because this dimension obviously can not be less than the diameter of the wire. Accordingly, for both progressive and retrogressive layering, only the negative sign is of practical significance in (5a) and (5b). The subscript may then be eliminated from the symbol s.

If the values of ϕ, h, and k, as determined from (1), (3), and (4), respectively, are substituted into (5), the result is[5]

$$4(1 - b^2)R^2 - 4n[1 \mp e(1 - b^2)]R + n^2[1 - a^2 \mp 2e + (1 - b^2)e^2] = 0, \tag{6}$$

and, similarly,

$$n^2[1 - a^2 \mp 2e + (1 - b^2)e^2]r^2 - 4n[1 \mp e(1 - b^2)]r + 4(1 - b^2) = 0, \tag{7}$$

where $r = 1/R$, and

$$a = s/qc \tag{8}$$

$$b = ns/q\pi d = s/\pi dv \tag{9}$$

$$e = p/nc. \tag{10}$$

The solution to the quadratic equation (6) is

$$R = (n/2)\left\{\frac{1 \mp e(1-b^2) \mp \sqrt{1-(1-a^2)(1-b^2)}}{(1-b^2)}\right\}. \tag{11}$$

Contrary to the usual custom, it is necessary to invert the sign before the radical in the above quadratic formula in order to adhere to the convention that the upper sign shall always be used for progressive layering. Since $R < n/2$ for progressive layering, the upper sign before the radical should not be positive. The solution to (7) is

$$r = (2/n)\left\{\frac{1 \mp e(1-b^2) \pm \sqrt{1-(1-a^2)(1-b^2)}}{1-a^2 \mp 2e+(1-b^2)e^2}\right\}, \tag{12}$$

and the sign before the radical is opposite to that of (11) because $r = 1/R$. This is the same as (15) given by Simon,[2] except for the inversion of the sign of the term $e(1-b^2)$. This inversion appears to be a misprint in his paper.[6]

Equation (11) is exact but complicated. However, it is readily apparent that (12) is even more complicated than (11), and yet both statements yield exactly the same information concerning the gear ratio. For practical computation, (11) may be simplified with very little error by setting $b^2 = 0$ since, ordinarily, $b^2 \ll 1$. With this approximation, (11) reduces to a simple linear equation convenient for rapid slide-rule calculation:

$$R = (n/2)[1 \mp (e + a)]. \tag{13}$$

On the other hand, the approximation to (12), as given by Simon, is[7]

$$r = (2/n)[1 \pm e \pm \sqrt{a^2 + b^2} + a^2 - e^2]. \tag{14}$$

The accuracy to be expected from (13) and (14), when these equations are used under the most unfavorable conditions likely to be met in practice, is of considerable interest. Let certain extreme values of a, b and e be chosen such that a condition of maximum error in the approximate equations is obtained for a coil which may still be considered practical. Then, for such values, it may be shown that (13) is in error by less than 2 per cent, whereas (14) is in error by almost 38 per cent.[8] If the same numerical values of a and b are retained, the error in (14) decreases as e diminishes until it finally becomes about 5 per cent. However, for exactly the same values of the parameters, (13) shows an error of slightly more than 1 per cent.

Thus (13), which is simple enough to be linear and free of radicals and squared terms, also is even more precise than the cumbersome (14). To obtain the best approximation to (12), it is only necessary to set $b^2 = 0$. If this is done, the result is the reciprocal of (13), namely, $r = 2/n[1 \mp (e+a)]$, which is, of course, just as accurate as (13).

[5] The intermediate mathematical steps in the derivation are given in Appendix I.

[6] Appendix II contains a list of typographical errata found in Simon's article.

[7] Equation (14) above was taken from footnote reference 2 and is equation (16) of that paper.

[8] Refer to Appendix III for the details of these calculations.

This approximation is a natural one because, in effect, it makes the gear ratio independent of the changing diameter of the coil, a condition which is physically apparent, especially in an ordinary universal coil. That a self-supporting coil of this kind can be produced at all is due entirely to the fact that the diameter has such a negligible effect upon the gear ratio.

The accurate and approximate equations defining the gear ratio for ordinary universal coils are, respectively,

$$R = (n/2)\frac{[1 \mp \sqrt{1-(1-a^2)(1-b^2)}]}{(1-b^2)} \tag{15}$$

and

$$R = (n/2)(1 \mp a), \tag{16}$$

which are obtained from (11) and (13) simply by setting $e=0$, because there is no progression p.

The terms n and e must be specified in the design equations if they are to be of practical use. Hitherto, n, the number of throws per coil revolution, has been determined by empirical formulas only. It will now be demonstrated that a theoretical expression for the maximum value of n can be derived for the ordinary universal winding. However, since the progression is usually quite small compared to the cam throw, the expression for n also may be applied to the progressive universal winding.

IV. Number of Throws per Turn

The following symbols are important in this section:

ϕ = winding angle for the ordinary universal coil, corresponding to n throws per turn

μ = coefficient of static friction between the surface of the coil form and the insulation of the wire

$\gamma = \mu d/c$.

In the ordinary universal coil, $\tan\phi = \pi d/nc$, or $n = \pi d/c \tan\phi$. Since n is constant during the winding process, it can be seen that, as the diameter increases, so does the winding angle. This natural increase in coil diameter dictates a value of n which corresponds to the minimum initial winding angle, in order that the winding may acquire its greatest height.

Fig. 4 is a sketch of a portion of the winding surface of the coil, showing only two complete throws of wire. There are three forces acting on this section of wire: the two equal tensile forces T at the ends of the wire, and f, the resultant of all the retaining frictional forces along the wire. Each tensile force acts in a plane which is tangent to the cylinder surface at the end of the wire. The two planes intersect in a line parallel to the cylinder axis and directly above the bend point of the wire. The lines of action of the tensile forces have been extended backward in the tangent planes until they intersect at their common point of application O, where each tensile force is resolved into three mutually perpendicular components. For clarity, the components of only one of the tensile forces are shown in Fig. 4.

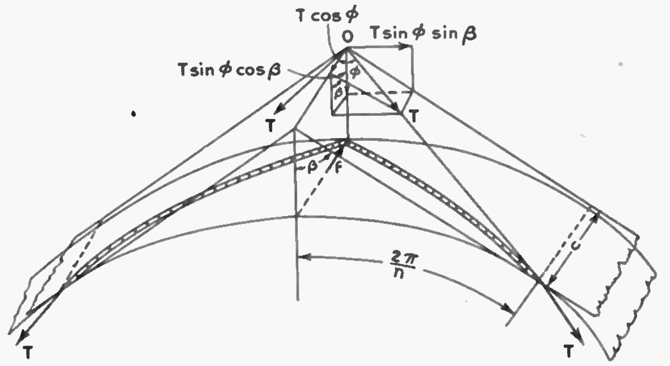


Fig. 4—Spatial view of a portion of an ordinary universal coil, showing all the mechanical forces acting on two throws of wire.

The vector representing the force T is the diagonal of a small rectangular parallelepiped. The winding angle ϕ is the angle in the tangent plane between T and its axial component, $T\cos\phi$. The angle β is the angle between the tangent plane and a plane passing through the axis of the cylinder and the bend point of the wire. Hence, β is the complement of the angle subtended at the axis by the projection of one throw of wire on the edge of the coil. This subtended angle is equal to $2\pi/n$ radians. The component of T normal to the surface of the coil is $T\sin\phi\cos\beta$, and the component of T in a direction tangent to the circumference of the coil is $T\sin\phi\sin\beta$, as shown in Fig. 4. Because there are two tensile forces acting at the point of application, the component $T\sin\phi\sin\beta$ is balanced by an equal and opposite force. Likewise, the sum of the normal components is $2T\sin\phi\cos\beta$, and the sum of the axial components is $2T\cos\phi$.

When the wire is on the verge of slipping toward the center of the coil surface, the resultant frictional force f must be just large enough to equal the sum of the axial components of the two tensile forces. Therefore,

$$f = 2T\cos\phi = \mu(2T\sin\phi\cos\beta) \tag{17}$$

where μ is the coefficient of friction. Since $\cos\beta = \sin(2\pi/n)$ and $\tan\phi = \pi d/nc$, (17) becomes

$$(2\pi/n)\sin(2\pi/n) = 2c/\mu d = 2/\gamma. \tag{18}$$

This equation can not be solved explicitly for n in closed form. It is correct for all values of $n>4$, or for $\gamma>4/\pi$, because then the central angle subtended at the axis of the coil does not exceed $\pi/2$. When $n=4$, the intersecting planes of Fig. 4 become parallel to each other, and the total normal force exerted at the bend of the wire is simply $2T\sin\phi$. When the central angle exceeds $\pi/2$, the analysis is continued by supposing that each throw of wire on either side of the bend in Fig. 4 is cut at the point that is one-fourth of the circumference from the bend. The tensile forces at these points in the wire are then parallel to each other, as in the case where $n=4$. Hence, the total normal force exerted at the bend of the wire is $2T \sin \phi$ and $f=2T \cos \phi=\mu(2T \sin \phi)$, from which

$$n = \mu\pi d/c = \pi\gamma. \qquad (19)$$

This equation states that n is a linear function of γ for all $\gamma \leqq 4/\pi$.

If (18) is solved for γ and then differentiated with respect to n, it is found that the slope of the curve at $n=4$ is the same as the slope of the straight line of (19). This indicates that (18) and (19) together give n as a continuous function of γ for all values of γ, as shown in the curve of Fig. 5. Hence, if γ is specified, the maxijum value of n may easily be determined from this chart. For the ordinary universal coil the maximum value of n is also the optimum value, if the winding is to be as high as possible. It is interesting to note that the preceding derivation shows that n, and consequently R, is independent of the magnitude of the winding tension. Representative values of the coefficient of friction for various materials have been determined and assembled in Table I.

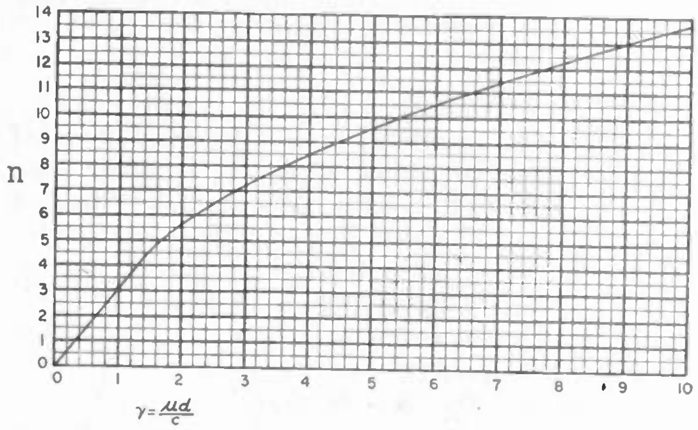


Fig. 5—A plot of the curve which determines the maximum number of throws per coil revolution as a function of the coefficient of friction, coil-form diameter, and cam throw.

V. Determination of Progression

In the sample calculation given by Simon,[2] the progression p is selected arbitrarily for use in the design equation. However, the specification of e, and consequently p, is not entirely a matter of free choice, but depends upon the desired geometry of the winding. Specifically, it will now be shown that the progression is determined by the ratio of the width of a helical ridge w_l to the mean distance between adjacent ridges s_r, as shown in Fig. 2. Let this ratio be denoted by the symbol σ. The angle between the direction of the helical ridges of the coil and the axis of the coil is called ψ, and the length of wire on the forward stroke is called L.

By definition,

$$\sigma = w_l/s_r, \qquad (20)$$

and, from Fig. 2,

$$w_l = c \sin \psi - (\pi d/n) \cos \psi. \qquad (21)$$

The distance s_r+w_l is $L \sin \alpha$ where $L=(c+k)/\cos \phi_1$ and $\alpha=\pi-\psi-\phi_1$. Therefore,

$$s_r + w_l = (c+k)(\sin \psi \cos \phi_1 + \cos \psi \sin \phi_1)/\cos \phi_1 \qquad (22)$$

or

$$s_r + w_l = (c+k)(\sin \psi + \tan \phi_1 \cos \psi). \qquad (23)$$

But from (1), $\tan \phi_1 = \pi d/2R(c+k)$, so that

$$s_r + w_l = c \sin \psi + k \sin \psi + (\pi d/2R) \cos \psi. \qquad (24)$$

From Fig. 1 or Fig. 3,

$$\tan \psi = h/k \qquad (25)$$

and, since $\sin \psi = \tan \psi \cos \psi$, (24) becomes

$$s_r + w_l = c \sin \psi + [(\pi d/2R) + h] \cos \psi. \qquad (26)$$

By substituting for h the expression given by (3), this simplifies to

$$s_r + w_l = c \sin \psi + (\pi d/n) \cos \psi. \qquad (27)$$

Therefore, (21) and (27) yield, as a final result,[9]

$$s_r = (2\pi d/n) \cos \psi. \qquad (28)$$

This equation can be deduced directly from Fig. 2, but the proof of its correctness is not as conclusive as this derivation. Although (28) above is derived for the retrogressive layering of Fig. 2, exactly the same result may be obtained for the progressive layering shown in Fig. 1.

Now it is possible to indicate just how e (and also p) is related to the ratio σ. From geometrical considerations, this ratio is bounded according to the following relationship:

$$0 < \sigma \leqq 1. \qquad (29)$$

When the ratio is less than unity, any two adjacent ridges of the coil are separated by an opening or depression similar to that of a space-wound solenoid. When the ratio equals unity, this opening vanishes and the ridges become contiguous, just like two turns of a close-wound solenoid. Dividing (21) by (28),

$$\sigma = (nc/2\pi d) \tan \psi - 1/2 \qquad (30)$$

or

$$\tan \psi = (2\pi d/nc)(\sigma + 1/2). \qquad (31)$$

From (10), $e=p/nc$; from (4), $p=2Rk$; and from (25), $k=h/\tan \psi$; so that

$$e = 2Rh/nc \tan \psi. \qquad (32)$$

[9] Equation (28) differs by a factor of $2/n$ from Simon's equation (24) of footnote reference 2. His result was obtained for the specific instance where $n=2$ and therefore is not suitable as a general formula. Consequently, his equations (26) and (27) are applicable only when $n=2$.

Substituting for h and $\tan\psi$ the expressions given by (3) and (31), respectively, (32) becomes

$$e = \pm (n - 2R)/2n(\sigma + 1/2). \qquad (33)$$

Now, if (11) and (33) are solved simultaneously for R and e by elementary algebraic manipulation, the results are:

$$R = (n/2)\left\{\frac{(2\sigma+1)[1 \mp \sqrt{1-(1-a^2)(1-b^2)}] - (1-b^2)}{2\sigma(1-b^2)}\right\} \qquad (34)$$

$$e = \frac{\sqrt{1-(1-a^2)(1-b^2)} \pm (1-b^2) \mp 1}{2\sigma(1-b^2)}. \qquad (35)$$

In (35), if it is recalled that $b^2 \ll 1$, then, approximately,

$$e = a/2\sigma. \qquad (36)$$

This last expression can be obtained more easily by employing the approximate (13), rather than (11), in the above simultaneous solution.

From (36), it is evident that once the geometric pattern of the coil is fixed by a suitable choice of the ratio σ, the value of the progression is uniquely determined. By setting $b^2 = 0$ in (34), or by substituting e, as given by (36), into (13), the following design formula (in terms of σ rather than e) is obtained:

$$R = (n/2)[1 \mp a(1 + 1/2\sigma)]. \qquad (37)$$

VI. Design Procedure

Coil-design procedure, as relating to the proper selection of gears for winding universal coils, can be outlined in the following steps:

Progressive Universal Coils

1. Determine the diameter of the wire, including the insulation. This dimension may then be designated as the wire spacing s, if a tight winding is desired. Practically, however, it is often necessary to maintain some separation between adjacent turns to allow for mechanical defects in the machine, and for variations in thickness of insulation and flexibility of the wire. It has been found that, for generally satisfactory results, the spacing s should exceed the wire diameter by about 25 per cent.

2. Select the appropriate coefficient of friction μ from Table I and compute $\gamma = \mu d/c$ where d is the coil-form diameter and c is the cam throw. The proper number of throws per revolution n is then obtained directly from Fig. 4. It should be selected as the nearest integer, or if this is not convenient, as the nearest integer plus a simple fraction. Such a choice of n avoids a complex winding pattern, a condition which may result in physically defective coils. An odd integer yields the simplest pattern. For progressive universal coils, it is recommended that n be selected as the nearest integer below the curve. Express n as a fraction, q/v, where q is an even integer of the least possible magnitude. (For example, if n is between 4 and 5, it may be chosen as $4\frac{1}{2}$ or 9/2. Then $q = 18$ and $v = 4$.)

3. Select a numerical value for σ, the ratio of the width of a helical ridge to the mean distance between ridges. Because this quantity influences the distributed capacitance, inductance, and physical dimensions of the coil, which cannot be predetermined, the designer is left with a free choice. A close-wound coil requires that $\sigma = 1$, causing the capacitance to be large. When $\sigma = 1/2$, the space between ridges becomes equal to the width of a ridge. This condition is approximately average for many coils. Having selected σ, calculate the rate of progression from the formula $p = s/2v\sigma$, obtained from (8), (10), and (36). Then determine the gears required to produce this amount of progression in inches per coil revolution.

4. Compute the gear ratio R from the equation

$$R = (n/2)[1 \mp a(1 + 1/2\sigma)]$$

where $a = s/qc$. The choice of plus or minus sign is optional, as either one yields satisfactory results.

5. Set either index of the C scale to the gear ratio on the D scale of the slide rule and move the indicator along until two gear numbers are found that coincide beneath the hairline. The drive gear number is located on the D scale and the cam gear number is on the C scale. If the machine is built with idler gears having a ratio other than unity, the computed value of R first must be multiplied by the idler gear ratio, and the result set on the slide rule in the manner just described.

Ordinary Universal Coils

The procedure in designing ordinary universal windings is the same as that for progressive universal windings, except that Part 3 is omitted and the formula for the gear ratio is changed to read as follows:

$$R = (n/2)(1 \mp s/qc).$$

APPENDIX I

Details of Gear-Ratio Derivation

Because only the smaller of the two wire spacings is of practical importance, (5) can be reduced to

$$s/q \cos\phi = h - k \tan\phi \qquad (38)$$

where s and ϕ represent s_1 and ϕ_1 for progressive layering, and s_2 and ϕ_2 for retrogressive layering. From (1) and (4),

$$\tan\phi = \pi d/(2Rc \pm p) \qquad (39)$$

$$\cos\phi = 1/\sqrt{1 + [\pi d/(2Rc \pm p)]^2}. \qquad (40)$$

If the expressions for h, k, $\tan\phi$, and $\cos\phi$, as given by (3), (4), (39), and (40), respectively, are substituted into (38), the result is

$$(ns/q\pi d)\sqrt{(2Rc \pm p)^2 + (\pi d)^2} = \pm c(n - 2R) - p. \quad (41)$$

After squaring and rearranging terms, this leads directly to (6).

APPENDIX II

Errata in Simon's Paper[2]

Page 868, second line of Table of Symbols: c should read e.

Page 869, line just above the figure: Fig. 3 should read Fig. 4.

Page 869, numerator of equation (15): first $\pm$ sign should read $\mp$.

Page 870, line below equation (20): (6) should read (5).

Page 870, equation (25) and preceding line: w_e should read w_t.

Page 871, equation (34): $\pm$ sign should precede the left-hand member.

APPENDIX III

Error Calculations

In order to estimate the maximum error which is apt to occur when the approximate equation (13) is used in preference to the exact equation (11), certain limiting values must be assigned to the terms a and b. The radical in (11) may be rearranged into the form

$$\mp a\sqrt{1 - b^2 + (b/a)^2}$$

from which it can be seen that the approximation in this expression is obtained by assuming the term under the radical sign to be unity. The error, however, is great when b is large and a is small. Now $a = s/qc = (ns/\pi dq)(\pi d/nc) = b(\pi d/nc)$, but $\pi d/nc = \tan\phi$ where ϕ is the angle shown in Fig. 4. From (18), $nc/\pi d = \cot\phi = \mu \sin(2\pi/n) \leq \mu$. Hence, by assuming that the coefficient of friction has a maximum value of about 0.40 (Table I), the term a cannot be less than $2.5b$.

Universal coils for radio purposes are seldom made with wire larger than No. 22, which has a diameter of about 0.025 inch. In addition, let it be assumed that the wire spacing s is limited to twice the wire diameter, and that the coil form diameter is usually not less than $\frac{1}{4}$ inch. Since the minimum magnitude of v is unity, the term b therefore has a maximum value of

$$b = s/\pi dv = 0.05/0.25\pi = 0.0636.$$

Hence, the term a has a minimum value of

$$a = 2.5b = 0.159.$$

In practice, if the opening between helical ridges exceeds about three times the width of a ridge, the coil generally is not acceptable because the winding becomes unecomical of useful space in which to attain the desired inductance. Such an extreme condition corresponds to $\sigma = 1/4$ and

$$e = a/2\sigma = 0.318.$$

Let $a = 0.159$, $b = 0.0636$, and $e = 0.318$ be substituted into (11), (13), and (14), considering only progressive layering at this time. The results are:

For (11),

$$2R/n = (1 - 0.3167 - 0.171)/0.996 = 0.514.$$

For (13),

$$2R/n = 1 - (0.318 + 0.159) = 0.523.$$

For (14),

$$2R/n = 2/nr = 1/(1 + 0.318 + 0.171 + 0.0253 - 0.101) = 1/1.4133 = 0.708.$$

Therefore, (13) is in error by

$$100(0.514 - 0.523)/0.514 = -1.75 \text{ per cent},$$

and (14) is in error by

$$100(0.514 - 0.708)/0.514 = -37.8 \text{ per cent}.$$

As the progression decreases, the errors become less, until finally, when $\sigma = 1$ and the coil is closely wound, (13) has an error of -1.13 per cent, while (14) yields an error of -4.65 per cent, these figures being computed in exactly the same manner as those above. For retrogressive layering, the results are so similar to those for progressive layering that they need not also be stated here.

TABLE I

Values of the Coefficient of Friction μ

Coil-Form Material	Insulation	
	Silk	Cotton
Cardboard (clean, dry)	0.20	0.21
Mica Ceramics (glass-bonded) (*Mycalex, Mykroy*, etc.)	0.24	0.26
Phenol Formaldehyde (molded) (*Bakelite, Resinox*, etc.)	0.18	0.21
Phenolic Laminates (*Formica, Phenolite*, etc.)		
Cloth Base	0.20	0.25
Paper Base	0.15	0.16
Polystyrene	0.16	0.17
Porcelain		
Glazed	0.12	0.14
Unglazed	0.36	0.39
Wood	0.21	0.22

A Vacuum-Tube-Type Transducer for Use in the Reproduction of Lateral Phonograph Recordings*

JAMES F. GORDON†, MEMBER, I.R.E.

Summary—A method is described wherein the lateral mechanical vibrations from a phonograph record are used to move a vacuum-tube element which creates variations in the anode current comparable to the anode-current variations caused by a change in grid voltage in the regular triode-type vacuum tube. An experimental-type movable-grid tube is shown, with the performance data. An applicable circuit is shown and other uses of the tube are mentioned.

Introduction

THE MOVABLE-ELEMENT vacuum tube is not new.[1] Phonograph pickups using a mechanically driven vacuum-tube element have not been previously considered, however, because of constructional difficulties and cost factors. The technical problems presented by relatively high noise level, microphonics, insufficient output voltage, fragility, and mechanical transmission difficulties have been overcome in experimental models.

It is possible to make a linear phonograph pickup from many low-mass generating devices, provided that the amplitude of movement is held to a low value. Unfortunately, this often results in too low an output from many possible devices for practical application.

For large vacuum-tube element movements, a considerable departure from linearity is usually experienced. It is desirable to use a structural design providing an output comparable to that obtained from other accepted phonograph pickups, while at the same time keeping the element motion small.

The Experimental Vacuum Tube

The sketch in Fig. 1 shows a parallel-plane triode tube with the grid structure communicating directly with the stylus. Assuming a zero or negative grid poten-

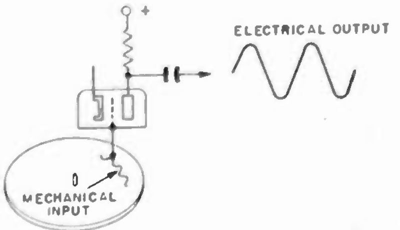


Fig. 1—Diagram of a simple triode phonotube arrangement.

tial, the anode current will normally decrease as the grid approaches the cathode. A positive grid will cause an increase in anode current as the grid approaches the cathode.

Because of the increased output obtainable, the tube here described is operated with a positive grid bias.

* Decimal classification: R339×621.375×621.385.971. Original manuscript received by the Institute, November 27, 1946; revised manuscript received, April 10, 1947.

† Bendix Radio Corporation, Baltimore 4, Md.

[1] U. S. Patent Nos. 1,871,253, G. F. C. Bauer; 2,290,531, G. F. Brett; 2,157,719, F. L. Pulaski; 1,936,922, T. W. Sukumlyn; RE 15,540, L. DeForest.

Assuming such a device to be linear, a sine displacement of the stylus due to a sine recorded groove deviation results in a sine increment of anode current. To accomplish this, a number of difficulties must be overcome. They are as follows:

(1) Anode-cathode spacing must be small enough to allow low anode-supply voltage; yet spacing must allow necessary clearance for grid motion.

(2) The grid structure must have low mass, but nevertheless must exhibit effective control over the electron stream.

(3) For normal stylus movement the grid should not operate too closely to the cathode, since nonlinearity increases. (See Fig. 10(a)).

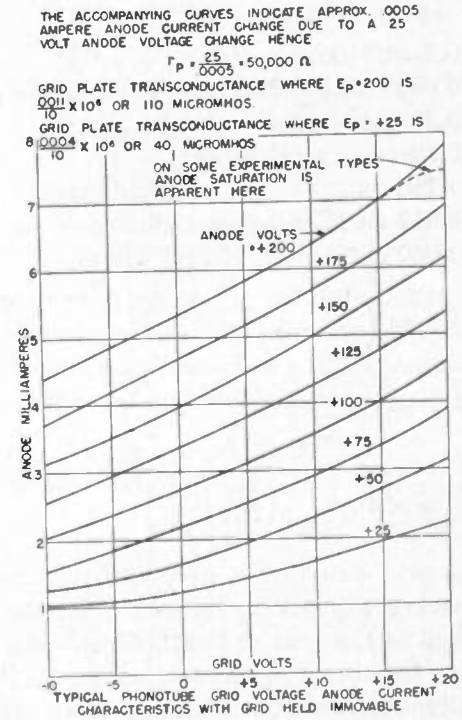


Fig. 2—By maintaining the grid in a fixed position at its neutral point, the static grid-voltage versus anode-current characteristics may be taken in a normal manner.

(4) A means of entering the envelope must be devised which makes a tight seal and yet offers no appreciable mechanical impedance to the motion of the stylus shaft.

(5) Sufficient vertical compliance (ease of motion) must be incorporated.

(6) Undesirable resonances must not occur within the usable audio spectrum.

(7) The vertical sensitivity should be low.

(8) The stylus must have a low lateral mechanical impedance.

(9) The vacuum tube must have a size comparable to accepted phonograph pickup cartridges.

(10) The entire structure should be of a design which can be produced by conventional manufacturing methods.

For early experimental tubes an arbitrary grid-cathode spacing was chosen to be 0.060 centimeter. Using an anode-cathode spacing of 0.160 centimeter, fairly linear operation could be obtained.

Curves of the linearity of operation to be expected are shown in Fig. 2. These indicate that for small excursions the harmonic distortion should be low.

An exact theoretical design approach toward the ultimate triode structure becomes complicated, especially where the sizes of elements involved are small with respect to the associated wires and supports. It is possible to use a simple form of Child's law, as applied to parallel-plane triode structures, to determine what may be expected with respect to linearity of operation, amplification factor, distortion, etc.

By means of the following expressions, determination of relative anode currents for a grid excursion of 0.020 inch, or plus or minus 0.025 centimeter, is as indicated in Fig. 3, where

$A_b = 0.2$ cm.^{2} (anode area)
$d_c = 0.035$ to 0.085 cm. (grid-cathode distance)
$e_b = 0.25$ volt (anode voltage)
$e_c = 0.20$ volt (grid voltage)
$d_b = 0.160$ cm. (anode-cathode distance)
$r_g = 0.012$ cm. (grid-wire radius)
$P = 0.050$ cm. (pitch of grid wires)

$$\mu = \frac{2.7 d_c \left(\dfrac{d_b}{d_c} - 1\right)}{P \log \dfrac{P}{2\pi r_g}}$$

$$I_P = 2.3 \times 10^{-6} \frac{A_b}{d_c^2} \left(\frac{e_b + \mu e_c}{1 + \mu}\right)^{3/2}.$$

Since the grid structure is pivoted from one end, the parallel spacing will not be uniform; i.e., the grid and anode as well as the grid and cathode spacing will vary as the space between two sides of a hinge.

For small movements of the grid, a mean point on the grid structure may be taken for measuring the excursion. If the tip of the grid structure is used as a measurement of the excursion, the calculated results would be as the solid line in Fig. 3, whereas the more exact conditions are shown by the dashed line.

Early tubes used approximately 0.12 cm.^{2} cathode area and 0.2 cm.^{2} anode area.

Wire mesh was first used as a grid structure, but was later abandoned in favor of the structure shown in Figs. 4 and 5.

A 0.002-inch-thick Kovar diaphragm was used, through which the stylus shaft passed. The diaphragm was 0.5 inch in diameter and was a satisfactory means of providing a flexible seal. A rib pressed at right angles to the grid shaft motion prevented undue motion of the stylus in a direction parallel to the record groove.

The air pressure acting on the diaphragm caused it to be externally concave. This automatically positioned the grid structure.

If the stylus were attached directly to the end of the

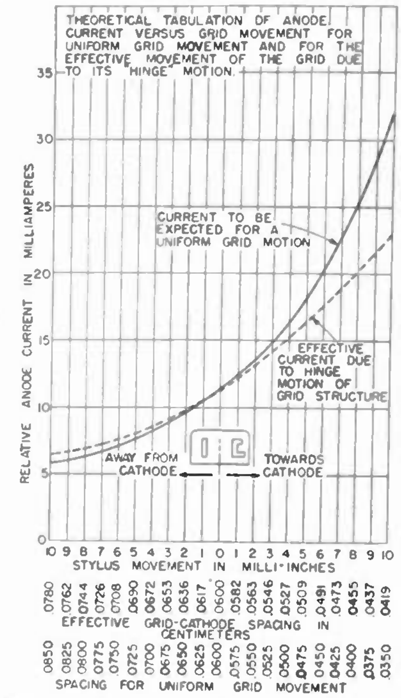
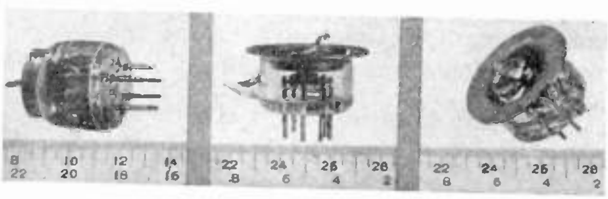


Fig. 3—The theoretical computation approximates practice for changes in anode current due to grid position. The variations from these computations where the grid draws current and the anode voltage is reduced are responsible for the increasing linearity of operation, as indicated by the curves of Fig. 10(a), (b), and (c).

grid shaft, there would be little vertical compliance. In order to keep distortion and "needle talk" as low as possible, it is desirable to have a certain amount of



(a) (b) (c)

Fig. 4—(a) Early phonotube model. (b) and (c) Two views of the experimental model discussed in the text.

vertical compliance.[2,3] This was accomplished by using a thin metal strap, as shown in Figs. 4 and 5, attached to the end of the stylus shaft.

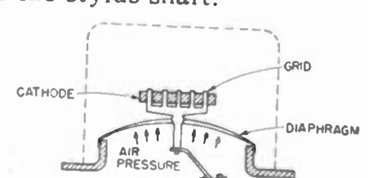


Fig. 5— Because of the forked structure and the vertically stiff diaphragm, plus the vertically compliant stylus strap, very little vertical motion actually reaches the grid structure.

It is generally recognized that a lateral pickup should have negligible vertical sensitivity. Since the signal to be transmitted to the pickup is entirely vested in the

[2] H. A. Frederick, Vertical Sound Records: Recent Fundamental Advances in Mechanical Records on "Wax," *Jour. Soc. Mot. Pic. Eng.*, vol. 18, pp. 141–164; February, 1932.

[3] J. A. Pierce and F. V. Hunt, "On distortion in sound reproduction from phonograph records," *Jour. Acous. Soc. Amer.*, vol. 10, July, 1938.

lateral displacement of the recorded groove, any motion of the stylus in a vertical direction which creates an output from the pickup element must be considered as

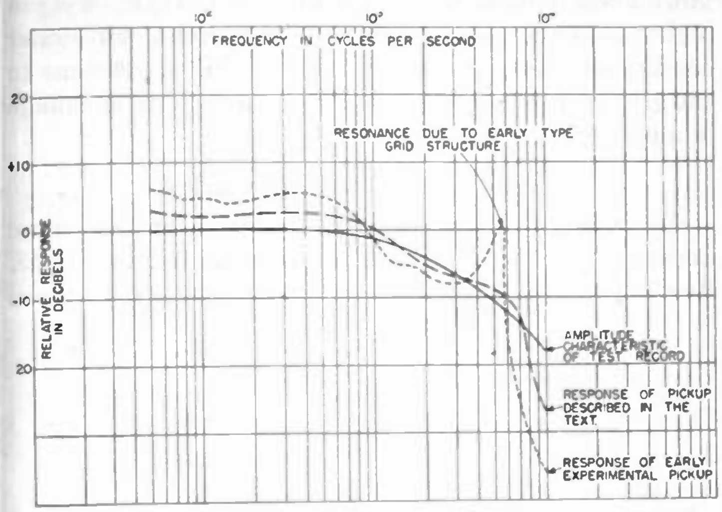


Fig. 6—Frequency response of an early type of pickup, along with that of the tube described in the text, compared with the amplitude characteristic of the test record used.

unwanted disturbance. This usually makes itself apparent as random noise, rumble, rattle, and harmonic distortion. This unwanted disturbance or increment will be present in the output circuits of the device if the vertical sensitivity is appreciable. In the experimental pickup, vertical sensitivity was reduced by the use of the compliant strap, a forked grid structure (Fig. 5), and by a vertically stiff diaphragm.

The difference between vertical and lateral sensitivity was measured to be in excess of 30 db.

Mechanical resonances occurring in the structure were not large, and so did not present a great damping problem. The greatest difficulty was encountered in preventing the grid structure within the tube from being resonant at a frequency within the usable audio spectrum. On the final models this resonance was approximately 12,000 c.p.s. A small circular piece of damping material with a slight projection extending against the stylus strap was satisfactory (see Fig. 4). Care must be exercised in the application of damping such that the mass is not increased beyond the point where satisfactory upper frequency response is obtained.

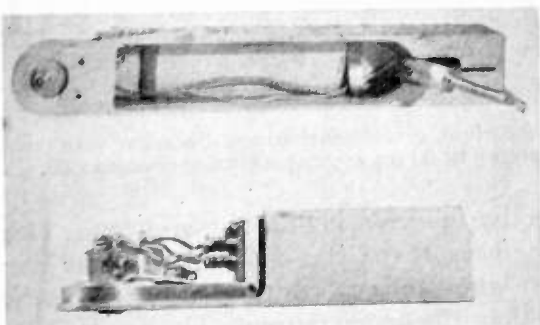


Fig. 7—The pickup arm showing method of mounting the vacuum tube for experimental operation. *Above*—Bottom view of experimental pickup arm. *Below*—Side view of phono pickup arm showing how the vacuum tube is mounted.

An amplitude-sensitivity curve for the final experimental model is shown in Fig. 6. Practically uniform response to nearly zero c.p.s. may be obtained, if desired.

The vacuum-tube pickup shown in Fig. 4(c) is of simple design and may be readily duplicated with conventional vacuum-tube manufacturing equipment. The cathode plate structures are welded directly to the mounting pins, which extend through the envelope as soldering lugs or miniature socket pins. The over-all height of the tube from stylus tip to the top of the base pins is 2.3 centimeters. This is small enough to allow styling in tone arms of normal design. An experimental pickup arm is shown in Fig. 7. The stylus force on the record was adjusted to be 15 grams. Satisfactory performance is obtained with this bearing weight; however, heavier weights may be used, if desired.

Pickup Applications

There are several methods for coupling the energy from the pickup tube. The simplest form is that of the resistance-capacitance coupling shown in Fig. 8. Ex-

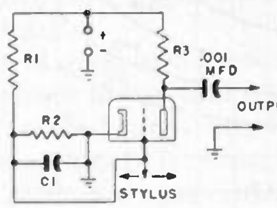


Fig. 8—A simple form of *RC* coupling which proved satisfactory in the experimental work with the tube.

amination of the curves of Fig. 10 shows that a plus or minus grid deviation does not generally give as linear a grid-current change as takes place in the anode circuit. It is desirable that no difference in actual grid potential

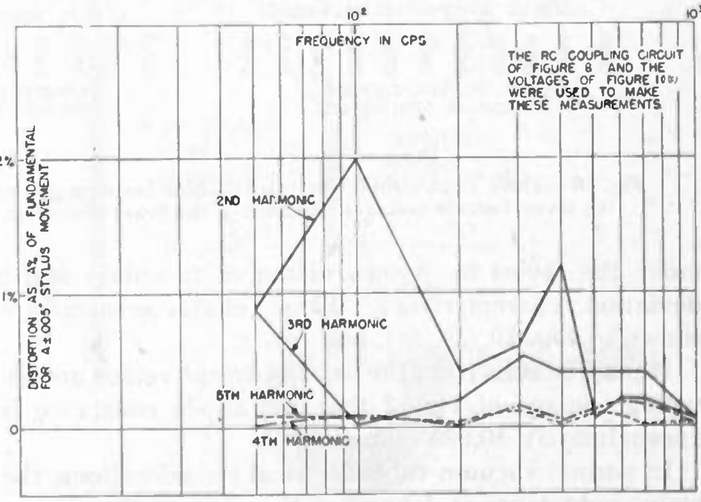


Fig. 9—Curves showing measured harmonic distortion where the stylus was driven by a voice-coil arrangement. In actual practice the stylus deviation does not reach ±0.001 inch with the result that the actual measured distortion is comparable to or less than that on the record itself.

occur between the grid and the cathode due to a variation in grid current, since this would introduce distortion which would be largely second-harmonic in order. The impedance between the grid and the cathode should be kept reasonably low at audio frequencies for this reason.

The R_1 and R_2 combination serves to bias the grid positive. A value of 0.1 megohm for the load resistance R_3 was satisfactory in obtaining voltage peaks of approximately 1 volt from standard shellac recordings. Transformer coupling is entirely satisfactory, and a voltage stepup may be obtained.

The tube may be operated as a radio-frequency oscillator, under which conditions both amplitude and frequency modulation of the output energy may be accomplished.

Static Characteristics of the Vacuum Tube

A test jig was set up using a micrometer screw to move the stylus by known distances to obtain static deviation characteristics. These characteristics are shown in Fig. 10 (a), (b), and (c).

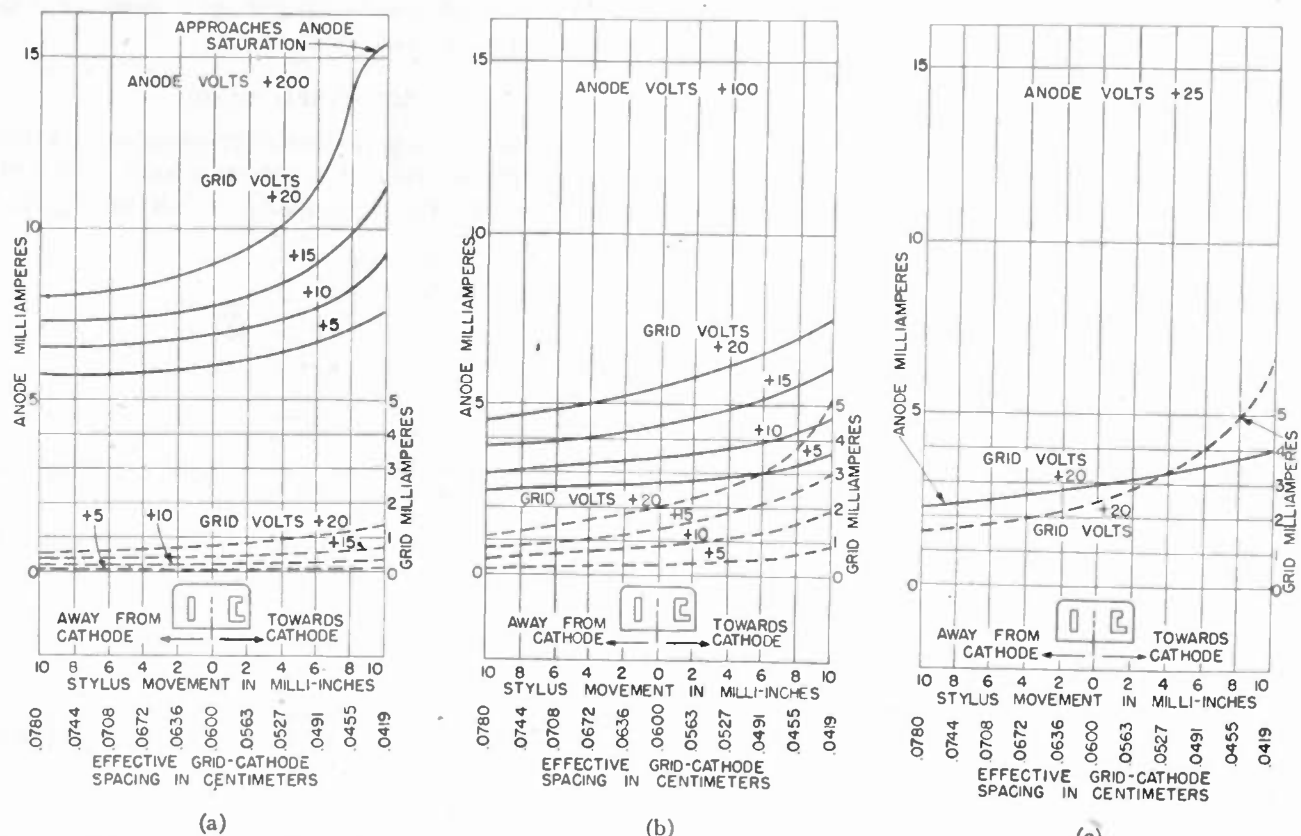


Fig. 10—These curves show the relationships between grid current, anode current, grid deviation, and linearity, with respect to several anode voltages. Because of the lower distortion, the curves shown in (c) are typical operating characteristics.

It may be seen from the anode-current versus anode-voltage curves of Fig. 2 that the anode resistance is approximately 50,000 ohms.

In normal vacuum-tube electrical considerations, the transconductance is defined as the ratio of the change in grid voltage to the resulting change in anode current, with all other conditions fixed. In the case of a vacuum tube where the grid and anode potentials remain fixed and the anode current depends upon the physical movement of the grid circuit, the grid-anode transconductance is electromechanical in nature. It should, therefore, be expressed as the ratio of grid deviation in inches, or some unit of measure, to the accompanying anode-current change.

It would be desirable to differentiate between this type of grid-anode transconductance and the strictly electrical type by expressing the grid-anode electromechanical transconductance in terms of, for example, vibromhos. This would be the ratio of the grid swing in inches or centimeters to the anode current, with other conditions fixed, or the ratio of the grid movement to the grid voltage required to provide an identical change in anode current.

Distortion Characteristics

An examination of Fig. 10 (c) indicates that more linearity is accomplished than is indicated in Fig. 3. As the positive grid approaches the cathode it becomes less effective in increasing the anode current. This is due in large measure to the condition in this particular structure where the grid current may be as great or even greater than the anode current. The result of this condition is to give a straighter anode-current versus grid-deviation curve (see Fig. 10 (c)). This indicates that, for small deviations of the grid structure such as are encountered in practice, the harmonic distortion will be low (see Fig. 9).

Other Uses of the Phonotube

The many applications of the phonotube are too numerous for consideration in this paper. The tube may be used in an oscillatory circuit as a source of frequency modulation where the grid structure is driven mechanically at the modulation frequency. By using the output of a discriminator to position the grid structure of a phonotube inherently connected into a reactance-tube-modulated oscillator, effective automatic frequency control may be obtained.

The tube may be used as an electromechanical mixer in applications where it is desirable to combine an electrical vibration with a mechanical one.

Conclusion

The electron tube which has a movable small-mass grid structure is practical as a phonograph pickup device. It possesses high performance characteristics with respect to fidelity, amplitude of output, and distortion.

The device is considered to be of a size comparable to existing accepted units used as a phonograph pickup.

Acknowledgment

The writer wishes to acknowledge the valuable assistance of Al Stuart and George S. Miles and their staff at the Eclipse-Pioneer Division of Bendix Aviation Corporation in the development of the tube described in this paper.

Field Measurements on Magnetic Recording Heads*

DONALD L. CLARK†, ASSOCIATE, I.R.E., AND LYNN L. MERRILL‡

Summary—A method is described for measuring relative values of the magnetizing force along the path traversed by the recording medium in passing through a magnetic recording or reproducing head. Field-distribution curves obtained by this method are shown. A method for calculating the frequency response of a reproducing head from field-distribution data is presented, and results of calculations are compared with a measured frequency response. In the recording process, the important part of the field lies in the air gap. The high-frequency response depends on the sharpness of cutoff on the "leaving" side of the gap, and is independent of the shape on the "approaching" side.

Introduction

IT IS THE PURPOSE of this paper to describe measurements of the magnetic fields produced by typical heads used for wire recording, and to correlate their performance with these measurements. The technique used in measuring the magnetic field depends upon the measurement of the electromotive force induced in a minute exploring coil as the coil is placed at various positions along the path of the recording medium. Results of measurements using this technique have been published by only one investigator,[1] to the authors' knowledge.

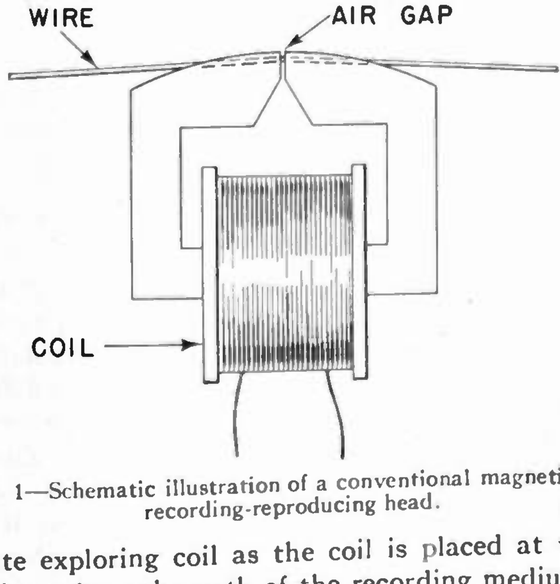


Fig. 1—Schematic illustration of a conventional magnetic-wire recording-reproducing head.

Experimental Technique

A schematic illustration of the type of head upon which most of the measurements were performed is given in Fig. 1. This is a conventional head having a high-permeability core, a close-fitting slot for the wire, and a short air gap.

The mechanical setup used for positioning the exploring coil with respect to the head is illustrated in Fig. 2. It is evident from the figure that the position of the coil is fixed, and that the head is movable with respect to it. The exploring coil is the most critical part of the setup, and several were tried before a reasonably satisfactory coil was obtained. It is desirable to make the coil very small in order to determine the field distribution with as much detail as possible. The most satisfactory type of coil used consisted of a single turn of No. 46 Formex-insulated wire wound around a 0.0012×0.015-inch phosphor bronze strip. The leads from the coil were twisted tightly together for a distance of several inches to minimize the effect of stray fields.

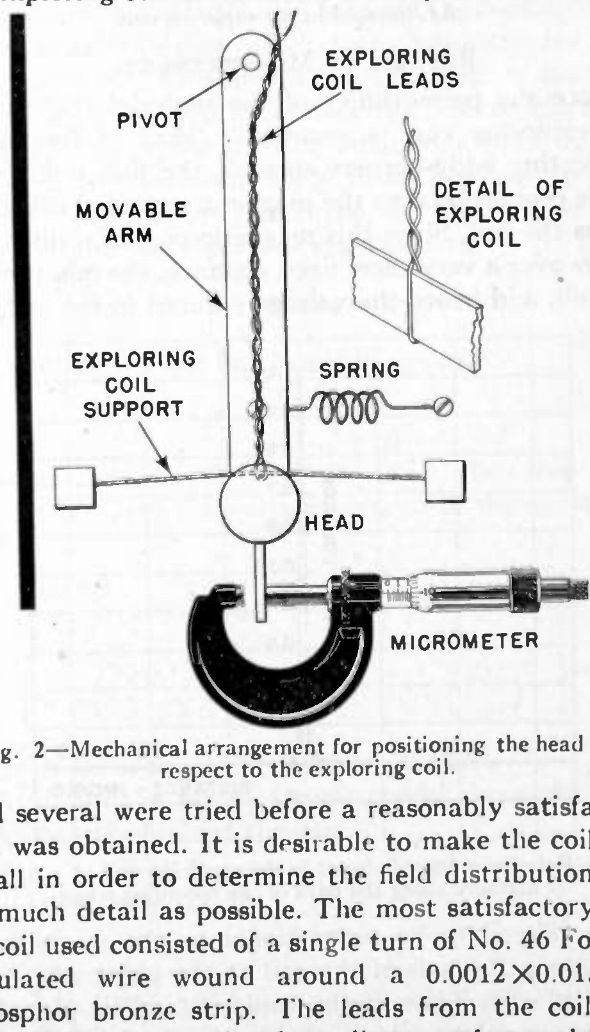


Fig. 2—Mechanical arrangement for positioning the head with respect to the exploring coil.

* Decimal classification: R365.35×681.843. Original manuscript received by the Institute, November 8, 1946; revised manuscript received, February 5, 1947.

† Formerly, Stromberg-Carlson Company, Rochester, N.Y.; now, University of Rochester, Rochester, N. Y.

‡ Stromberg-Carlson Company, Rochester, N. Y.

[1] Von Heinz Lübeck, "Magnetische schallaufzeichnung mit filmer- und ringkopfen," *Akus. Zeit.*; vol. 2, pp. 273–295; November, 1937.

A block diagram of the electrical setup for measuring the e.m.f. induced in the exploring coil is shown in Fig. 3. A current of any desired amplitude and frequency is supplied to the head under test by the audio oscillator. The induced e.m.f. is amplified and measured with a wave analyzer tuned to the oscillator frequency. With this arrangement, voltages as small as 10^{-8} volt were detectable.

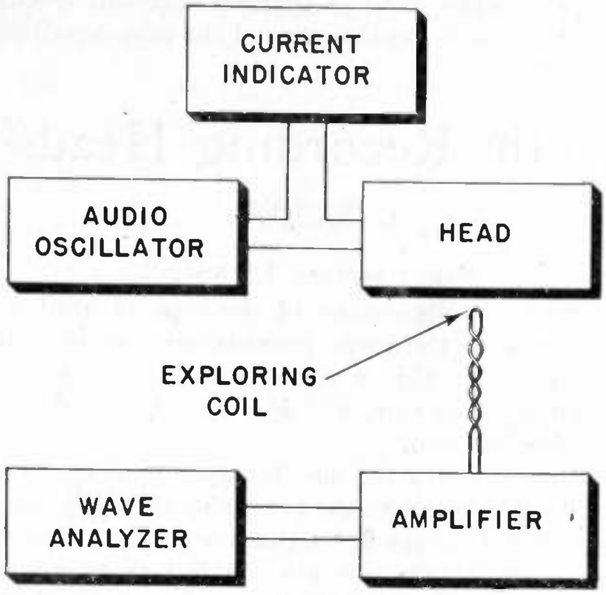


Fig. 3—Block diagram of the electrical setup for measuring e.m.f. induced in the exploring coil.

Results of Measurements

Since the permeability of the material enclosed by the exploring coil is practically that of free space (neglecting eddy-current effects), the flux linking the coil is proportional to the magnetic potential difference across the coil. Since this magnetic potential difference occurs over a very short fixed distance, the flux through the coil, and hence the voltage induced in the coil, can

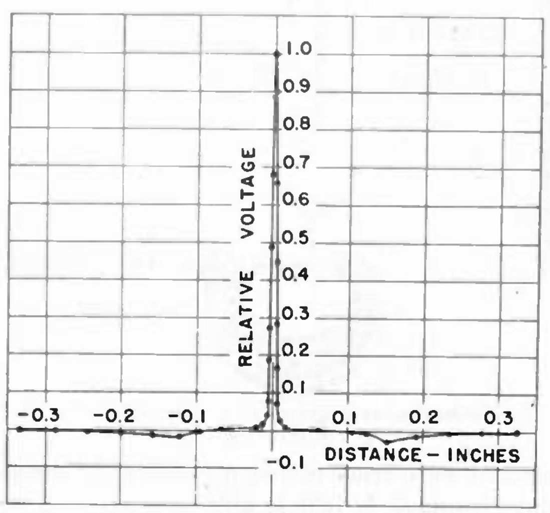


Fig. 4—Relative voltage induced in the exploring coil as a function of distance along the path of the recording wire.

be considered to be proportional to the magnetizing force along the axis of the coil at the point where the coil is placed. Thus, if the exploring coil is placed at various positions along the path of the recording medium, a measure of the relative magnetizing force to which the recording medium is subjected at each point can be obtained. It is assumed that the field distribution is not appreciably affected by the presence of the recording medium in a practical case, because the permeability of the recording medium is very small compared with the permeability of the core.

A typical graph of voltage induced in the exploring coil as a function of distance along the path of the recording wire is shown in Fig. 4. The large peak at the origin is obtained when the coil is in the air gap, and the smaller peaks when the coil is entering and leaving the slot. Measurements on different types of heads indicate that the shape of the field-distribution curve varies considerably with the design of the head. In particular, the shape of the small peaks on either side of the curve shows large variations for different heads, this shape depending upon the configuration of the slot in the regions where the recording wire enters and leaves the head.

The current required to saturate the core of a head can be found by placing an exploring coil in the air gap, and measuring the induced voltage as a function of the exciting current. Correlation of the current required to saturate the magnetic circuit of the head with that required for recording at saturation levels on wires of known coercivity shows that saturation of the head is not appreciable for wires having coercivities less than 600 oersteds.

Magnetizing force as a function of frequency can be measured in a similar manner. What variation is found is due to resonance of the inductance of the coil with its own distributed capacitance. The variation is not necessarily detrimental, and for conventional heads is small in the audio-frequency range.

No important variation in the shape of the field-distribution curve has been found when the amplitude or frequency of the exciting current is varied, or when bias current is present.

Theoretical Determination of Frequency Response of a Reproducing Head

It is desirable in gaining an understanding of the operation of a reproducing head, and in designing heads for improved performance, to be able to calculate the frequency response from a measurement of the field distribution. A method for accomplishing this follows.

The field-distribution curve of Fig. 4 was obtained by exciting the coil of the head with a known current and measuring the voltage in the exploring coil. By the reciprocity theorem it would be possible to interchange the current and the voltage; that is, if the same current were used to excite the exploring coil, the same voltage would be read across the coil of the head. Under these conditions the exploring coil can be considered to be a source of magnetic potential difference acting over the effective length of the coil. If the effective length is sufficiently small, the exploring coil can be considered to be the source of a certain magnetic potential difference per unit length; that is, a magnetizing force, applied at a given point. Knowing the voltage induced in the coil of the head with this applied magnetizing force, and knowing the frequency and the number of turns in the coil, it is possible to calculate the flux in the lower leg of the core. By this reasoning, the curve of Fig. 4 represents relative values of flux in the lower leg of the core of the head resulting from a certain magnetizing force applied at any point along the path of the recording wire.

Noting that the curve of Fig. 4 exhibits symmetry about the center line of the gap and approximate local symmetry about the small peaks on either side of the gap, the curve can be represented by the following empirical expressions:

$$\begin{aligned}
f(x) &= Ae^{-\alpha_1 x} && \text{for } 0 \leqq x \leqq a;\\
f(x) &= -Be^{-\alpha_2(c-x)} && \text{for } b \leqq x \leqq c;\\
f(x) &= -Be^{-\alpha_3(x-c)} && \text{for } c \leqq x \leqq d;\\
f(x) &= Ae^{+\alpha_1 x} && \text{for } -a \leqq x \leqq 0\\
f(x) &= -Be^{-\alpha_2(c+x)} && \text{for } -c \leqq x \leqq -b\\
f(x) &= -Be^{-\alpha_3(-x-c)} && \text{for } -d \leqq x \leqq -c
\end{aligned} \qquad (1)$$

where

$f(x)$ = the magnetizing force at the distance x from the center of the gap
A = the magnetizing force at the central peak
B = the magnetizing force at the lateral peaks
a = the abscissa to the point where $e^{-\alpha_1 x}$ becomes negligible
b = the abscissa to the point where $e^{-\alpha_2(-x+c)}$ becomes negligible
c = the abscissa to the center line of the small peak
d = the abscissa to the point where $e^{-\alpha_3(x-c)}$ becomes negligible.

Assume a sinusoidally magnetized wire to be drawn through the slot of the reproducing head at an arbitrary constant velocity, the wavelength and amplitude of the sinusoidal distribution being arbitrary. The effect of each element of wire in sending flux through the lower leg of the core can be found by multiplying the magnetizing force produced by the element of wire by the ordinate to the field-distribution curve at the point where the element is situated. Applying the principle of superposition, the total effect of the wire in sending flux through the lower leg of the core can be found by integrating the effects of all the elements whose influence is appreciable.

Neglecting demagnetization, the applied magnetizing force may be taken as

$$m = M \sin \frac{\pi}{L}(x - vt) \qquad (2)$$

where

m = instantaneous magnetizing force (ampere turns/inch)
M = maximum magnetizing force (ampere turns/inch)
L = one-half wavelength $= v/2f$ (inch)
v = velocity of the wire (inch/second)
x = distance from center line of gap (inch)
t = time (second)
f = frequency.

The flux through the core of the head is then

$$\phi = M \int_{-d}^{+d} f(x) \sin \frac{\pi}{L}(x - vt)dx, \qquad (3)$$

or, integrating and combining terms,

$$\phi = 2M \sin \frac{\pi}{L} vt \left[B \frac{-\frac{\pi}{L}(\alpha_2^2 - \alpha_3^2) \sin \frac{\pi c}{L} + (\alpha_2 + \alpha_3)\left\{\left(\frac{\pi}{L}\right)^2 + \alpha_2\alpha_3\right\} \cos \frac{\pi c}{L}}{\left\{\alpha_2^2 + \left(\frac{\pi}{L}\right)^2\right\}\left\{\alpha_3^2 + \left(\frac{\pi}{L}\right)^2\right\}} - A \frac{\alpha_1}{\alpha_1^2 + \left(\frac{\pi}{L}\right)^2}\right]. \qquad (4)$$

To establish a relationship between the coefficients A and B, consider a unit pole to be moved along the path of the recording wire from a point far outside the head on one side to a point far outside the head on the other side and back to the starting point by a path well removed from the head. Since there has been no net change in the magnetic potential, the net area between the curve $f(x)$ and the x-axis must be zero. Thus

$$\int_{-d}^{+d} f(x)dx = 0. \qquad (5)$$

Substituting for $f(x)$ and integrating yield

$$B = \frac{A}{\alpha_1} \frac{\alpha_2\alpha_3}{\alpha_2 + \alpha_3}. \qquad (6)$$

Making use of (4) and (6) and the fact that $e = Nd\phi/dt(10)^{-8}$, the voltage induced in the coil of the head is

$$e = 4\pi(10^{-8})NMAf \cos 2\pi ft \left[\frac{\alpha_2\alpha_3}{\alpha_1} \frac{\frac{-2\pi f}{v}(\alpha_2 - \alpha_3) \sin \frac{2\pi fc}{v} + \left\{\left(\frac{2\pi f}{v}\right)^2 + \alpha_2\alpha_3\right\} \cos \frac{2\pi fc}{v}}{\left\{\alpha_2^2 + \left(\frac{2\pi f}{v}\right)^2\right\}\left\{\alpha_3^2 + \left(\frac{2\pi f}{v}\right)^2\right\}} - \frac{\alpha_1}{\alpha_1^2 + \left(\frac{2\pi f}{v}\right)^2}\right] \qquad (7)$$

where N is the number of turns in the coil.

If $f(x)$ may be assumed symmetrical about the small peaks on either side of the gap

$$\alpha_3 = \alpha_2$$

and (7) becomes

$$e = 4\pi(10^{-8}) \frac{NMAf \cos 2\pi ft}{\alpha_1} \left[\frac{\alpha_2^2 \cos \frac{2\pi fc}{v}}{\alpha_2^2 + \left(\frac{2\pi f}{v}\right)^2} - \frac{\alpha_1^2}{\alpha_1^2 + \left(\frac{2\pi f}{v}\right)^2}\right]. \qquad (8)$$

Frequency-response curves have been calculated using (7) and (8) with values of the constants determined from the data plotted in Fig. 4. There was good agreement between the curves obtained from the two equations. Since (8) is the simpler of the two and is adequate to illustrate the principles involved, it will be used in the discussion that follows.

The values of the constants used in (8) are as follows:

$$\alpha_1 = 375$$
$$\alpha_2 = 17$$
$$c = 0.14 \text{ inch}$$
$$v = 24 \text{ inches per second.}$$

The empirical field-distribution curve obtained when these constants are substituted in (1), and the measured field-distribution curve, are shown in Fig. 5 plotted to expanded scales. The frequency-response curve calculated with (8) is shown in Fig. 6, together with a measured frequency-response curve for the same head.

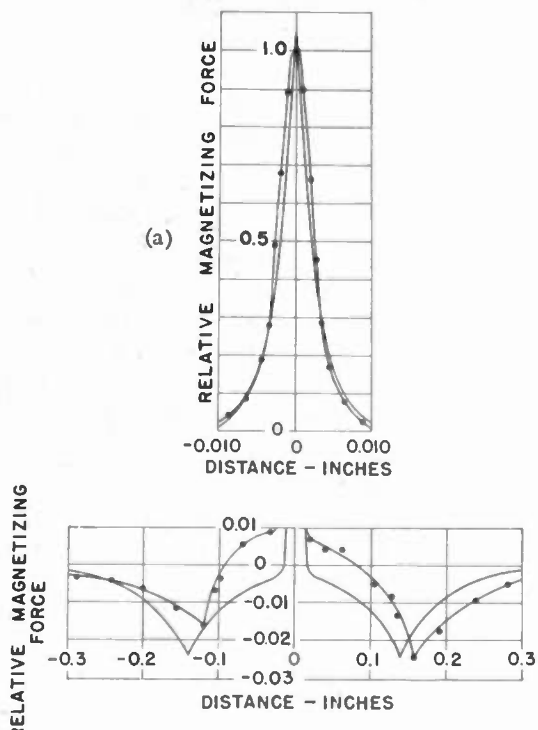


Fig. 5—Measured and empirical field-distribution curves plotted with expanded scales to show comparison. Dots are measured points. (a) Expanded horizontally. (b) Expanded vertically.

The characteristics of the frequency-response curve can be accounted for by considering (8). At very low frequencies each term in the brackets is nearly equal to unity, and the difference between them is very small. Thus the voltage induced in the coil of the head is very small for frequencies corresponding to wavelengths which are large compared with the dimensions of the head.

As the frequency increases, the first term in the brackets oscillates and decreases in amplitude. The second term remains practically constant throughout the low-frequency range. Thus, in the frequency region where the wavelength is of the same order of magnitude as the dimensions of the head, there are undulations in the frequency response which gradually become imperceptible with increasing frequency. On the average, the response will increase about in proportion to frequency in the low-frequency range. Because of the undulations, it will rise somewhat faster in certain regions, particularly at frequencies lower than that at which the first maximum occurs.

For frequencies corresponding to wavelengths which are small compared with the dimensions of the head, the first term in the brackets of (8) will be negligible. The response is then due entirely to the discontinuity at the gap. As the frequency increases and the wavelength becomes very small, the measured response reaches a maximum and then falls off rapidly. The calculated response behaves quite similarly.

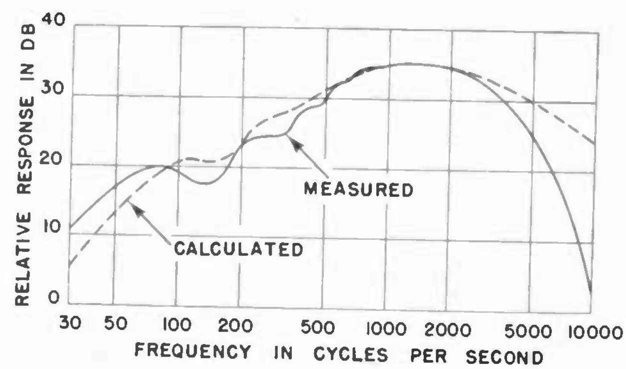


Fig. 6—Frequency-response curve calculated from field-distribution data compared with measured frequency-response curve.

The lack of agreement between the calculated and measured response in the high-frequency region is due to three factors: first, the neglect of self-demagnetization; second, the finite dimensions of the exploring coil which made it impossible to determine the field-distribution curve in sufficiently fine detail for accurate calculations; and third, the inaccuracy in the empirical representation of the data in the respects that determine the performance of the head near its high-frequency limit.

Effect of Field Distribution on Performance of a Recording Head

The reproducing process differs from the recording process in that in the former all elements operate in an essentially linear fashion, while in the latter the operation of the recording medium is decidedly nonlinear. This greatly complicates analysis of the recording process. Without detailed analysis, however, it is possible to show qualitative correlation betwen the field distribution and the performance of a recording head.

When a signal is recorded on a magnetic medium, the process consists essentially of subjecting each element of the medium to a peak value of magnetizing force so related to the signal that the flux remaining in the element is proportional to the instantaneous value of the signal. The proper relation between the signal and the peak magnetizing force is established by the use of a biasing field superimposed upon the signal field.[2] In order that the flux remaining in the element may depend only upon the peak magnetizing force, it is essential that the element be subjected to no subsequent reversals of magnetizing force which are comparable in magnitude with its coercive force.

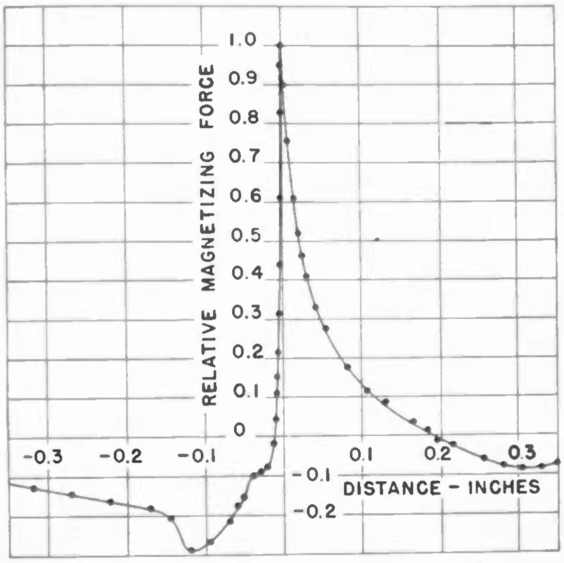


Fig. 7—Relative magnetizing force as a function of distance along the path of the recording medium for an experimental recording head.

Measurements on typical recording media show that the coercive force is roughly one-half the maximum applied magnetizing force when the medium is working below saturation. Thus, after an element of the medium has been subjected to a certain magnetizing force, subsequent applications of magnetizing forces having perhaps one-fourth this value would have little effect on the flux remaining in the element. According to this reasoning, then, the portion of the field of a recording head which is important in the recording process is that part in which the magnetizing force is greater than about 20 or 30 per cent of the maximum value of the magnetizing force.

In order that an element of the recording medium shall not be subjected to a reversal of magnetizing force when recording signals of high frequency, it is necessary to remove the element from the influence of the recording field in a time equal to or less than approximately half the period of the highest frequency to be recorded. In practice this is accomplished by having the field of the recording head decrease as rapidly as possible in the direction in which the medium leaves the head, and by moving the medium with sufficiently high velocity. When these conditions are not met, the medium is subjected to one or more reversals of magnetizing force, partial erasing takes place, and the high-frequency response of the system is impaired.

The manner in which the magnetizing force applied to an element builds up to its peak value is relatively unimportant. The reason for this is that the flux remaining in each element of the medium is, for practical purposes, determined by the maximum magnetizing force to which it has been subjected. This means that the shape of the field of a recording head is relatively unimportant in the region where the field applied to an element of the medium is increasing to its maximum value as the element moves through the head.

[2] L. C. Holmes and D. L. Clark, "Supersonic bias for magnetic recording," *Electronics*, vol. 18, pp. 126–136; July, 1945.

Fig. 7 shows the field distribution of an experimental recording head. Fig. 8 shows frequency-response curves measured using this head for recording and a conventional head for reproducing. The upper curve represents the response obtained when the recording wire was traversing the head from right to left, and the lower curve from left to right, as referred to Fig. 7.

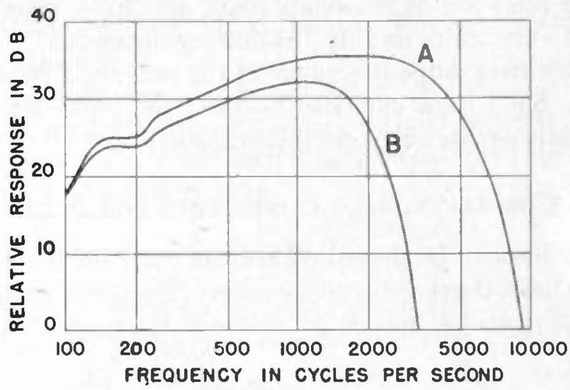


Fig. 8—Comparison of frequency-response curves for (A) an abrupt change in the field, and (B) a gradual change in the field as the medium leaves the recording head.

It is evident that, as the wire traverses the head from right to left, the recording field builds up slowly to a maximum and then drops abruptly. The shape of the frequency-response curve measured under these conditions is indistinguishable from that measured with a conventional recording head, with the remainder of the elements of the system unchanged. Thus the slow build up of the recording field and the large irregularity in the decay of the field have no appreciable effect on the recording process.

On the other hand, when the recording wire traverses the head from left to right, as referred to Fig. 7, the recording field builds up rapidly and drops slowly. The lower curve of Fig. 8 represents the frequency response measured under these conditions. It is evident that somewhat more than an octave has been lost in the high-frequency region. This loss is attributed to the erasing action which takes place when the medium is subjected to several reversals of decreasing magnetizing force while within the influence of the recording field.

Acknowledgment

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Video Delay Lines*

J. P. BLEWETT†, ASSOCIATE, I.R.E., AND J. H. RUBEL‡, ASSOCIATE, I.R.E.

Summary—**Continuous coaxial transmission lines are described in which the velocity of propagation is about one one-thousandth of the velocity of light. These lines include a solenoidal inner conductor and a Litz-braid outer conductor. Phase and amplitude distortions in such lines are discussed, and design procedures are presented to yield lines of optimum performance under various conditions.**

I. Introduction

THIS PAPER describes delay transmission lines for delaying video signals by periods from a fraction of a microsecond up to 2 or 3 microseconds. These lines are of reasonable size and have properties which are not unduly frequency-dependent. Their characteristic impedances are of the order of a thousand ohms. Such lines can also be used for pulse forming, impedance matching,[1] or electrical filtering.[2]

II. Construction of Continuous Delay Lines

The lines to be described are the result of increasing to the limit the number of sections in the usual synthetic line of cascaded tee or pi network elements. The in-

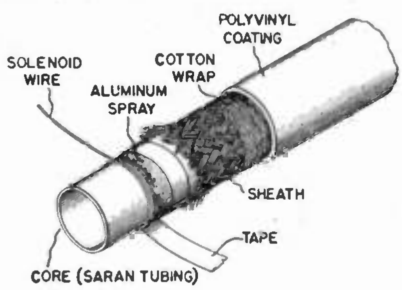
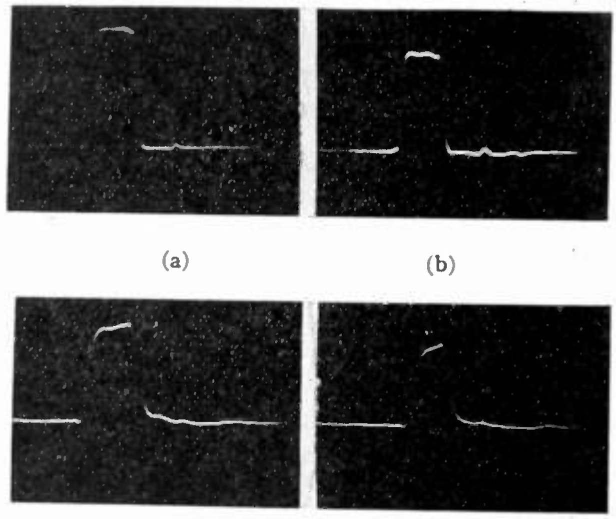


Fig. 1—Details of construction of continuous delay line.

ductance elements merge into a continuous coil and the capacitive elements are replaced by the distributed capacitance between the turns of this coil and an outer shield. Physically, this structure takes the form shown in Fig. 1. The solenoid is wound on a flexible insulating core of polyvinylidene chloride ("Saran") about $\frac{3}{16}$ inch in diameter. The coil is close-wound of 3- or 4-mil Formex-insulated copper wire. A layer of insulating tape serves as the dielectric between the conductors of the line, and the outer conductor is a braid of insulated wires connected together at one end of the line. A cotton covering and an outer shell of polyvinyl tubing complete the line. In use, a ground connection is made at one end to the outer conductor. Input and output leads are at opposite ends of the internal coil. It has been found, for reasons which will appear in Section IV, that a layer of aluminum paint over the inner coil has the effect of decreasing phase distortion in these lines.

If the outer conductor of the line also takes the form of a coil, wound in the opposite direction to the inner coil, the inductance per unit length will be increased by a factor approaching four. It will be shown below, however, that the higher delay per unit length achieved in this way may result in serious phase distortion, so



(a) (b)

(c) (d)

Fig. 2—Effect of delay on a 1-microsecond pulse. (a) 1-microsecond pulse generated by discharge of a double-layer delay line. (b) 1-microsecond pulse delayed 1 microsecond in a single-layer line. (c) 1-microsecond pulse delayed 3 microseconds. (d) 1-microsecond pulse delayed 5 microseconds.

this feature of the double-layer line is not necessarily an advantage. Since, moreover, it is impossible to ground both input and output of the double-layer line, this line has not found much application and has not been carried into production. In the laboratory the double-layer lines have been used chiefly as pulse-forming lines, since in this application they seem to be somewhat better than single-layer lines. When discharged through an 884 tube into its characteristic impedance, the double-layer line yields square pulses such as that shown in the oscillograph trace of Fig. 2(a).

III. Performance of Continuous Delay Lines

In a typical single-layer line, the mechanical characteristics were as follows:

Core diameter: 3/16 inch

* Decimal classification: R117.2. Original manuscript received by the Institute, December 13, 1946; revised manuscript received, May 26, 1947.

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[1] H. E. Kallmann, "High-impedance cable," Proc. I.R.E., vol. 34, pp. 348–351; June, 1946.

[2] H. E. Kallmann, "Transversal filters," Proc. I.R.E., vol. 28, pp. 302–310; July, 1940.

Over-all wire diameter: 3.6 mils (No. 40 A.W.G., Formex-covered)
Number of turns per inch: 277
Tape covering: $3/8\times 0.0015$ inch aceto-butyrate tape, single wrap, with 50 per cent overlap
Litz-braid outer conductor: 192 strands of No. 36 A.W.G. Formex-covered wire braided in 8-strand strips at a pitch of 1.9 inches.

The electrical characteristics of such a line are:
Inductance per inch length: 51 μh.
Capacitance per inch length: 42 $\mu\mu$fd.
Characteristic impedance: 1100 ohms
Delay per foot length: 0.55 microsecond
Attenuation: 1.3 db per microsecond delay.

As will be demonstrated in the next section, delay lines are increasingly subject at high frequencies to phase and amplitude distortion. Around 3 or 4 Mc., the wavelength along the line just described becomes short enough that a turn of wire begins to find itself magnetically coupled with turns which are appreciably out of phase with it. This results in an apparent decrease in inductance and increase in propagation velocity with frequency, and composite signals will undergo "phase distortion." Amplitude distortion is due to skin effects and dielectric losses, both of which become important at high frequencies. Amplitude distortion may mask phase distortion to some extent by attenuating the high-frequency components which have suffered phase distortion. Both amplitude and phase distortion can be varied within wide limits by changes in design parameters (see Section IV (E)).

Fig. 2 shows oscilloscope traces of a 1-microsecond pulse delayed 0, 1, 3, and 5 microseconds in a line of the design discussed above.

IV. Theory of Delay Lines

A. Line Inductance, Capacitance, Delay and Characteristic Impedance

The low-frequency inductance per unit length L_0 and capacitance per unit length C_0 of a delay line follow from standard formulas. For the single-layer line with close spacing,

$$L_0 = 10^{-9}\pi^2 n^2 D^2 \text{ henry/cm.} \tag{1}$$

$$C_0 = AkD/s \text{ farad/cm.} \tag{2}$$

where D = line diameter in centimeters
n = number of turns per centimeter
s = separation between inner and outer conductor in centimeters
k = effective dielectric constant of material separating inner and outer conductor
A = geometrical factor of the order of 5×10^{-13}.

If we neglect losses, variation of inductance with frequency, and distributed series-capacitance effects, the delay per unit length is given by the transmission-line relation

$$T_0 = \sqrt{L_0 C_0} = 10^{-4}\,(AkD^3n^2/s)^{1/2} \text{ sec./cm.} \tag{3}$$

and the characteristic impedance by

$$Z_0 = \sqrt{L_0/C_0} = 10^{-4}\left(\frac{Dsn^2}{Ak}\right)^{1/2} \text{ ohms.} \tag{4}$$

B. Variation of Inductance with Frequency

In Section III above it was noted that part of the inductance of a turn of the delay line winding may derive from coupling with turns which are not in phase. At higher frequencies, when the wavelength along the line becomes short, this out-of-phase coupling may result in a material decrease in the effective inductance per unit length of the winding. The actual amount of this decrease will now be determined. We are indebted to H. Poritsky and Mrs. M. H. Blewett for the treatment which follows.

For the purposes of this computation, it is assumed that current flows only in the azimuthal direction in a thin sheet of infinite conductivity. It is assumed further that all fields include a factor

$$e^{j\omega(t-z/v)}$$

where z measures distance along the z axis of a cylindrical co-ordinate system coaxial with the delay line, and where v, the velocity of propagation along the line, is much less than c, the velocity of light in free space.

We now set up Maxwell's equations in cylindrical co-ordinates and solve them subject to the above assumptions. As usually happens in problems of this sort with cylindrical symmetry, the various field components prove either to be zero or to be expressible in Bessel functions. The arbitrary coefficients which appear in the general solution are evaluated from the boundary conditions, which are as follows:

1. All components are finite along the axis.
2. All components vanish at infinity.
3. The axial component of magnetic field has a discontinuity at the conducting surface, proportional to the current flowing in the surface.
4. All other components are continuous through the conducting surface.

The final result when these procedures are completed will now be tabulated. The nonvanishing field components prove to be the axial and radial magnetic fields and the azimuthal electric field. Inside the solenoid these components are as follows:

$$\left.\begin{aligned} H_z &= -\frac{4\pi\omega anI}{v} K_1(\omega a/v)I_0(\omega r/v)e^{j\omega(t-z/v)} \\ H_r &= -\frac{4\pi j\omega anI}{v} K_1(\omega a/v)I_1(\omega r/v)e^{j\omega(t-z/v)} \\ E_\phi &= 4\pi j\omega anI \times 10^{-7} K_1(\omega a/v)I_1(\omega r/v)e^{j\omega(t-z/v)} \end{aligned}\right\}. \tag{5}$$

Outside the solenoid:

$$\left.\begin{aligned} H_z &= \frac{4\pi\omega a n I}{v} I_1(\omega a/v)K_0(\omega r/v)e^{j\omega(t-z/v)} \\ H_r &= -\frac{4\pi j\omega a n I}{v} I_1(\omega a/v)K_1(\omega r/v)e^{j\omega(t-z/v)} \\ E_\phi &= 4\pi j\omega a n I \times 10^{-7} I_1(\omega a/v)K_1(\omega r/v)e^{j\omega(t-z/v)} \end{aligned}\right\} \quad (6)$$

where

a = the radius of the solenoid in meters
n = the number of turns per meter of the solenoid
I = the current in amperes in a turn of the solenoid
I_0, I_1, K_0 and K_1 = the modified Bessel functions of the first and second kinds, respectively.[3]

The above solution is in m.k.s. units.

We are now in a position to evaluate the impedance of the coil, since the voltage drop per turn is $2\pi a E_\phi(r=a)$. The impedance is

$$X = \frac{2\pi a n E_\phi}{I}(r = a) \text{ ohms/meter}$$

$$= j\omega 8\pi^2 a^2 n^2 \times 10^{-9} I_1(\omega a/v)K_1(\omega a/v) \text{ ohms/cm.} \quad (7)$$

This is the reactance produced by an inductance of

$$L = 8\pi^2 a^2 n^2 \times 10^{-9} I_1(\omega a/v)K_1(\omega a/v)$$

$$= 2\pi^2 D^2 n^2 \times 10^{-9} I_1(\pi D/\lambda)K_1(\pi D/\lambda) \text{ henries/cm.} \quad (8)$$

where D = line diameters in meters, and λ = wavelength measured along the line in centimeters.

At low frequencies, i.e., for small values of $\pi D/\lambda$, the quantity $I_1(\pi D/\lambda)\, K_1(\pi D/\lambda)$ is approximately $\frac{1}{2}$ and

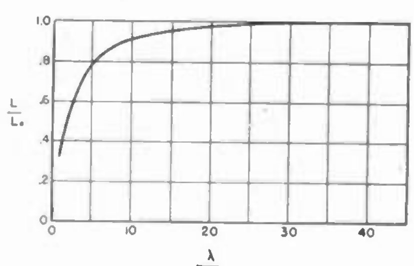

Fig. 3—Variation of inductance with frequency.
$\frac{\text{inductance}}{\text{low-frequency inductance}}$ vs. $\frac{\lambda}{D}$
λ = wavelength along the line
D = diameter of line
L = inductance per unit length
L_0 = inductance per unit length at low frequency.

(8) becomes identical with (1). At higher frequencies the value given for L by (8) falls below the low-frequency value. For $\lambda/D = 16$, L has decreased by 5 per cent; for $\lambda/D = 4$, the value of L is only about half of its low-frequency value. The complete relation between L/L_0 and λ/D is plotted in Fig. 3. The decrease in inductance for the line described in Section III, at 4 Mc., will be about 2 per cent.

[3] A. Gray, G. B. Matthews, and T. M. MacRobert, "Bessel Functions," Macmillan Co., New York, N. Y., 1922. In the deduction of the above results it is necessary to invoke the theorem which states that

$$I_0(x)\, K_1(x) + K_0(x)\, I_1(x) = 1/x.$$

From (5) and (6) it is possible also to show that the fields outside the line are weak compared with those inside, over the low-frequency range. This makes it possible to coil up delay lines with turns in close proximity without noticeable effects on the propagation.

C. Phase Distortion

Phase distortion, or variation of propagation velocity with frequency, arises in delay lines, for the most part, in three ways. In the video range by far the most important cause is the variation of inductance with frequency, discussed in the preceding section. At higher frequencies, distributed-series-capacitance phase distortions become appreciable, and at low frequencies (of the order of 100 kc. and below) phase distortions appear due to the resistive component of the line impedance.

The effects of the variation of inductance with frequency will be considered first. The fractional delay error due to this effect will be

$$\frac{\delta T}{T_0} = \frac{\sqrt{L_0C_0} - \sqrt{LC_0}}{\sqrt{L_0C_0}} = 1 - \sqrt{2I_1(\pi D/\lambda)K_1(\pi D/\lambda)}. \quad (9)$$

But $1/\lambda$ = delay per cm. (T) times the signal frequency (f), so that

$$D/\lambda = fTD = fT_0D(T/T_0)$$

$$= fT_0D\sqrt{2I_1(\pi D/\lambda)K_1(\pi D/\lambda)}. \quad (10)$$

By graphical or numerical methods D/λ can now be eliminated between (9) and (10), so that $\delta T/T_0$ can be plotted as a function of fT_0D. Such a plot is given in Fig. 4. The delay error predicted by this curve agrees within 2 or 3 per cent with that determined experimentally for a number of lines with a variety of design parameters.

Phase distortions due to distributed capacitance result in delay errors having the opposite sign to those caused by the variation of inductance with frequency. If series capacitance C_D per unit length is included in the computation of the effective series element, it is evident that L_0 in (3) must be replaced by

$$\frac{L_0}{1 - \omega^2 L_0 C_D}.$$

The net effect will be an increase in delay with frequency up to the point at which the line becomes self-resonant. For the lines discussed here the resonant frequency will be of the order of 100 Mc., and series-capacitance-effects will be negligible in the video range. C_D may be increased by artificial means, however, to the point at which its effects help to compensate for the effects due to the variation of inductance with frequency. The aluminum-spray coating mentioned in Section II serves this purpose for some types of production delay lines. More refined compensation techniques are described by Kallmann.[4]

[4] H. E. Kallmann, "Equalized delay lines," Proc. I.R.E., vol. 34, pp. 646–657; September, 1946.

The inclusion of a resistance R per unit length in the line element will change (3) to

$$T = \sqrt{L_0C_0}\left(1 + \frac{R^2}{4\omega^2 L_0^2}\right) \text{sec./cm., approximately.} \quad (11)$$

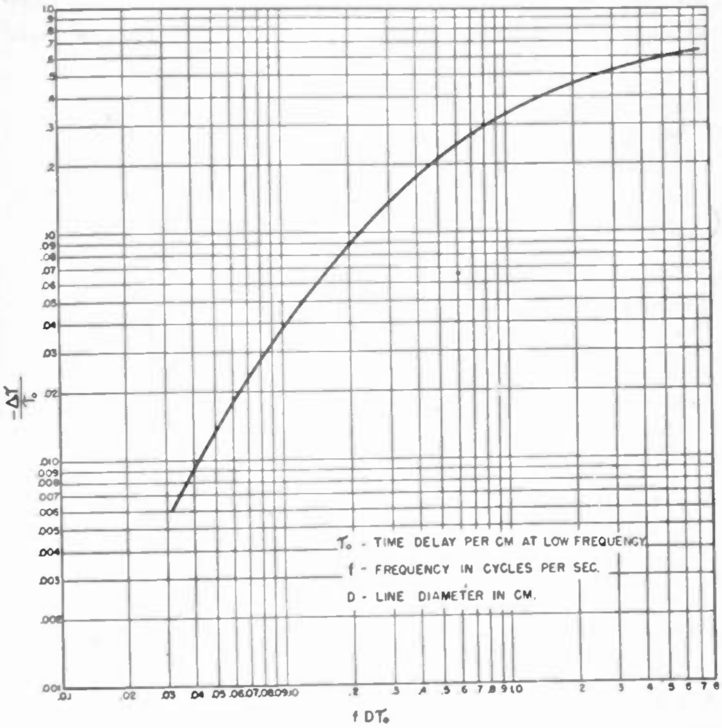

Fig. 4—Delay error as a function of frequency, low-frequency delay, and line diameter.

The correction term is negligible for lines of the type described above, provided the frequency is above 100 kc. or so.

D. *Amplitude Distortion*

If the attenuation constant of a delay line is frequency-dependent, amplitude distortion will result. If the attenuation is not too high, it can be expressed by the usual transmission-line formula:

$$\alpha = \tfrac{1}{2}\sqrt{LC}\,(R/L + G/C) \text{ nepers per unit length.} \quad (12)$$

Since $\sqrt{LC}$ is delay per unit length, we can also express attenuation by:

$$\begin{aligned} A &= \tfrac{1}{2}(R/L + G/C) \text{ nepers/sec. delay} \\ &= 4.343(R/L + G/C) \text{ db/sec. delay.} \end{aligned} \quad (13)$$

It is evident from (13) that amplitude distortion will follow from variations with frequency of resistance R (skin effects), inductance L, or shunt conductance G (dielectric losses).

An approximate expression for the high-frequency resistance can be obtained by assuming that the line winding is replaced by a thin sheet whose thickness is equal to the wire diameter d, and which carries circumferential current. A calculation similar to that presented by Ramo and Whinnery[5] gives the ratio of high-frequency resistance R to d.c. resistance R_0:

$$\frac{R}{R_0} = \frac{d}{\delta}\left(\frac{\sinh(2d/\delta) + \sin(2d/\delta)}{\cosh(2d/\delta) - \cos(2d/\delta)}\right) \quad (14)$$

where δ, the "depth of penetration," is given for copper by

$$\delta = 6.60/\sqrt{f} \text{ cm.} \quad (15)$$

For d greater than 3 mils and f higher than 700 kc., this expression does not deviate by more than 10 per cent from the relation:

$$R/R_0 = d/\delta = 0.152 d\sqrt{f}. \quad (16)$$

But for round copper wire,

$$R_0 = 5.87 \times 10^{-6}\, nD/d^2 \text{ ohms/cm. of line.} \quad (17)$$

Therefore, from (16) and (17),

$$R = 1.04 \times 10^{-6}\, nD\sqrt{f}/D \text{ ohms/cm. of line.} \quad (18)$$

The second term in (13) can be expressed in terms of the power factor F through the approximate relation

$$F = G/(\omega C). \quad (19)$$

Now, from (13), (1), (18), and (19) and the relation $nd = 1$, we obtain

$$A = \frac{4.55 \times 10^{-4}\sqrt{f}}{D}\frac{L_0}{L} + 2.72 \times 10^{-5} fF \text{ db per microsecond delay.} \quad (20)$$

The parameter in the first term of this formula will be in error because of the simplifying assumptions. The actual experimental results are best described by

$$A = \frac{7 \times 10^{-4}\sqrt{f}}{D}\frac{L_0}{L} + 2.7 \times 10^{-5} fF \text{ db per microsecond delay} \quad (21)$$

where F is about 1 per cent. Evidently, below 20 Mc. the resistive loss is the dominant one.

It is worthy of note that, to a first approximation, A is independent of the wire diameter d. It is possible, therefore, to plot universal attenuation curves for all lines having the same coil diameter D. Fig. 5 is a chart of this type which describes the experimental results within the experimental errors of about 10 per cent.

E. *Design Charts*

Figs. 5, 6, and 7 are charts which can be used in designing single-layer delay lines in which the outer conductor is a "closed" braid having a pitch of 1.9 inches or more, made up of 0.005-inch Formex-insulated round copper wire, and in which the inner and outer conductors are separated by a single wrap of $\frac{3}{8} \times 0.0015$-inch aceto-butyrate tape with 50 per cent overlap. Fig. 7 is a plot

[5] S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, New York, N. Y., Section 6.11, 1944.

of reciprocal of delay per unit length against characteristic impedance. Curves **I**, **II** and **III** are curves of constant coil diameter D, and are associated with the attenuation curves of Fig. 5. The dashed lines are lines of constant wire diameter d. Curves A, B, and C connect points having the same phase distortion as given by Fig. 6.

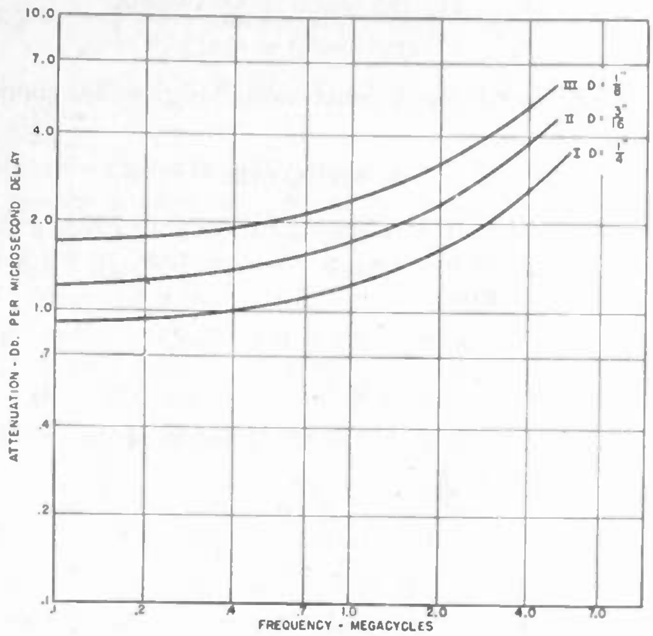

Fig. 5—Attenuation as a function of frequency of curves I, II, and III on the design chart, Fig. 7.

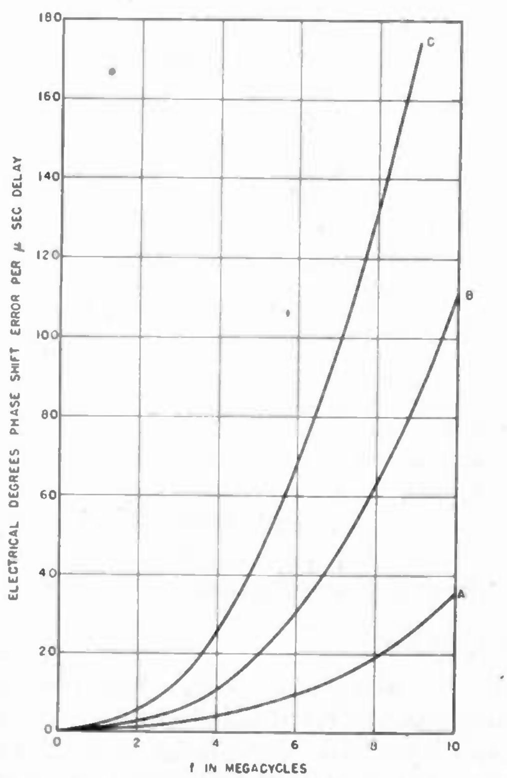

Fig. 6—Delay error as a function of frequency. Phase distortion of curves A, B, and C on the design chart, Fig. 7.

The choice of d and D is influenced by four factors: characteristic impedance, delay per unit length, phase distortion, and attenuation. However, any two of these factors uniquely determine d and D and the other two factors. The best design procedure seems to be to make a choice of values for the two factors having the most stringent requirements. d and D are then fixed, and small compromises can be made to arrive at suitable values for the other two factors.

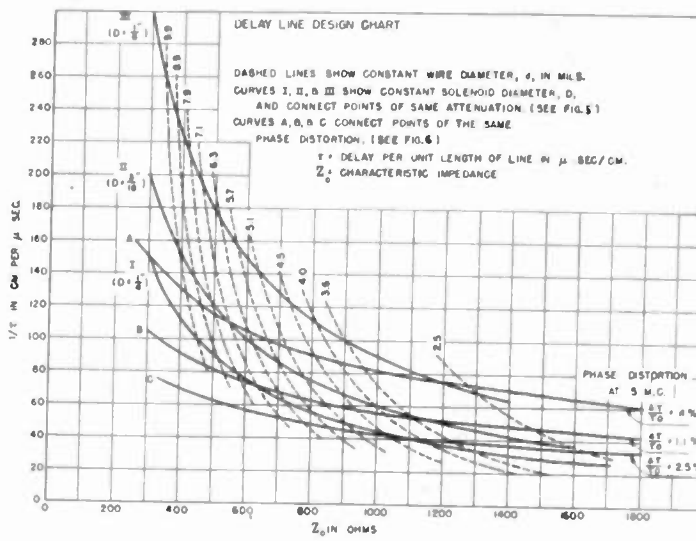

Fig. 7—Delay-line design chart.

V. Measurement of Delay-Line Characteristics

Delay per unit length is determined accurately and easily by resonating the line, either open- or short-circuited, and computing delay from several resonance points. Other methods are described by Kallmann.[4]

As Kallmann points out, measurement of attenuation is difficult. The measurements used in Fig. 5 were obtained by vacuum-tube-voltmeter readings at the input and ouput of a line. Even with the most careful attention to correct termination of the line, however, periodic fluctuations of the order of ± 10 per cent always appeared.

VI. Acknowledgments

The authors are indebted to H. E. Stevens for data on attenuation and for development of the design charts; to R. B. Nelson for early analyses of phase and dielectric distortions; to Mrs. M. H. Blewett and H. Poritsky for the analysis of variation of inductance with frequency; to R. E. Troell for many helpful suggestions including the use of Litz braid for a shield; and to J. M. Pettit of the Radio Research Laboratory for assistance in obtaining experimental data. F. J. Moles and G. Waggoner of the General Engineering Laboratory of the General Electric Company developed the use of aluminum spray for improvement of phase distortion, and evolved the various changes in technique which brought a laboratory product up to production standards and quantity.

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John P. Blewett (A'43) was born in Toronto, Canada, in 1910. He received the B.A. and M.A. degrees from the University of Toronto in 1932 and 1933, respectively, and the Ph.D. degree in physics from Princeton University in 1936. After a year at the Cavendish Laboratory in Cambridge, England, as a Royal Society of Canada Fellow, he joined the staff of the Research Laboratory of the General Electric Company in Schenectady, N. Y.

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Subsequently, Mr. Roe has participated in the development of orthicon field equipment and other types of television pickup equipment. During the war he was engaged in the development of radiating systems for Block equipment, and is currently in charge of the product line development of television terminal equipment in the RCA Victor Division, Camden, N. J. He is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

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For a photograph and biography of LYNN L. MERRILL, see the January, 1947, issue of the PROCEEDINGS OF THE I.R.E.

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Mr. Smith is the inventor of the commonly used transmission-line matching stub, and the optimum-impedance coaxial line (for high-frequency power transmission), as well as the cloverleaf antenna. He is the originator of the circular transmission-line reflection chart commonly identified with his name. Mr. Smith is at present serving as a member of the Antenna Committee of the I.R.E.

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Abstracts and References

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ACOUSTICS AND AUDIO FREQUENCIES

016:534 3386

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534.21 3387

The Propagation of an Acoustic Wave along a Boundary—I. Rudnick. (*Jour. Acous. Soc. Amer.*, vol. 19, pp. 348–356; March, 1947.) The sound field of a point source near the boundary of two media cannot be obtained by assuming plane waves and using the plane-wave reflection coefficient. A more rigorous solution is obtained similar to that given by Sommerfeld for the analogous electromagnetic case. The discussion of the solution is restricted to cases in which the sound source is at the boundary. It is shown that when the boundary medium has a high real specific acoustic impedance, nonzero fields are obtained at all points along the boundary. Calculations of the sound pressure as a function of height, when the media are air above and Quietone (an absorbing material) below, reveal a minimum some distanec above the boundary.

534.213 3388

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534.231:621.396.11 3389

A Device for Plotting Rays in a Stratified Medium—A. W. Lawson, P. H. Miller, Jr., and L. I. Schiff. (*Rev. Sci. Inst.*, vol. 18, pp. 117–120; February, 1947.) A description of an instrument used during the war for the computation of sound fields in water; it could also be applied to the propagation of radar signals in a stratified atmosphere.

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Measurement and Specification of Ultrasonic Lenses—P. J. Ernst. (*Jour. Acous. Soc. Amer.*, vol. 19, p. 474; May, 1947.) Methods analogous to those approved in optics are suggested.

534.321.9:621.317.49 3392

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534.321.9:[666.1+669.71 3393

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534.41+534.781 3395

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534.41+534.781]:535.37 3397

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534.41+534.781]621.385.832 3398

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534.43:621.395.61 3400

Measurement of Mechanical Compliance and Damping of Phonograph Pickups—B. B. Bauer. (*Jour. Acous. Soc. Amer.*, vol. 19, pp. 319–321; March, 1947.) The apparatus described can be adapted to measure vertical as well as lateral compliance and resistance of pickups, and the mechanical impedance of certain types of structures and damping materials.

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534.64 3402

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621.396.645:534.78 3403

The Theory and Design of Speech Clipping Circuits—Dean. (See 3477.)

534.782 3404

On an Artificial Voice for Acoustic Measurements—P. Chavasse. (*Compt. Rend. Acad. Sci.* (Paris), vol. 224, pp. 1620–1622; June 9, 1947.) Essentially a source of current with a continuous and uniform spectrum, formed by a neon tube working in a zone of unstable equilibrium. The tube is polarized by a d.c. voltage through a resistor and capacitor favoring l.f. oscillations. Change from a male voice, with a maximum voltage near 550 c.p.s. to a female voice, with maximum near 1100 c.p.s. is effected by a simple switch.

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High-Fidelity Reproduction of Music—E. Toth. (*Electronics*, vol. 20, pp. 108–113; June, 1947.) Practical suggestions for reducing various types of distortion which occur in a.m. and f.m. receivers and in phonograph record reproduction.

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A Distortion Reducing Stylus for Disk Reproduction—E. F. McClain, Jr. (*Jour. Accous. Soc. Amer.*, vol. 19, pp. 326–328; March, 1947.)

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Recording Studio Acoustics—L. Green, Jr., and J. Y. Dunbar. (*Jour. Acous. Soc. Amer.*, vol. 19, pp. 412–414; May, 1947.) Summary noted in 2307 of September. A general discussion, with some details of the treatment of several studios to obtain improved reverberation characteristics.

534.861/.862].1 3410
Convex Wood Splays for Broadcast and Motion Picture Studios—M. Rettinger. (*Jour. Acous. Soc. Amer.*, vol. 19, pp. 343–345; March, 1947.) Summary noted in 2307 of September.

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A Review of Criteria for Broadcast Studio Design—H. M. Gurin and G. M. Nixon. (*Jour. Acous. Soc. Amer.*, vol. 19, pp. 404–411; May, 1947.) Summary noted in 2307 of September.

534.862.3"1857/1926" 3412
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621.395.61 3413
Mechano-Electronic Transducers—H. F. Olson. (*Jour. Acous. Soc. Amer.*, vol. 19, pp. 307–319; March, 1947.) A system whereby a voltage is developed by the direct action of acoustical vibrations on one of the electrodes of a tube. The vibrating element can be made very small, with a low mechanical impedance. Various electrode arrangements are discussed and expressions derived for their electrical characteristics. For small amplitudes, the output voltage is proportional to the displacement. Equivalent mechanical circuits are shown and discussed in relation to the amplitude frequency characteristic. A successful transducer, consisting of a small triode in which the anode is the vibrating element, is illustrated. Details are also given of a phonograph pickup and two microphones. Their response characteristics are derived theoretically from the equivalent mechanical circuits, and measured characteristics are shown. Summary noted in 2307 of September. See also 2624 of September.

621.395.625.2 3414
Technics of Sound Recording with Embossed Groove Methods—L. Thompson. (*Tele-Tech*, vol. 6, pp. 48–51 and 115; May, 1947.) Discusses the advantages of these methods for business use. Reasonable fidelity and volume are obtained at slow track speed. Photographs and frequency-response curves are given.

621.395.625.3 3415
Some Factors Influencing the Choice of a Medium for Magnetic Recording—L. C. Holmes. (*Jour. Acous. Soc. Amer.*, vol. 19, pp. 395–403; May, 1947.) Summary noted in 2307 of September. A definition of signal-to-noise ratio for magnetic recording systems is offered to stimulate discussion. Modulation noise, background noise, cross talk, and uniformity are considered. The ratio of coercivity to retentivity is suggested as a figure of merit for evaluating the h.f. response of a recording medium.

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621.395.625.6:534.862.4 3417
A New Method of Counteracting Noise in Sound-Film Reproduction—W. K. Westmijze. (*Philips Tech. Rev.*, vol. 8, pp. 97–104; April, 1946.) When the incident light beam is replaced by a series of equidistant light spots moving with high velocity, perpendicular to the sound track, noise arising from dust and scratches on the film is reduced considerably.

621.395.645:621.395.614]:621.395.623.8 3418
Microphone Pre-Amplifier—Selby. (*See* 3457.)

621.395.813:621.395.66:621.396.97 3419
Compensation of Temperature Effects on Music Circuits—F. J. Stringer and G. Stannard. (*B.B.C. Quart.*, vol. 2, pp. 41–50; April, 1947.) An account of apparatus and technique developed by the British Broadcasting Corporation and the Government Printing Office to compensate for changes in response characteristics of long-distance line circuits due to seasonal changes in temperature. These changes in response characteristics are calculated theoretically and compared with actual measurements made on particular circuits.

AERIALS AND TRANSMISSION LINES

621.315.687 3420
Cable Terminations—D. B. Irving. (*Jour. I.E.E.* (London), Part II, vol. 94, pp. 123–128; April, 1947.) Discussion on 3360 of 1945.

621.392+537.291]:[621.385.029.63/.64 3421
On the Theory of Progressive-Wave Amplifiers—A. Blanc-Lapierre, P. Lapostolle, J. P. Voge, and R. Wallauschek. (*Onde Élec.*, vol. 27, pp. 194–202; May, 1947.) An integration and amplification of previous papers (1317 and 1330 of June, 1999 and 2003 of August).

621.392.029.64 3422
Wave-Guide Coupler—G. Ashdown. (*Jour. Sci. Inst.*, vol. 24, p. 79; March, 1947.) For aligning and joining sections.

621.392.029.64 3423
Transmission in Waveguides—A. M. Woodward. (*Wireless Eng.*, vol. 24, pp. 192–196; July, 1947.) "The theory of transmission of an H_{01} wave in a rectangular waveguide, containing longitudinal slabs of solid dielectric, is developed. Formulas are given for phase constant and attenuation which take into account imperfect conductivity of the guide walls as well as losses in the dielectric. Numerical values are given for polythene as dielectric. At high frequencies, most of the energy traveling down the guide is confined to the dielectric slabs. The phase constant and attenuation are then nearly equal to their values for a completely filled guide."

621.392.2 3424
Use of [Transmission] Lines as Resonant Circuits—A. Fournier. (*Rev. Sci.* (Paris), vol. 84, pp. 624–629; December 1–15, 1946.) The lines are considered as having lumped constants and the equations of propagation are derived. Other sections discuss (a) characteristic impedance, (b) reflection of waves and stationary waves, (c) impedance transformation, (d) equivalence of quarter-wave line and resonant circuit, (e) lines matched to tube capacitances, (f) charged lines, (g) coupled lines, and (h) link coupling.

621.396.67 3425
Circularly-Polarised Omnidirectional Antenna—G. H. Brown and O. M. Woodward, Jr. (*RCA Rev.*, vol. 8, pp. 259–269; June, 1947.) The combination of a vertical dipole and horizontal loop requires too critical adjustment. Four dipoles spaced around the circumference of a horizontal circle and inclined to its plane give a good approximation. A theoretical discussion is followed by the results of tests over a frequency range of 106 to 134 Mc.

621.396.67 3426
Antenna Focal Devices for Parabolic Mirrors—G. Reber. (Proc. I.R.E., vol. 35, pp. 731–734; July, 1947.) The measured characteristics of cone and cylindrical aerials in parabolic mirrors are given. The relationships between geometrical and electrical characteristics are discussed.

621.396.67 3427
Gain vs. Element Spacing in Parasitic Arrays—R. G. Rowe. (*QST*, vol. 31, pp. 30–35; April, 1947.) Results of measurements are given, showing the relation between gain and spacing under controlled conditions and with a definite technique. Wider spacing than is commonly used is shown to give greater gain.

621.396.67 3428
Input Impedance of a Folded Dipole—W. van B. Roberts. (*RCA Rev.*, vol. 8, pp. 289–300; June, 1947.) A theoretical discussion of the cases in which the dipole has equal and unequal elements. Consideration is also given to finite spacing of the elements.

621.396.67:517.512.2 3429
Fourier Transforms in Aerial Theory: Part 2—Ramsay. (*See* 3561.)

621.396.67:621.317.772.029.64 3430
Phase-Front Plotter for Centimeter Waves—Iams. (*See* 3590.)

621.396.67:621.396.712:621.396.619.13 3431
Antennas for F.M. Broadcasting: Part 2—N. Marchand. (*Communications*, vol. 27, pp. 24–26 and 37; May, 1947.) Discussion of the principles of operation and construction of the clover-leaf, slot, and turnstile aerials. (For Part 1 see 3030 of November.)

621.396.67.002.72:621.397.5 3432
The WABD Super-Turnstile TV Antenna Installation—A. W. Deneke. (*Communications*, vol. 27, pp. 12–15 and 38; May, 1947.) Consideration of the factors influencing the installation and testing of an aerial and feeder system on an 80-foot tower on the roof of a skyscraper.

621.396.67.029.62:621.396/.397].62 3433
Aerials for Ultrashort Waves: Part 1—A Double Dipole for Television and F.M.—R. D. A. Maurice. (*B.B.C. Quart.*, vol. 2, pp. 59–62; April, 1947.) The dipole receives 45- and 90-Mc. transmissions simultaneously. Its

performance when suitably oriented differs little from that of an ordinary dipole.

621.396.672.029.62:621.396.65 3434
Aerials for Ultrashort Waves: Part 2—A Simple Omnidirectional Aerial with Concentric Feeder—H. L. Kirke. (*B.B.C. Quart.*, vol. 2, pp. 62–64; April, 1947.) An end-fed, vertical $\lambda/4$ folded dipole is used to ensure correct impedance matching to the feeder; this is particularly important for the receiving aerial in u.s.w. radio links. A circular polar diagram is obtained by means of an "artificial earth."

621.396.675:550.837 3435
Electric Field of an Oscillating Dipole on the Surface of a Two-Layer Earth—A. Wolf. (*Geophys.*, vol. 11, pp. 518–534; October, 1946.) "The electric field of a low-frequency oscillator placed on the surface of a two-layer earth is determined in two special cases: namely, the case in which the conductivities of the two layers are nearly equal, and the case in which the lower layer is a perfect insulator; in the latter case, only terms of zero and first order in frequency are considered. It is shown that, when the upper layer is sufficiently thin or is very thick, the mutual inductance of two wire elements on the surface of a two-layer earth has the same value as for a homogeneous earth. In the case of an insulated layer, it is shown that the maximum departure of the value of mutual inductance of two colinear wire elements from the corresponding value on a homogeneous earth is 35 per cent."

621.396.675:550.837 3436
Electric Field of an Oscillating Dipole at the Surface of a Two-Layer Earth—W. B. Lewis. (*Geophys.*, vol. 11, pp. 535–537; October, 1946.) Discussion on 3435 above. Field measurements of the transverse component E_y of the dipole field show that its value is dependent on frequency, whereas according to Wolf's rigorous solution (above) E_y should be independent of frequency. The experimental results are in agreement with the solution of the problem of a dipole oscillator in a homogeneous medium. The disagreement of the measurements with the more rigorous theory and the agreement with theory that neglects the air-earth boundary is paradoxical.

621.396.679.4 3437
Choosing a Transmission Line—R. M. Purinton. (*QST*, vol. 31, pp. 39–44 and 118; June, 1947.) A comparison of various types of feeders and considerations governing the choice for particular installations.

621.396.677:621.396.97 3438
Directional Antennas [Book Review]—C. E. Smith, Cleveland Institute of Radio Electronics, Cleveland, 1946, 298 pp., $15.00 (PROC. I.R.E., vol. 35, p. 706; July, 1947.) "... Of particular interest to those concerned with the design of vertical-tower directional antennas for broadcast stations. A thorough, systematic engineering treatment. ..."

CIRCUITS AND CIRCUIT ELEMENTS

621.314.23:621.396.69 3439
Toroidal Transformers—A. L. Morris. (*Electronic Eng.*, vol. 19, pp. 218–219; July, 1947.) These can effect a considerable saving in space and weight. Special winding machines are required and insulation and winding difficulties may be encountered in high-voltage windings, but toroids are very suitable for multiwinding filament transformers, especially in aircraft equipment where supply frequencies of 400 or 1600 c.p.s. are used and the number of turns in each winding can be reduced correspondingly. Developments in silicon-iron core material may make toroids advantageous for 50 c.p.s. working. For an account of a winding machine see 3672 below.

621.316.722.1.076.7 3440
Diode-Controlled Voltage Regulators—H. Helterline. (*Electronics*, vol. 20, pp. 96–97; June, 1947.) Circuit details of a bridge-type regulator, in which the filament of a temperature-limited diode acts as control element. A.c. output voltage is constant within 0.2 per cent over 10 to 1 load variation. The stability of a d.c. version equals that of batteries.

621.316.89+621.315.59 3441
Properties and Uses of Thermistors—Thermally Sensitive Resistors—J. A. Becker, C. B. Green, and G. L. Pearson. (*Bell Sys. Tec. Jour.*, vol. 26, pp. 170–212; January, 1947.) Reprint of 765 of April.

621.318.323.2.042.15 3442
Permeability of Dust Cores—Legg. (*See* 3555.)

621.318.572 3443
Self-Switching R.F. Amplifier—H. M. Wagner and J. F. Herrick. (*Electronics*, vol. 20, pp. 128–131; June, 1947.) A twin-pentode multivibrator circuit amplifies and automatically switches two circuits into a common indicator for direction-finding purposes. Design considerations, switching ratios, and input-resistance variations are discussed.

621.318.572 3444
Vane-Actuated Controller—W. H. Wannamaker, Jr. (*Electronics*, vol. 20, pp. 117–119; June, 1947.) Movement of a vane between the coils of a double-triode r.f. oscillator, causes a sudden change of anode current, which actuates an output relay. Full circuit details are given, with several industrial applications.

621.318.572:621.317.755 3445
A Laboratory Four-Channel Electronic Switch—F. S. Replogle, Jr., and V. M. Albers. (*Rev. Sci. Instr.*, vol. 18, pp. 114–117; February, 1947.) Permits the simultaneous presentation of four signals of frequency up to 100 kc. on a c.r.o. screen. See also 1780 of 1946 (Moerman).

621.319.53 3446
A Simple Pulse Converter for Gas-Tube Applications—L. Reiffel and K. Rothschild. (*Rev. Sci. Instr.*, vol. 18, pp. 181–183; March, 1947.) Pulses of either polarity are fed through a resistance network to the screen grid of a thyratron, whose electrodes are biased to give monopolar output pulses. The anode circuit is used to provide pulse shaping.

621.392.2 3447
Use of [Transmission] Lines as Resonant Circuits—Fournier. (*See* 3424.)

621.392.21 3448
On the Short-Circuiting of a Charged Transmission Line—V. L. Ginzburg. (*Bull. Acad. Sci.* (U.R.S.S.) sér. phys., vol. 10, pp. 57–64; 1946. In Russian.) A transmission line with uniformly distributed circuit parameters is considered. Initially, it is open-circuited and charged. A general equation is derived for the self-oscillations in the line when short-circuited through a loading inductance, and solutions are found for two particular values of this inductance.

621.392.43:621.396.615.141.2 3449
Microwave Generator—W. C. Brown. (*Tele-Tech*, vol. 6, p. 59; May, 1947.) Summary of an Institute of Radio Engineers' paper. The effect of a mismatched transmission line on the frequency stability and power output of a magnetron is studied by means of an equivalent circuit.

621.392.5:621.396.622.6 3450
Crystal Networks—L. Apker, E. Taft, and J. Dickey. (*Tele-Tech*, vol. 6, p. 54; May, 1947.) Summary of an Institute of Radio Engineers' paper. Discusses the case where the insertion loss of a nonlinear quadripole is different for power transmitted in opposite directions. Results of tests on 20 Si and Ge crystals are given.

621.392.5.015.3 3451
Network Distortion—M. J. DiToro. (*Tele-Tech*, vol. 6, pp. 55–56; May, 1947.) Summary of an Institute of Radio Engineers' paper. The transient response of a network to a step signal may be used as a measure of the distortion to be expected. Curves to facilitate the study of transient response are shown together with design data for correcting networks.

621.392.52 3452
A Simplified Analysis of the Parallel-T Null Network—M. P. Givens and J. S. Saby. (*Rev. Sci. Instr.*, vol. 18, pp. 342–346; May, 1947.) "The parallel-T resistance-capacitance null network is analyzed algebraically. General conditions for null, and expressions for network impedance and sharpness of null, are obtained in a convenient form for application to practical design problems. Vector diagrams are used to illustrate the variations of phase and amplitude of output voltage with frequency."

621.392.52 3453
Analysis of a Resistance-Capacitance Parallel-T Network and Applications—A. E. Hastings. (PROC. I.R.E., vol. 35, p. 694; July, 1947. Correction to 1464 of 1946.

621.394/.397].645 3454
Cathode-Follower Circuit—H. L. Krauss. (PROC. I.R.E., vol. 35, p. 694; July, 1947.) Comment on 1025 of May (McIlroy). See also 1373 of June.

621.394/.397]645:518.3 3455
Cathode-Follower Nomograph for Pentodes—M. B. Kline. (*Electronics*, vol. 20, p. 136; June, 1947.) Gives relation between gain, transconductance, and cathode-load resistance.

621.394/.397].645.34 3456
A Variation on the Gain Formula for Feedback Amplifiers for a Certain Driving-Impedance Configuration—T. W. Winternitz. (*Bell Sys. Tech. Jour.*, vol. 26, p. 216; January, 1947.) Summary of 50 of February.

621.395.645:621.395.614]:621.395.623.8 3457
Microphone Pre-Amplifier—R. Selby. (*Wireless World*, vol. 53, pp. 239–240; July, 1947.) Cathode-follower circuit suitable for public-address work.

621.395.661 3458
Mica Capacitors for Carrier Telephone Systems—A. J. Christopher and J. A. Kater. (*Bell Sys. Tech. Jour.*, vol. 26, p. 213; January, 1947.) Summary of 374 of March.

621.396.611.3.029.56 3459
Coupled-Circuit Oscillators—D. K. Cheng. (*Tele-Tech*, vol. 6, pp. 58–59; May, 1947.) Summary of an Institute of Radio Engineers' paper. Measurements of wavelength and loading characteristics of a 2-Mc. coupled-circuit oscillator show good correlation with theoretical predictions. Conclusions concerning the optimum degree of coupling and magnitude of the external load resistance are stated.

621.396.611.4 3460
The Simplest Design Calculations of Certain Cavity Resonators—V. M. Lopukhin. (*Bull. Acad. Sci.* (U.R.S.S.) sér. phys., vol. 10, pp. 111–116; 1946. In Russian.) Approximate formulas are derived for calculating the Q and impedance of the following resonators: simple toroidal (Fig. 1), quasi-toroidal (Fig. 2), π type (Fig. 3), and cylindrical (Fig. 4).

621.396.611.4 3461
On the Self-Excitation of a Cavity Reso-

nator Traversed by an Electron Beam—S. Gvozdover and V. Lopukhin. (*Bull. Acad. Sci.* (U.R.S.S.) sér. phys., vol. 10, pp. 29–36; 1946. In Russian.)

621.396.611.4 3462

Coupling between Cavity Resonators Through Small Apertures—V. B. Brodski. (*Bull. Acad. Sci.* (U.R.S.S.) sér. phys., vol. 10, pp. 17–22; 1946. In Russian.) A mathematical investigation of the effect on the fields inside two resonators of a small aperture in the common wall.

621.396.611.4 3463

Flat Cavities as Electrical Resonators—C. G. A. von Lindern and G. de Vries. (*Philips Tech. Rev.*, vol. 8, pp. 149–160; May, 1946.) The characteristic vibrations of Lecher systems short-circuited at one end are first considered. It is then shown that in the case of conical flat cavity resonators, short-circuited around their outer edge, the rotation-symmetrical vibrations correspond exactly to those of the short-circuited Lecher systems. The rotation-symmetrical vibrations of flat resonators of more general forms are discussed and curves are given for the variation of current and voltage with the radius. Resonance resistance and quality factor are calculated; these can be improved by making the cavity resonators thicker than those for which the theory given applies unconditionally. Examples are given of practical resonators and of their use for h.f. stabilization or as output and input electrodes for short-wave transmitting tubes.

621.396.611.4 3464

End-Plate and Side-Wall Currents in Circular Cylinder Cavity Resonator—J. P. Kinzer and I. G. Wilson. (*Bell Sys. Tech. Jour.* vol. 26, pp. 31–79; January, 1947.) Formulas are given for the calculation of the current streamlines and intensity in the walls of a circular cylindrical cavity resonator. Tables are given which permit calculation for many of the lower order modes.

The integration of $\int_0^x [J_l(x) | J_l'(x)] dx$ is discussed and the integral is tabulated for $l=1$, 2, and 3.

The current distribution for a number of modes is shown by plates and figures.

621.396.611.4:621.396.662.3.029.64 3465

Cavity Resonators—M. W. Wheeler. (*Tele-Tech*, vol. 6, p. 60; May, 1947.) Summary of an Institute of Radio Engineers' paper. Discussion of their characteristics and use as u.h.f. band-pass filters.

621.396.611.4.029.64:621.396.662 3466

3-cm. Resonant Cavity—R. R. Reed. (*Tele-Tech*, vol. 6, pp. 54–55; May, 1947.) Summary of an Institute of Radio Engineers' paper. A transmission-type cavity for use in an automatic frequency-control circuit. The resonant frequency is nearly independent of temperature and humidity effects and may be pretuned accurately. Source and load are coupled to the cavity by "Kovarglass" windows; temperature compensation is effected by altering the length of the cavity by a flexible diaphragm moved by the differential expansion of copper and invar.

621.396.615:621.316.726.078.3 3467

Synchronization of Oscillators—R. D. Huntoon and A. Weiss. (*Jour. Res. Nat. Bur. Stand.*, vol. 38, pp. 397–410; April, 1947.) An analysis of the behavior of any self-limiting oscillator when a sinusoidal current or voltage of small but consistant magnitude is injected into it. The synchronization band is proportional to the injected voltage. The theory was checked by measurements on a small Hartley oscillator at 11.5 Mc. The analysis includes the mutual synchronization of two oscillators. For synchronization measurements the driving oscillator must be more powerful than the test oscillator. Applications of the synchronized oscillator include (a) linear voltmeter for small voltages, (b) field-intensity meter, (c) linear a.m. demodulator for small signals, (d) f.m. demodulator, (e) f.m. synchronous amplifier limiter. In these applications, microwave generators can be used as well as the more conventional triode oscillators.

621.396.615.14 3468

The Excitation of Resonant Circuits by Electron Currents in the Transit-Time Domain—F. W. Gundlach. (*Rev. Sci.* (Paris), vol. 85, pp. 19–28; January 1, 1947.) Translation into French of paper to appear in *Hochfrequenztechnik*. A method is described which gives the magnitude of both the in-phase and quadrature components of the circuit current induced by an electron current through a tube grid. The intensity and velocity of the electron current may vary in any periodic manner with time. The method is applicable to all possible cases and a series of abacs is provided. Application is made to the Barkhausen-Kurz oscillator.

621.396.615.14 3469

Band-Switched Exciter—P. W. J. Gammon. (*Short Wave Mag.*, vol. 5, pp. 16–20; March, 1947.) A switched-coil oscillator and frequency-doubler circuit for driving high-powered output stages on frequencies between 1.7 and 28 Mc.

621.396.615.142 3470

The Principles of a General Theory of the Generation of Electron Oscillations at Ultra-High Frequencies—V. I. Kalinin. (*Bull. Acad. Sci.* (U.R.S.S.) sér. phys., vol. 10, no. 1, pp. 93–102; 1946. In Russian.) "Electron oscillations" are defined as oscillations, the excitation of which depends ultimately on the inertia of electrons. A general scheme of the different phases in the excitation of such oscillations is presented in a graphical form (Fig. 1) and using Brüche and Recknagel's conception of "phase focusing" (2325 of 1938), the foundations are laid of a theory which would not only explain the oscillation mechanism but also answer questions relating to the energy balance in an electron oscillator.

The first two fundamental equations (I and II) cover the kinetic side of the problem and determine respectively the current in the oscillator and the condition necessary for the formation of a focus. Two energy equations (III and IV) determining respectively the power output and efficiency of the oscillator are also derived.

The main factor determining the character of any particular modification of the oscillating system is the distribution and behavior of potentials in the transformation zone where the velocity-modulated beam is subjected to the action of an electric field with a potential varying in space and time in a known manner. If the electric field is absent and the velocity-modulated beam is moving by inertia, the simplest case corresponding to a two-circuit klystron is obtained. Conclusions reached in studying this case are enumerated briefly and as a further illustration of the proposed theory, the operation of an oscillator with a retarding field in the transformation zone is discussed in detail.

621.396.615.17:621.396.663 3471

On a Standardized Aperiodic Pulse Generator and Its Application to the Statistical Recording and Radiogoniometry of Atmospherics—F. Carbenay. (*Compt. Rend. Acad. Sci.* (Paris) vol. 224, pp. 1624–1626; June 9, 1947.) Apparatus similar in principle to that used by R. Bureau for atmospheric recording, but including a standardized variable inductive coupling between the capacitor discharge circuit and either the aerial or the input circuit of the receiver-recorder. Some circuit details are given and the methods of use and standardization are described briefly.

621.396.621.54 3472

The Inversion of the Autodyne Principle—F. C. Saic. (*Elek. Nach. Tech.*, vol. 64, pp. 16–24; January and February, 1947.) A new type of heterodyne arrangement is described in which the oscillator frequency is fixed and the i.f. variable. Numerous advantages are claimed. A scheme is given for an all-wave receiver incorporating the new principle and giving an appreciable increase in output power.

621.396.622.71 3473

The Ratio Detector—Seeley and Avins. (*See* 3643.)

621.396.645 3474

Theory of Grounded-Grid Amplifiers—A. van der Ziel. (*Philips. Res. Rep.*, vol. 1, pp. 381–399; November, 1946.) In part 1 a survey of the existing triode theory at u.h.f. is given. Neglecting lead effects, the four characteristic impedances of a grounded-grid triode at u.h.f. can be described by the "cold" tube capacitances, the amplification factor μ and the transconductances S_1 and S_2 in the cathode-grid lead and grid-anode lead, respectively (the moduli and the phase angles of these transconductances can be measured). Shot effect in triode tubes can be described completely by assuming two mutually dependent fluctuating currents i_1 and i_2 to be flowing in the cathode-grid lead and in the grid-anode lead, respectively; at u.h.f. i_2 is delayed in phase with respect to i_1 (the introduction of these mutually dependent fluctuating currents is a direct consequence of Fourier analysis of the shot effect). It is shown that the introduction of the "equivalent noise resistance" of the tube may cause serious errors in the calculation of the signal-to-noise ratio.

In part 2 this theory is applied to grounded-grid amplifiers. The input resistance R_1 of the tube when the output is short-circuited, and the output resistance R_2 of the tube when the input is short-circuited, are of special importance in this case. Denoting the transformed aerial resistance by R_1' and the transformed input resistance of the next stage by R_2', the power gain g is calculated as a function of R_1'/R_1 and R_2'/R_2. It is shown that the internal feedback of the tube makes it impossible to match, at the same time, the aerial to the input of the amplifier and its output to the input of the next stage. The best results are obtained by using a high value of R_1'/R_1 (loose aerial coupling) and matching the output of the amplifier to the next stage. The theoretical gain limit is $(\mu+1)$; values between 0.5 $(\mu+1)$ and 0.8 $(\mu+1)$ may be obtained easily. For wide-band amplifiers $R_1'/R_1=1$ for maximum gain, whereas it is shown that a wide anode-grid spacing will give a higher gain. It is shown that electronic transit times cannot account for the drop in power gain at u.h.f.; this drop must be due to the impedance of the electrode leads. At u.h.f. instability may occur; a stability condition is given, from which it can be seen that careful shielding and narrow electrode spacings result in a better stability of the amplifier. Finally the signal-to-noise ratio of the grounded-grid amplifier is calculated and it is shown that the grounded-grid amplifier contributes only slightly to the noise, especially for large values of R_1'/R_1. This result is verified experimentally.

621.396.645 3475

Nonstationary Processes in Tuned and Band-Pass Amplifiers—A. N. Shchukin. (*Bull. Acad. Sci.* (U.R.S.S.) sér. phys., vol. 10, pp. 37–48; 1946. In Russian.) A mathematical investigation of the processes taking place in amplifiers when a constant or an alternating e.m.f. is suddenly applied.

621.396.645 3476
A Note on a Paper by Faust and Beck—W. M. Stone. (*Jour. Appl. Phys.*, vol. 18, pp. 414–416; April, 1947.) "An infinite sum transformation is defined and applied to a system of linear difference equations discussed by Faust and Beck in their paper on single-tuned amplifiers [677 of April]. Some transforms of the more common functions are given and points of superiority of the transform method over the classical methods of solution of difference equations are emphasized."

621.396.645:534.78 3477
The Theory and Design of Speech-Clipping Circuits—M. H. Dean. (*Tele-Tech*, vol. 6, pp. 62–65 and 119; May, 1947.) The action of a compressor in preventing overmodulation of a.m. transmitters is described, and is shown to be less effective than might be expected. Clipping speech peaks squarely at a predetermined level, and inserting a low-pass filter to eliminate any harmonics caused thereby, is considered better, and the design of a clipper suitable for good commercial speech transmissions is given.

621.396.645.029.3 3478
A Portable Two-Channel Amplifier and Ink Recorder—W. Grey Walter and A. A. Brooks. (*Electronic Eng.*, vol. 19, pp. 221–226; July, 1947.)

621.396.645.029.62 3479
Broad-Band Amplifiers—A. M. Levine and M. G. Hollabaugh. (*Tele-Tech*, vol. 6, p. 58; May, 1947.) Summary of an Institute of Radio Engineers' paper. Calculations for input damping and instability due to feedback are outlined. Calculated and measured values are given for various tube types throughout the 30- to 300-Mc. range. Measurements taken on actual amplifiers are displayed graphically.

621.396.645.029.63 3480
550-Megacycle Amplifier—R. C. Petrich. (*Tele-Tech*, vol. 6, p. 59, May, 1947.) Summary of an Institute of Radio Engineers' paper. A gain of 10 db for each of 5 stages has been obtained for a 20 Mc. bandwidth using a lighthouse triode in a grounded-grid amplifier circuit.

621.396.645.029.64:621.396.615.142.2 3481
On U.H.F. Amplification and on the Resonance Method for Suppressing Noise in a Klystron—Yu. A. Katsman. (*Bull. Acad. Sci.* (U.R.S.S.) sér. phys., vol. 10, no. 1, pp. 23–28; 1946. In Russian.) The use of klystrons at the input stage of a radio receiver at frequencies greatly exceeding 600 Mc. is limited by the high level of tube noise. To overcome this difficulty it is suggested that a resonant oscillatory circuit absorbing the noise energy should be connected between the cathode and the input electrodes of the tube.

621.396.645.35:621.317.71/.72 3482
Stabilized D.C. Amplifier with High Sensitivity—H. S. Anker. (*Electronics*, vol. 20, pp. 138, 140; June, 1947.) Designed to measure very small currents or voltages from a high-impedance source.

621.396.645.37 3483
Feed-Back Amplifiers—J. A. Rado, A. M. Levine, and H. G. Hollabaugh. (*Tele-Tech*, vol. 6, p. 55; May, 1947.) Summary of an Institute of Radio Engineers' paper. Analysis of the generalized feed-back amplifier shows that it can be regarded as a ladder network with negative conductance shunt arms. Mathematical analysis shows that these amplifiers have a gain-bandwidth capacitance equal to that of an ideal amplifier. In actual amplifiers, the ideal has been approached very closely.

621.396.645.37.029.3 3484
A Stable Selective Audio Amplifier—J. M. Sturtevant. (*Rev. Sci. Instr.*, vol. 18, pp. 124–127; February, 1947.) A narrow-band amplifier using both frequency-dependent and independent degeneration to secure stability and linearity.

621.396.662 3485
Electronic Attenuators—F. W. Smith, Jr., and M. C. Thienpont. (*Communications*, vol. 27, pp. 20–22; May, 1947.) Continuously variable attenuation over a wide frequency range is achieved by varying the cathode load of a cathode follower. The influence of various factors on design is discussed and a typical attenuator described.

621.396.662.3.029.3 3486
Tuned A.F. Filters: Part 1—H. E. Styles. (*Wireless World*, vol. 53, pp. 242–244; July, 1947.) General considerations and design formulas.

621.396.69+621.317.7+621.38 3487
The Physical Society's Exhibition—(*See* 3581.)

621.396.69:621.315.3 3488
Stamped Wiring—W. MacD. (*Electronics*, vol. 20, pp. 82–85; June, 1947.) Basically, a series of vertical and horizontal conducting strips, separated by a thin sheet of insulator, with interconnection by eyelets or pins.

621.397.335 3489
New Techniques in Synchronizing-Signal Generators—E. Schoenfeld, W. Brown, and W. Milwitt. (*RCA Rev.*, vol. 8, pp. 237–250; June, 1947.) The pulse edges are established by means of a terminated artificial transmission line carrying 31.5-kc. trigger impulses and their number during each framing interval is determined by an electronic counter. The locked-in relationship between line and field scanning frequencies makes use of the cascade-binary type of frequency divider.

621.38/.39].01 3490
Fundamentals of Industrial Electronic Circuits [Book Review]—W. Richter. McGraw-Hill Book Co., London and New York, 1947, 569 pp., $4.50. (*Elec. Rev.* (London), vol. 140, p. 930; June 6, 1947; and PROC. I.R.E., vol. 35, p. 707; July, 1947.) The interest of the book is not confined to industrial electronics. The explanations of fundamentals and tubes are so good that those mainly interested in radio would do well to study it. Intended as a text of intermediate standard for use in evening classes.

GENERAL PHYSICS

535.13:512.831 3491
On the Matrix Form of Maxwell's Equations—J. Baudot. (*Compt. Rend. Acad. Sci.* (Paris) vol. 224, pp. 1622–1624; June 9, 1947.) See also 2475 of September.

535.215:621.383.4 3492
Lead Sulphide Photoconductive Cells—Sosnowski, Starkiewicz, and Simpson. (*See* 3709.)

536.21:517.942.9 3493
Heat Conduction in Elliptical Cylinder and an Analogous Electromagnetic Problem—N. W. McLachlan. (*Phil. Mag.*, vol. 37, p. 216; March, 1946.) Correction to 3570 of January.

536.422 3494
The Escape of Molecules from a Plane Surface into a Still Atmosphere—K. J. Brookfield, H. D. N. Fitzpatrick, J. F. Jackson, J. B. Matthews, and E. A. Moelwyn-Hughes. (*Proc. Roy. Soc. A*, vol. 190, pp. 59–67; June 17, 1947.)

536.483 3495
A Helium Cryostat—S. C. Collins. (*Rev. Sci. Instr.*, vol. 18, pp. 157–167; March, 1947.) For temperatures down to 2°K. Three types of expansion device are described.

537.122:538.3 3496
The Electron and Electromagnetic Theory—G. Darrieus. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 249–264; May, 1947.) Certain difficulties of the classical theory are discussed and an outline is given of a modified Born-Infeld nonlinear theory.

537.228.1:512.9 3497
First-and-Second-Order Equations for Piezoelectric Crystals Expressed in Tensor Form—W. P. Mason. (*Bell Sys. Tech. Jour.*, vol. 26, pp. 80–138; January, 1947.) The phenomena occurring on application of electric fields, stresses and temperature changes are examined. The nine first-order effects are considered for the 32 types of crystal and measurement methods are discussed. Second-order effects dealt with are: elastic constants dependent on the applied stress and electric displacement, the electrostrictive effect, piezoelectric constants dependent on the applied stress, and the piezo-optical and electrooptical effects. These second-order equations may be used to examine the phenomena occurring in ferro-electric type crystals, and are applied to the case of Rochelle salt.

537.32 3498
Thermoelectric Properties of Conductors: Part 1—L. Gurevich. (*Jour. Phys.*, (U.S.S.R.) vol. 9, no. 6, pp. 477–488; 1945.) A new possible mechanism for thermoelectric e.m.f. is the carrying of electrons by the phonon current created by the temperature gradient. In a certain temperature range, this e.m.f. may greatly exceed that to be expected from the usual theory, and observed anomalies may be due to the transition from one mechanism to another.

537.525.5+621.314.65 3499
On the Mechanism of Dielectric Ignition and Resistance Ignition in Mercury Arc Rectifiers [Thesis]—Warmoltz. (*See* 3670.)

537.527 3500
Similarity of High-Pressure Discharges of the Convection-Stabilized Type—W. Elenbaas. (*Philips Res. Rep.*, vol. 1, pp. 339–359; November, 1946.) Similarity conditions are deduced for discharges in free air, or in tubes so wide that the walls have no effect. Pressures considered are so high that energy loss by radiation, dissociation and diffusion is negligible. Similarity conditions for discharges in various gases, discharges stabilized by forced convection and discharges in closed tubes filled with various gases are also considered.

538 3501
Magnetism—R. M. Bozorth. (*Rev. Mod. Phys.*, vol. 19, pp. 29–86; January, 1947.) A general descriptive account of the whole subject, taken from the American edition of the Encyclopedia Britannica. Magnetic theory is treated historically down to the modern "electron-spin" theory. A section is devoted to the measurement of magnetic quantities and references are given to modern papers on this subject. A bibliography of textbooks is appended.

538.532:621.316.974:621.318.4 3502
The Field of a Coil between Two Parallel Metal Sheets—E. B. Moullin. (*Jour. I.E.E.* (London), Part I, vol. 94, p. 158; March, 1947.) Summary of 2077 of August.

538.56 3503
Relative Directions of the Electric and Magnetic Vectors in Electromagnetic Waves in Vacuo—N. S. Japolsky. (*Nature* (London), vol. 159, pp. 580 and 817; April 26, and June

14, 1947.) In general, the electric vector E and the magnetic vector B will not be perpendicular for electromagnetic waves in vacuo. They are perpendicular in the special cases of (a) nonrotating vectors, (b) circularly polarized plane waves and (c) spherical or cylindrical waves.

538.56:535.13 3504
The Reflection of an Electromagnetic Plane Wave by an Infinite Set of Plates: Part 2—A. E. Heins and J. F. Carlson. (*Quart. Appl. Math.*, vol. 5, pp. 82–88; April, 1947.) In part 1 (2756 of October), the case was treated in which only one component of the electric field was excited, the incident electric field being parallel to the edges of the plates. Fourier transform technique is again used when the excitation is by a plane wave which has only a single component of the magnetic field parallel to the edges of the plates. In this case, it is found that the reflection and transmission coefficients are independent of the wavelength and depend only on the angle of stagger of the plates and the angle of incidence of the waves.

538.566 3505
[One-Dimensional] Propagation of a Perturbation, of Narrow Frequency Range, in a Nonabsorbing Dispersive Medium—A. Blanc-Lapierre and P. Lapostolle. (*Rev. Sci.* (Paris), vol. 84, pp. 579–595; December 1–15, 1946.) The type of perturbation considered is that of filtered background noise or quasimonochromatic light. The harmonic analysis of such perturbations leads to a representation analogous to a Fourier integral. The notions of phase velocity and group velocity are analyzed and their limits of validity are given as functions of the width of spectrum considered and of the dispersive properties of the medium in the neighborhood of the mean frequency. The results of the analysis are summarized and discussed.

538.567.2 3506
The Biased Ideal Rectifier—W. R. Bennett. (*Bell Sys. Tech. Jour.*, vol. 26, pp. 139–169; January, 1947.) Methods of solution and results are given for the frequency response of devices with sharply defined transitions between the conducting and nonconducting portions of their characteristics.

538.569.4.029.64:546.171.1 3507
The Inversion Spectrum of Ammonia at Centimetre Wavelengths—B. Bleaney and R. P. Penrose. (*Proc. Roy. Soc. A*, vol. 189, pp. 358–371; May 1, 1947.) Measurement technique and results for wavelengths between 1.1 and 1.6 centimeters. 29 lines have been identified, each corresponding to a different rotational quantum state. An accurate formula is given for the wave numbers of these lines. See also 2622 of 1946 and 3096 of November (Strandberg *et al.*).

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.72:621.396.822.029.62 3508
Solar Radio Noise: Part I—E. V. Appleton and J. S. Hey. (*Phil. Mag.*, vol. 37, pp. 73–84; February, 1946.) An account of experiments carried out during a period of sunspot activity. Ground-level measurements of the noise spectrum at meter wavelengths are described; the shape of the spectral curve at the longer wavelengths is deformed by ionospheric influences. Enhancement of the noise has been observed to occur simultaneously with solar flares and short-wave radio fadeouts. See also 402 of March (Appleton) and back references.

523.746:538.12 3509
The Growth and Decay of the Sunspot Magnetic Field—T. G. Cowling. (*Mon. Not. R. Astr. Soc.*, vol. 106, No. 3, pp. 218–224; 1946.) The rapid growth and decay, a matter of days, is thought to be due to convection and an initial magnetic field, otherwise the time taken would be about 300 years.

523. 746."1947.03/.04" 3510
Recent Solar Activity—(*Nature* (London), vol. 159, p. 549; April 19, 1947.) Report on the return of the sunspot group mentioned in 2760 of October. In April, it had a peak area of 5400 millionths of the sun's hemisphere, and there were no associated magnetic storms. See also 3107 of November.

523.854:[621.396.822.029.58/.6 3511
Interpretation of Radio Radiation from the Milky Way—C. H. Townes. (*Astrophys. Jour.*, vol. 105, pp. 235–240; March, 1947.) A discussion of the emission of radiation by ionized interstellar gas. Measurements between 30,000 and 9.5 Mc. are analyzed and compared with theory. The radiation is explicable on the basis of an electron gas density of about 1 electron per cubic cm. and a temperature of 100,000 to 200,000°K, which is much higher than that indicated previously. See also 402 of March, and 3598 and 3599 of January.

537.591 3512
Further Cosmic-Ray Experiments above the Atmosphere—S. E. Golian and E. H. Krause. (*Phys. Rev.*, vol. 71, pp. 918–919; June 15, 1947.)

537.591 3513
Slow Cosmic Ray Mesons at Sea-Level—G. R. Evans and T. C. Griffiths. (*Nature*, (London), vol. 159, pp. 879–800; June 28, 1947.)

537.591 3514
Recent Research in Meson Theory—G. Wentzel. (*Rev. Mod. Phys.*, vol. 19, pp. 1–18; January, 1947.)

537.591 3515
The Production of Nucleons by the Cosmic Radiation—S. A. Korff and B. Hamermesh. (*Phys. Rev.*, vol. 71, pp. 842–845; June 15, 1947.)

551.510.53 3516
The Temperature of the Upper Atmosphere—R. Penndorf. (*Bull. Amer. Met. Soc.*, vol. 27, pp. 331–342; June, 1946.) Translation of paper in *Met. Zeit.*, vol. 58, pp. 1–10; January, 1941. Summary noted in 3492 of 1942. A critical review of work published up to 1940. It is concluded that the probable thermal structure for latitudes 45 to 55 degrees may be represented approximately by the following points:—10 km., 220°K; 35 km., 230°K; 50 km., 320°K; 80 km., 200°K; 100 km., 330 to 370°K; 230 km. 430 to 830°K. Data for high latitudes are also discussed briefly.

551.510.53 3517
The Constitution of the Stratosphere—R. Penndorf. (*Bull. Amer. Met. Soc.*, vol. 27, pp. 343–345; June, 1946.) Translation of paper in *Met. Zeit.*, vol. 58, no. 3, pp. 103–105; 1941. Summary noted in 3492 of 1942.

The pressure/height relation used is $\log_e (p_2/p_1) = -A(h_2-h_1)/T$. Assuming the values of T as a function of height given in 3516 above, pressure is calculated in 1-km. steps up to 100 km. where the value agrees with that given by Martyn and Pulley (2073 of 1936) based on ionospheric data. The values of p are used to derive other parameters as a function of height and the results are tabulated for 10 km. intervals up to 100 km.

551.510.535 3518
Ionospheric Clouds—H. G. Wells. (*Tele-Tech*, vol. 6, pp. 53–54; May, 1947.) Summary of an Institute of Radio Engineers' paper. A description of a motion-picture pulse recording equipment.

551.510.535:621.396.11 3519
Radio Investigation of the Ionosphere—C. J. Bakker. (*Philips Tech. Rev.*, vol. 8, pp. 111–120; April, 1946.) A general survey of the physical constitution and properties of the ionosphere and their bearing on radio communication.

551.510.535:621.396.11 3520
The Role of the Ionosphere in the Propagation of Radio Waves—R. Jouaust. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 265–270; May, 1947.) Discussion on 1447 of June.

551.510.535:621.396.11 3521
Radiation Angle Variations from Ionosphere Measurements—Hallberg and Goldman. (*See* 3625.)

551.547+551.524.7 3522
Pressure and Temperature of the Atmosphere to 120 km.—N. Best, R. Havens, and H. La Gow. (*Phys. Rev.*, vol. 71, pp. 915–916; June 15, 1947.) Results of measurements made using a V-2 rocket and discussion of their accuracy.

LOCATION AND AIDS TO NAVIGATION

621.396.663:621.396.615.17 3523
On a Standardized Aperiodic Pulse Generator and Its Application to the Statistical Recording and Radiogoniometry of Atmospherics—Carbenay. (*See* 3471.)

621.396.93:519.2 3524
A Problem on the Summation of Simple Harmonic Functions of the Same Amplitude and Frequency but of Random Phase—Horner. (*See* 3566.)

621.396.93:551.594.6 3525
The Location of Thunderstorms by Radio Direction-Finding—F. Adcock and C. Clarke. (*Jour. I.E.E.* (London), Part I, vol. 94, p. 237; May, 1947.) Summary of 2799 of October.

621.396.93:621.396.677 3526
The Development and Study of a Practical Spaced-Loop Radio Direction-Finder for High Frequencies—W. Ross. (*Jour. Instn. Elec. Eng.*, Part I, vol. 94, p. 235; May, 1947.) Summary of 2780 of October.

621.396.93:621.396.677 3527
The Use of Earth Mats to Reduce the Polarization Error of U-Type Adcock Direction-Finders—R. L. Smith-Rose and W. Ross. (*Jour. I.E.E.* (London), Part I, vol. 94, p. 234; May, 1947.) Summary of 2781 of October.

621.396.93:621.396.677.029.58 3528
Site and Path Errors in Short-Wave Direction-Finding—W. Ross. (*Jour. I.E.E.* (London), Part I, vol. 94, p. 235; May, 1947.) Summary of 2782 of October.

621.396.93:621.396.677.029.62 3529
An Experimental Spaced-Loop Direction-Finder for Very-High Frequencies—F. Horner. (*Jour. I.E.E.* (London), Part I, vol. 94, p. 233; May, 1947.) Summary of 2783 of October.

621.396.93:621.396.677.029.63 3530
Some Experiments on Conducting Screens for a U-Type Spaced-Aerial Radio Direction-Finder in the Frequency Range 600–1200 Mc.—R. R. Pearce. (*Jour. I.E.E.* (London), Part I, vol. 94, p. 236; May, 1947.) Summary of 2784 of October.

621.396.932 3531
Radar for Merchant Marine Service—F. E. Spaulding, Jr. (*RCA Rev.*, vol. 8, pp. 312–330; June, 1947.) "Discusses the technical features of a new 3-cm. merchant marine radar equipment. Factors relating to the basic design are treated and operation of the various circuits is explained by reference to functional block diagrams. The physical form of the apparatus

is shown and plan-position-indicator (PPI) photographs are included to illustrate the navigational data furnished by this instrument. Specifications defining the performance characteristics are also included." See also 2194 of 1946 (Byrnes).

621.396.933 3532
P.I.C.A.O. Report on Navigational Aids [Book Notice]—Obtainable from E. M. Lewis, North Atlantic Regional Office, 7 Fitzwilliam Place, Dublin, 3s.9d.—(*Engineer* (London), vol. 183, p. 369; May 2, 1947.) Final report covering the first session in Montreal.

621.396.933.2 3533
The Theory and Application of the Radar Beacon—R. D. Hultgren and L. B. Hallman, Jr. (PROC. I.R.E., vol. 35, pp. 716–730; July, 1947.) The functions of the various components of a typical beacon and its applications.

621.396.933.2:621.396.615.141.2 3534
Stabilized Magnetron for Beacon Service: Part I—Development of Unstabilized Tube—Donal, Cuccia and Brown. (*See* 3736.)

621.396.933.2:621.396.615.141.2 3535
Stabilized Magnetron for Beacon Service: Part 2—Engineering of Tube and Stabilizer—Vogal and Dodds. (*See* 3737.)

621.396.96:621.317.79 3536
Echo Boxes for Radar Testing—Marshal. (*See* 3592.)

621.396.96:623.827 3537
Electronics in Submarine Warfare—C. A. Lockwood. (PROC. I.R.E., vol. 35, pp. 712–715; July, 1947.)

621.396.96(52) 3538
Short Survey of Japanese Radar: Part 1—R. I. Wilkinson. (*Bell Sys. Tech., Jour.*, vol. 26, p. 215; January, 1947.) Summary of part 1 of 424 of March.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.37 3539
Thermal and X-Ray Analyses of Some Common Phosphors—R. Nagy and Chung Kwai Lui. (*Jour. Opt. Soc. Amer.*, vol. 37, pp. 37–41; January, 1947.) The structural changes which occur in the formation of phosphors are explained and the correct firing temperatures for maximum fluorescence determined.

535.61-15:666.112.3:535.34 3540
Infrared Absorption Spectra of Some Experimental Glasses Containing Rare Earth and Other Oxides—R. Stair and C. A. Faick. (*Jour. Res. Nat. Bur. Stand.*, vol. 38, pp. 95–101; January, 1947.) Transmission data for soda lime glasses from 0.7 to 4.5 ohms.

535.61-15/-2:679.5 3541
Plastic Filters for the Visible and Near InfraRed Regions—J. H. Shenk, E. S. Hodge, R. J. Morris, E. E. Pickett, and W. R. Brode. (*Jour. Opt. Soc. Amer.*, vol. 36, pp. 569–575; October, 1946.) Discussion of the combination of dyes and plastics to give filters capable of resisting heat, intense light, and weather effects, and possessing specified transmission characteristics.

538.21:669.14–41 3542
Medium-Frequency Magnetization of Sheet Steel—R. Pohl. (*Jour. I.E.E.* (London), Part II, vol. 94, pp. 118–123; April, 1947.) Discusses the interdependence of hysteresis, eddy currents, and magnetic utilization; and gives simple expressions and curves for eddy-current loss, apparent flux and utilization factor. Summary *ibid.*, Part I, vol. 94, p. 278; July 1947. For earlier work see 2047 of 1945.

620.197 3543
Protective Finishing of Electrical Equipment—(*Engineering*, London, vol. 163, p. 330; April 25, 1947.) Summary of an Institute of Radio Engineers' paper by F. Widnall and R. Newbound. A general survey covering self-protective materials, electrolytic and chemical finishing, paint spraying, vitreous enameling, metal spraying, and test methods. For another account see *Elec. Times*, vol. 111, p. 246; March 6, 1947.

621.314.63 3544
Remarks on the Operation and Construction of Barrier Layer Rectifiers—M. Leblanc. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 202–208; April, 1947.) Discussion on 1468 of June.

621.314.63 3545
Applications of Dry Rectifiers—J. M. Girard. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 202–208; April, 1947.) Discussion on 1469 of June.

621.315.59+537.311.33 3546
Semiconductors and Their Applications—A. F. Ioffe. (*Bull. Acad. Sci.* (U.R.S.S.) sér. phys., vol. 10, no. 1, pp. 3–14; 1946. In Russian.) The work in progress at the Physico-Technical Institute of the Academy is surveyed under the following headings: (a) The determination of the number of the conductivity electrons and of their mobility; experimental curves for various types of semiconductors are shown. (b) Electrical conductivity in strong electric fields: deviations from Ohm's law are discussed. (c) The mechanism of conductivity: factors determining the direction of the current passed by a semiconductor are examined. (d) Boundary layers: the penetration into semiconductors of the field set up by the contact potential difference is considered. (e) Photoelectric phenomena: spectral sensitivity characteristics are plotted for various semiconductors. (f) Applications of semiconductors: particulars of the Ag_2S and Tl_2S photo cells of Soviet manufacture are given in Table 2.

621.315.61.011.5:546.431.823:537.228.1 3547
Dielectric and Piezoelectric Properties of Barium Titanate—S. Roberts. (*Phys. Rev.*, vol. 71, pp. 890–895; June 15, 1947.) Description and discussion of measurements of dielectric constant and loss at biasing field strengths from 0 to 5 mv./m., at temperatures from −50 to +135 degrees C and at frequencies from 0.1 to 25 Mc. The transverse and longitudinal piezoelectric effects have been measured directly.

621.315.611.011.5+537.226.3 3548
The Relation between the Power Factor and the Temperature Coefficient of the Dielectric Constant of Solid Dielectrics: Part 3—M. Gevers. (*Philips Res. Rep.*, vol. 1, pp. 298–313; August, 1946.) A new theory is presented which explains why, for most of the commercial dielectrics, the ratio between the temperature coefficient (T.C. of the dielectric constant and the power factor tan δ has a value of about 0.6. Hence the value of the T.C. can be predicted from measurements of tan δ at two different frequencies and two temperatures. For a mixture of dielectrics a simple linear relation is found to exist between its T.C. and those of the components. Thus a mixture of 90 per cent CeO_2 (T.C. $+100\times10^{-6}$) and 10 per cent TiO_2 (T.C. -880×10^{-6}) has a T.C. approximately zero. The linear relation does not apply to the power factors of the components of a composite dielectric and no general law relating T.C. and tan δ can be given for mixtures, particularly if some of the components have a positive and others a negative T.C. For part 4 see 3572 below.

621.315.611.011.5+537.226.3 3549
The Relation Between the Power Factor and the Temperature Coefficient of the Dielectric Constant of Solid Dielectrics: Part 4—Gevers. (*See* 3572).

621.315.612.2 3550
Alkaline Earth Porcelains Possessing Low Dielectric Loss—M. D. Rigterink and R. O. Grisdale. (*Jour. Amer. Ceram. Soc.*, vol. 30, pp. 78–81; March 1, 1947.) Porcelain bases for deposited carbon resistors, prepared from mixtures of clay, flint, and synthetic fluxes consisting of clay calcined with at least three alkaline earth oxides. These white porcelains have excellent dielectric properties and low coefficients of thermal expansion.

621.315.612.4.011.5 3551
Properties of Barium-Strontium Titanate Dielectrics—E. N. Bunting, G. R. Shelton, and A. S. Creamer. (*Jour. Res. Nat. Bur. Stand.*, vol. 38, pp. 337–349; March, 1947. *Jour. Amer. Ceram. Soc.*, vol. 30, pp. 114–125; April 1, 1947.) Results are given for various properties, including dielectric constant and power factor reciprocal for frequencies of 50 to 20,000 kc. together with some measurements at 3000 Mc.

621.316.89+621.315.59 3552
Properties and Uses of Thermistors—Thermally Sensitive Resistors—J. A. Becker, C. B. Green, and G. L. Pearson. (*Bell Sys. Tech. Jour.*, vol. 26, pp. 170–212; January, 1947.) Reprint of 765 of April.

621.318.2:621.775.7 3553
Sintered Permanent Magnets—S. J. Garvin. (*Engineering* (London), vol. 163, pp. 445–446 and 465–467; May 30, and June 6, 1947. A review of the development of sintering and a detailed account of recent methods for producing accurately shaped magnets of alnico or alcomax. Such methods involve the use of a "master alloy" of 48 per cent Fe, 52 per cent Al, which has a wetting point about 100 degrees C below the sintering temperature. This alloy is brittle and can be crushed readily to a fine powder. It is also much less prone to oxidation than pure Al, so that commercial hydrogen can be used as the atmosphere during the sintering process.

621.318.22:669.144.25 3554
Vicalloy—A Workable Alloy for Permanent Magnets—G.W.O.H. (*Wireless Eng.*, vol. 24, p. 192; July, 1947.) Editorial comment on a Bell Telephone System monograph by E. A. Nesbitt. Vicalloy is a new alloy of Fe, Co, and Va, which can be rolled and drawn and has been used as a tape 0.05 by 0.002 inch for speech recording in the Western Electric mirrorphone.

621.318.323.2.042.15 3555
Permeability of Dust Cores—V. E. Legg. (*Wireless Eng.*, vol. 24, pp. 218–219; July, 1947.) Comment on 35 of February. The value of an empirical formula for the permeability of molybdenum permalloy cores given in 4424 of 1940 (Legg & Given) is discussed. An empirical treatment is stated to be more profitable than a mathematical analysis of the magnetic behavior of such cores, as the shape and size of the magnetic particles and their relative dispositions in the insulating material are not simple and depend on the grain structure of the permalloy as originally cast and on its subsequent treatment. See also 1692 and 1693 of July and 2816 of October.

666.2:621.327.3 3556
Ultraviolet-Transmitting Glasses for Mercury-Vapor Lamps—M. E. Nordberg. (*Jour. Amer. Ceram. Soc.*, vol. 30, pp. 174–179; June 1, 1947.) The ultraviolet transmitting properties of Vycor glasses no. 791 and no. 7911 are compared with those of certain other glasses and fuzed silica. Transmission loss with age is much less than with other glasses.

678+546.26]:621.317.331 3557
Electrical Conductivity of GR-S and Natural Rubber Stocks Loaded with Shawinigan and R-40 Blacks—P. E. Wack, R. L. Anthony,

and E. Guth. (*Jour. Appl. Phys.*, vol. 18, pp. 456–469; May, 1947.)

679.5 3558

A Plastics Primer for Engineers—K. Rose. (*Materials and Methods*, vol. 25, pp. 119–138; April, 1947.) Description and characteristics are given for thermosetting resins of the phenolic, amino-formaldehyde, aniline-formaldehyde, and allyl ester groups and the effects of various fillers upon their properties are discussed. Thermoplastic groups include cellulosics, vinyls, acrylics, polyamides, polystyrenes, polyethylene, polytetrafluorethylene, caseins, and silicones. The trade names by which the principal plastics are known in the United States are tabulated.

679.5:621.315.616 3559

Teflon—An Improved Plastic for R.F. Use—W. S. Penn. (*Electronic Eng.* (London), vol. 19, p. 220; July, 1947.) Electrical, mechanical and dielectric properties of polytetrafluorethylene (Teflon) and polythene are compraed. See also 1121 of May and 3169 of November (Johnson).

MATHEMATICS

513.732.6:621.396.615.141.2 3560

A Flux Plotting Method for Obtaining Fields Satisfying Maxwell's Equations, with Applications to the Magnetron—P. D. Crout. (*Jour. Appl. Phys.*, vol. 18, pp. 348–355; April, 1947.) Flux plotting methods previously applied to fields satisfying Laplace's and Poisson's equations are here extended to fields satisfying Maxwell's equations. The method is applied to the hole-and-slot type of magnetron operating in its main mode, and to the vane type of magnetron. For previous work see Radiation Laboratory Report 1047.

517.512.2:621.396.67 3561

Fourier Transforms in Aerial Theory: Part 2—J. F. Ramsay. (*Marconi Rev.*, vol. 10, pp. 17–22; January to March, 1947.) Graphs of the Fourier sine transforms of a square wave, a saw-tooth wave, a sine wave, and a sine-squared wave are given. For part 1 see 2680 of October.

518.5 3562

An Electrical Network for the Solution of Secular Equations—R. H. Hughes and E. B. Wilson, Jr. (*Rev. Sci. Instr.*, vol. 18, pp. 103–108; February, 1947.) A network of suitable coils and capacitors is assembled whose resonance equation is identical with the secular determinant. For determining the latent roots of a real symmetric matrix of n rows and columns, the network has n junctions which are interconnected by reactive admittances and grounded by equal variable admittances which represent the unknown in the equation. The values of the variable admittances corresponding to maxima of the voltages of the junctions with respect to ground give the desired roots. Usually an accuracy better than 1 per cent is obtained, for $n \leqq 6$.

518.5 3563

Electrical Analogue Computing: Parts 1 and 2—D. J. Mynall. (*Electronic Eng.*, vol. 19, pp. 178–180 and 214–216; June and July, 1947.) The fundamental circuits used for electro-mechanical addition, multiplication, division, differentiation and integration are described. To be continued.

518.5:621.385 3564

Electrostatic Storage—Rajachman. (*See* 3712.)

518.61 3565

Some Improvements in the Use of Relaxation Methods for the Solution of Ordinary and Partial Differential Equations—L. Fox. (*Proc. Roy. Soc. A.*, vol. 190, pp. 31–59; June 17, 1947.) The standard use of relaxation methods is extended by the inclusion of terms usually neglected in the finite difference equations involved. Eight examples of the method are given, illustrating the high accuracy obtainable with reduced labor.

519.2:621.396.93 3566

A Problem on the Summation of Simple Harmonic Functions of the Same Amplitude and Frequency but of Random Phase—F. Horner. (*Phil. Mag.*, vol. 37, pp. 145–162; March, 1946.) The problem treated is the determination of the probability $P_n(s)$ that the amplitude of an arbitrarily chosen component of the resultant shall lie between the limits s and $s+ds$. Curves of $P_n(s)$ are given as functions of s for $n=1$, 2, 3, and 7 where n is the number of harmonic functions involved in the summation. For large values of n, the distribution is of Gaussian form, and it seems likely that the lowest value of n for which the Gaussian curve gives a reasonably good fit is 5. The r.m.s. values of s are the same for the true and normal distributions for all values of n.

MEASUREMENTS AND TEST GEAR

531.76:621.317.755 3567

Precision Device for Measurement of Pulse Width and Pulse Slope—H. L. Morrison. (*RCA Rev.*, vol. 8, pp. 276–288; June, 1947.) Direct measurement in microseconds for pulses having the same repetition rate.

531.761:621.317.39 3568

An Electronic Millisecond Timer—S. S. West and L. C. Bentley. (*Electronic Eng.*, vol. 19, pp. 207–210; July, 1947.) The circuit comprises an electronic trigger arrangement, which can be operated by a photo cell or from an external source.

It will measure short time intervals in the range 0.5 to 100 milliseconds with an accuracy of ± 2 per cent. The accuracy depends almost entirely on the stability of a standard capacitor and on the calibrating resistance.

531.765:621.317.755 3569

Spiral Chronograph for Measurement of Single Millisecond Time Intervals with Microsecond Accuracy—R. J. Emrich. (*Rev. Sci. Instr.*, vol. 18, pp. 150–157; March, 1947.) The spiral timebase is controlled by a crystal; the c.r.t. beam is held off in the steady state and is switched on by the pulse marking the beginning of the time interval to be measured. The beam brilliance is modulated to provide 5-microsecond markers and is turned off by the pulse marking the end of the required time interval, an Eccles-Jordan trigger circuit being used. The screen is of long persistence, the trace being measured directly, or photographed as a "still" record. A pulse sharpening circuit is used to eliminate uncertainty as to the exact time of operation of the on-off trigger.

538.569.4.029.64:546.171.1 3570

The Inversion Spectrum of Ammonia at Centimetre Wavelengths—Bleaney and Penrose. (*See* 3507.)

621.3.012.3 3571

Microwave Impedance-Plotting Device—W. Altar and J. W. Coltman. (Proc. I.R.E., vol. 35, pp. 734–737; July, 1947.) The device is used in conjunction with a Smith Impedance Chart (1372 of 1939) and computes the angular position of the load point directly from the observed data.

621.315.611.011.5+537.226.3 3572

The Relation Between the Power Factor and the Temperature Coefficient of the Dielectric Constant of Solid Dielectrics: Part 4—M. Gevers. (*Philips Res. Rep.*, vol. 1, pp. 361–379; November, 1946.) The power factor is measured by determining the increase in damping of a tuned circuit when a capacitor formed from the dielectric is connected in parallel. The details of this method and the apparatus used for determining the temperature coefficient of the dielectric constant are discussed. Sources of error and the necessary corrections are indicated. For previous parts see 125 of February 1476 of June and 3548 above.

621.317.1.011.5 3573

A Method for Measuring Certain Electric Constants at Centimetre Wavelengths—K. G. Knorre. (*Bull. Acad. Sci.* (U.R.S.S.), sér. phys., vol. 10, no. 1, pp. 117–123; 1946. In Russian.) A rectangular cavity resonator is considered with ideally conducting walls and divided into three zones each representing a different dielectric medium (Fig. 1). The discussion is limited to H-waves and systems of (3) and (7) are derived determining the field in each zone. On the basis of the results obtained, a method is proposed for measuring the dielectric constant of a medium. The method is based on obtaining resonance by moving one of the end walls of the resonator. The damping of a resonator containing a dielectric is discussed and also the possibility of measuring losses in the dielectric.

621.317.3:621.396.611.21 3574

Electrical Characteristics of Quartz-Crystal Units and Their Measurement—W. D. George, M. C. Selby, and R. Scolnik. (*Jour. Res. Nat. Bur. Stand.*, vol. 38, pp. 309–328; March, 1947.) Q-meters and r.f. bridges were used. Measurement methods and their relative merits and limitations are discussed. Antiresonance impedance up to 5 megaohms was measured to within ± 5 per cent. Constancy of electrical characteristics, secondary responses and changes with amplitude of vibration and temperature were investigated for many 8.7-Mc. BT-cut crytsal units and a few 50-kc. and 100-kc. units. A graphical method of representing the electrical characteristics of normal crystal units is suggested.

621.317.333.82:621.319.4 3575

Overvoltage Testing of Capacitors—R. J. Hopkins. (*Electronics*, vol. 20, pp. 105–107; June, 1947.)

621.317.336.1:[621.385.3/.5 3576

The Measurement of Dynamic Mutual Conductance of Valves using the Grounded-Grid Triode Mode of Operation—F. Gutmann. (*Jour. Sci. Instr.*, vol. 24, pp. 94–95; April, 1947.) The tube under test is connected as a cathode-loaded grounded-grid triode and a known alternating voltage in series with a current indicator is applied across the load. The measured current is approximately proportional to the dynamic mutual conductance.

621.317.361 3577

F.C.C. Frequency Measurement Techniques—A. K. Robinson. (*Electronics*, vol. 20, pp. 114–116; June, 1947.) The system depends on a primary 50-kc. standard, of accuracy greater than 1 in 10^7, with frequency subdivision to 50 c.p.s. for comparison check with standard time signals. 10-kc. markers, derived from the standard, extend up to 500 Mc. by use of a high-gain harmonic amplifier having individual harmonic output constant with frequency. The external signal to be checked is made to beat with the nearest marker. The beat note is measured by an audio interpolation oscillator, range 0 to 5 kc., accuracy ± 2 c.p.s., used with an oscilloscope. Provision is made for sense determination. Signals of short duration are checked by means of a heterodyne frequency meter.

621.317.382:621.385.831 3578

Power Measurement of Class B Audio Amplifier Tubes—D. P. Heacock. (*RCA Rev.*, vol. 8, pp. 147–157; March, 1947.) An accurate method, particularly suitable for production tests, of determining the performance by a simple measurement of the anode current of one of the two tubes.

621.317.44.025 3579

An Alternating Current Probe for Measurement of Magnetic Fields—E. C. Gregg, Jr. (*Rev. Sci. Instr.*, vol. 18, pp. 77–80; February, 1947.) A summarized account of this probe was noted in 3184 of November. Advantage is taken of the fact that the a.c. permeability of most magnetic alloys changes with superposed steady-state d.c. magnetic fields.

621.317.49:534.321.9 3580

Comparison of Supersonic Intensities by Means of a Magnetostriction Gauge—A. W. Smith and D. K. Weimer. (*Rev. Sci. Instr.*, vol. 18, pp. 188–190; March, 1947.) The gauge consists essentially of a small rod of Ni wound with a few turns of wire. The acoustic pressures present in the liquid produce changes in length of the rod which in turn, because of the inverse magnetostrictive effect, induce voltages in the coil.

621.317.7+621.38+621.396.69 3581

The Physical Society's Exhibition—(*Engineer* (London), vol. 183, pp. 328–331, 352–353, and 383–385; April 18 and 25, and May 2, 1947.) *Engineering* (London), vol. 163, pp. 364–366; May 2, 1947.) Descriptions of further selections of the exhibits in the trade and research sections. See also 3185 of November and 2494 of September.

621.317.7.029.62/.63:[621.396.81+621.396.822 3582

A V.H.F./U.H.F. Noise and Field Intensity Meter—L. W. Martin. (*Communications*, vol. 27, pp. 32–35, 44; June, 1947.) Description, with complete circuit diagram, of equipment for noise measurement in mv. or db. and field-intensity measurement in mv./m. in the frequency range 88 to 400 Mc.

621.317.725 3583

A Very High Impedance R.M.S. Voltmeter for Iron Testing—K. A. Macfadyen, D. C. Gall and F. C. Widdis. (*Jour. Sci. Instr.*, vol. 24, p. 109; April, 1947.) Discussion of 1514 of June. Input impedance up to 80 megaohms is achieved by returning the grid leak to a positive voltage. The relative merits of current and voltage feedback are considered.

621.317.725.027.7 3584

The Design of an Ellipsoid Voltmeter for the Precision Measurement of High Alternating Voltages—F. M. Bruce. (*Jour. I. E. E.* (London), Part II, vol. 94, pp. 129–137; Discussion, pp. 149–154; April, 1947.) High alternating voltages are measured by timing the oscillations of a small ellipsoid suspended by a thread in the uniform electric field between two parallel vertical disks. By accurate mechanical construction, and correction of the results for known sources of error, the voltage between the disks is deduced with an estimated error of less than ± 0.03 per cent. See also 3585 below. Summary, *ibid.*, Part I, vol. 94, p. 279; June, 1947.

621.317.728.089.6 3585

Calibration of Uniform-Field Spark-Gaps for High-Voltage Measurement at Power Frequencies—F. M. Bruce. (*Jour. I.E.E.* (London), Part II, vol. 94, pp. 138–149; April, 1947. Discussion, pp. 149–154.) Description of spark gaps using electrodes shaped to give a uniform field in the gap. A calibration between 9 and 315 kv. is given, agreeing with a simple empirical formula to within ± 0.2 per cent. See also 3584 above. Summary *ibid.*, Part I, vol. 94, pp. 279–280; June, 1947.

621.317.75 3586

A Rotary Periodograph—G. B. Moncrieff-Yeates. (*Jour. Sci. Instr.*, vol. 24, pp. 35–40; February, 1947.) An instrument for the rapid analysis of disturbed periodic functions. The variance, obtained photoelectrically, of the sum of two ordinates is plotted against their separation to produce a "characteristic curve" containing many of the constants required.

621.317.755:621.3.015.3:621.311.1 3587

Oscillographs for Rapid Transient Phenomena. Their Application to the Study of Overvoltages in Grid Systems—P. Grassot. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 95–101; February, 1947.) Short descriptions of various modern instruments, with applications to surge investigations.

621.317.761+[621.396.615.12:621.317.79 3588

A V.H.F. Signal Generator or Frequency Meter—J. G. Ratcliff. (*R.S.G.B. Bull.*, vol. 22, pp. 118–122; February, 1947.) Describes a compact heterodyne frequency meter, covering the range 5 to 250 Mc., which can also be used as a tone-modulated source of r.f. voltage. Six plug-in coils are used and crystal check points are provided at 2- and 10-Mc. intervals. Transitron multivibrators give division down to 500 or 100 kc.

621.317.763.029.64 3589

Direct Reading Wavemeters—G. E. Feiker and H. R. Meahl. (*Tele-Tech.*, vol. 6, p. 59; May, 1947.) Summary of an Institute of Radio Engineers' paper. An account of loop-coupled quarter-wave coaxial wavemeters for use in the 8 to 12 and 12 to 17 cm. wavelength ranges. Resonance is indicated by a crystal-tube voltmeter. Accuracy is within 0.1 per cent.

621.317.772.029.64:621.396.67 3590

Phase-Front Plotter for Centimeter Waves—H. Iams. (*RCA Rev.*, vol. 8, pp. 270–275; June, 1947.) The area to be plotted is scanned by a motor-actuated probe. The energy picked up by the probe at any point is combined with a reference signal and applied to a detector. The detector output, which varies with the phase difference between the probe and reference signals, is amplified and applied to a stylus directly below the probe; a sheet of current-sensitive paper is thus darkened in proportion to the detector output and a record is obtained showing which parts of the area have the same phase. The plotter was used to test centimeter-wave aerials.

621.317.79:621.396.822 3591

Distortion-Noise Meter—C. W. Clapp. (*Tele-Tech.*, vol. 6, p. 61; May, 1947.) Summary of an Institute of Radio Engineers' paper. A bridge-T type filter covering 50 to 15,000 c.p.s. rejects the fundamental while passing harmonics. The circuit can be adapted for use as a noise meter.

621.317.79:621.396.96 3592

Echo Boxes for Radar Testing—R. W. Marshall. (*Bell Lab. Rec.*, vol. 25, pp. 111–113; March, 1947.) A short account of typical constructions and their use to indicate over-all performance of radar installations.

621.396.619.083:621.397.5 3593

A Method of Measuring the Degree of Modulation of a Television Signal—T. J. Buzalski. (*RCA Rev.*, vol. 7, pp. 265–271; June, 1946.) The double sideband output of the transmitter energizes a linear diode monitor, the output of which contains a d.c. component in addition to the visual signal. This composite signal is short-circuited periodically by a "Vibroswitch," thus establishing a zero level. The amplitudes of the components of the resultant trace may be read directly on an oscilloscope and the modulation percentage calculated.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.321.9:531.717:621.436–222 3594

Ultrasonic Measurement of Wall Thickness in Diesel Cylinder Liners—F. W. Struthers and H. M. Trent. (*Jour. Acous. Soc. Amer.*, vol. 19, pp. 368–371; March, 1947.

535.61–15:621.389 3595

Infrared Equipment for Military Purposes—(*Engineering* (London), vol. 163, p. 258; April 4, 1947.) For another account see 2862 of October.

535.61–15:621.391.64:621.327.032.196 3596

Cesium Vapor Lamps—N. C. Beese. (*Jour. Opt. Soc. Amer.*, vol. 36, pp. 555–560; October, 1946.) Structural details and characteristics of lamps giving infrared radiation and capable of 100 per cent current modulation throughout the a.f. range.

537.533.73:539.2 3597

Electron Diffraction. Apparatus used in France and Abroad. Possibilities of the Method for Crystal Analysis of Thin Plates and for Surface Structure—J. Devaux. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 111–115; February, 1947.)

538.71.001.8+538.71:[623.26+623.95 3598

Bomb and Mine Location: Peace-Time Applications—(*Beama Jour.*, vol. 54, pp. 139–140; April, 1947.) A short account of the E.R.A. mumetal magnetometer and balanced-coil locator, with applications to the locations of sunken anchor buoys and other equipment in the Seine estuary.

550.837:621.396.675 3599

Electric Field of an Oscillating Dipole on the Surface of a Two-Layer Earth—Wolf: Lewis. (*See* 3435 and 3436.)

550.837.7 3600

On the Use of Electromagnetic Waves in Geophysical Prospecting—C. W. Horton. (*Geophys.*, vol. 11, pp. 505–517; October, 1946.) The response of the earth to a d.c. step function is analyzed for the case in which displacement currents are negligible. It is shown that under typical conditions the depth of an electrical interface 6000 feet below ground can be measured by means of e.m. waves, even thin layers of salt water or oil-bearing sand giving measurable effects.

550.837.7:621.396.9 3601

Use of the Broadcast Band in Geologic Mapping—L. Kerwin. (*Jour. Appl. Phys.*, vol. 18, pp. 407–413; April, 1947.) A description of field equipment designed to study the effect of geologic anomalies on e.m. field intensity, with experimental results.

551.46.018.3:621.317.39 3602

The Measurement of Sea-Water Velocities by Electromagnetic Induction—R. W. Guelke and C. A. Schoute-Vanneck. (*Jour. I.E.E.* (London), Part I, vol. 94, p. 232; May, 1947.)

620.191.33:534.321.9 3603

The Detection of Internal Leaks in Aircraft Hydraulic Systems—R. G. Nuckolls and H. M. Trent. (*Jour. Acous. Soc. Amer.*, vol. 19, pp. 364–367; March, 1947.) A crystal pickup and amplifier which monitor "the ultrasonic vibrations produced in the system by the leaking tube, which vibrations are most intense at the defective element."

621.318.572 3604

Vane-Actuated Controller—Wannamaker. (*See* 3444.)

621.365.5 3605

Temperature Charts for Induction and Constant-Temperature Heating—M. P. Heisler. (*Trans. A.S.M.E.*, vol. 69, pp. 227–236; April, 1947.) "Charts are presented for determining complete temperature histories in spheres, cylinders, and plates."

621.38/.39].001.8 3606

Radar Techniques in an Industrial Control—W. D. Cockrell. (*Elec. Eng.*, vol. 66, pp. 365–

368; April, 1947.) A description of u.h.f. methods for register control in the printing and paper industries.

621.38:6(048) 3607
Industrial Electronic Equipment Uses: Part 2—W. C. White. (*Elec. Ind.*, vol. 1, p. 6; April, 1947.) Continuation of 2520 of September. A further list of 123 references.

621.384.6 3608
Atomic Artillery—J. Stokley. (*Gen. Elec. Rev.*, vol. 50, pp. 9–19; June, 1947.) An outline of the various types of electron and ion accelerators which have been used to produce streams of atomic particles of high energy. The principles of operation of the cyclotron, synchro-cyclotron, betatron, and synchrotron are described, and mention is made of a new type of linear accelerator expected to produce particles with an energy of 40 megaelectron volts, in which radar pulse transmitters will provide the energy source.

621.384.6:621.316.7 3609
The Synchronization of Auxiliary Apparatus with a Betatron—G. C. Baldwin, G. S. Klaiber, and A. J. Hartzler. (*Rev. Sci. Instr.*, vol. 18, pp. 121–124; February, 1947.) Automatic control by external apparatus, such as a cloud chamber, of the production of X-rays by a betatron is described. Upon receiving an initiating signal from the cloud chamber a relay and thyratron circuit permits injection of electrons into the betatron vacuum tube only during a single cycle. Three thyratrons furnish a series of synchronizing signals.

621.385.833 3610
Present Status and Future Possibilities of the Electron Microscope—J. Hillier. (*RCA Rev.*, vol. 8, pp. 29–42; March, 1947.)

621.385.833 3611
The Electron Microscope—P. Grivet. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 102–110; February, 1947.) A description of some of the special features of the C.S.F. electrostatic instrument, with microphotographs of widely differing objects. See also 3706 of January.

621.385.833 3612
The Electron Optical System of the Electron Microscope—M. E. Haine. (*Jour. Sci. Instr.*, vol. 24, pp. 61–66; March, 1947.) Theoretical and practical considerations in the design and use of the microscope.

621.385.833 3613
On the Limit of Resolution of the Electron Microscope. Unsymmetrical Lens—H. Bruck. (*Compt. Rend. Acad. Sci.* (Paris), vol. 224, pp. 1628–1629; June 9, 1947.) Formulas for the limit are derived, which depend on the lack of symmetry in the objective lens. Similar formulas are given for the optical case. The formulas hold in all cases where lack of symmetry is a more serious defect than spherical aberration. See also 3238 of November.

621.385.833 3614
Conditions for Extending the Resolution Limit of the Electron Microscope—V. E. Cosslett. (*Jour. Sci. Instr.*, vol. 24, pp. 40–43; February, 1947.) The limiting resolution obtainable with magnetic lenses of existing type may be reduced from 10 Å to perhaps 6 Å by the use of a sufficiently high accelerating voltage, provided that the lens power is maintained at the value which gives minimum aberration. Further improvement can only be obtained by correction of lens aberrations.

621.385.833 3615
Preparation and Uses of Silica Replicas in Electron Microscopy—C. H. Gerould. (*Jour. Appl. Phys.*, vol. 18, pp. 333–343; April, 1947.) A method is described for preparing silica replicas of specimens which cannot be subjected to the temperatures and pressures of the ordinary technique.

621.386.1 3616
X-Ray Generators for 1000 and 2000 kv.—J. Saget. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 273–274; May, 1947.) Discussion on 1547 of June.

621.386.1:615.849 3617
A 400 Kilovolt Installation for X-Ray Therapy—W. H. Boldingh and W. J. Oosterkamp. (*Philips Tech. Rev.*, vol. 8, pp. 105–110; April, 1946.) A novel construction, with the anode earthed and the focus at the end of a long earthed metal tube projecting through a partition into the irradiation chamber.

621.791.736.31 3618
Precision Energy-Storage Spot Welder—R. Briggs and H. Klemperer. (*Electronics*, vol. 20, pp. 102–104; June, 1947.)

623.454.25:621.396.9 3619
Radio Proximity Fuze—(*Tech. Bull. Nat. Bur. Stand.*, vol. 31, pp. 3–8; January, 1947.) An account of the development of the fuse at the National Bureau of Standards, Washington. See also 623, 624, and 1627 of 1946.

PROPAGATION OF WAVES

538.566+621.396.11 3620
On the Propagation of Electromagnetic Waves Through the Atmosphere—B. K. Banerjea. (*Proc. Roy. Soc. A*, vol. 190, pp. 67–81; June 17, 1947.) "A general method of tackling the problem of the propagation of electromagnetic waves in the ionosphere has been developed and the current methods of Appleton, Hartree, Saha, Rai, and Mathur, etc., have been deduced as special cases from the general results. The different assumptions by Appleton, Hartree, Bose, Booker, and Rai, as regards the condition of reflexion of the waves from the ionosphere, have been shown to be identical. A symbol-correspondence chart for the different symbols used by the different workers has been given to facilitate the understanding of the parallelism between the different methods. Polarization of the radio waves has been discussed fully."

538.566 3621
[One-Dimensional] Propagation of a Perturbation, of Narrow Frequency Range, in a Non-Absorbing Dispersive Medium—Blanc-Lapierre and Lapostolle. (*See* 3505.)

551.510.535:621.396.24 3622
Application of the Theories of Indirect Propagation to the Calculation of Links Using Decametre Waves—Aubert. (*See* 3654.)

621.396.11 3623
On the Problem of Efficient Long-Distance Wireless Power Transmission—S. Tetelbaum (*Jour. Phys.* (U.S.S.R.), vol. 9, no. 6, pp. 505–514; 1945.)

621.396.11:534.231 3624
A Device for Plotting Rays in a Stratified Medium—Lawson, Miller, Jr., and Schiff. (*See* 3389.)

621.396.11:551.510.535 3625
Radiation Angle Variations from Ionosphere Measurements—H. E. Hallborg and S. Goldman. (*RCA Rev.*, vol. 8, pp. 342–351; June, 1947.) "The heights of the F and F_2 layers at Washington, D. C., and San Francisco, Calif., and their variability ranges are studied for the year 1945. These data are applied to determine the optimum radiation angle ranges for various hop modes on the New York-San Francisco circuit. Wide diurnal and seasonal variations are indicated. Practical applications to effective antenna design are discussed."

621.396.11:551.510.535 3626
The Role of the Ionosphere in the Propagation of Radio Waves—Jouaust. (*See* 3520.)

621.396.11:551.510.535 3627
Radio Investigation of the Ionosphere—Bakker. (*See* 3519.)

621.396.11.029.62/.63 3628
Propagation Studies on 45.1, 474, and 2800 Megacycles Within and Beyond the Horizon—G. S. Wickizer and A. M. Braaten. (Proc. I.R.E., vol. 35, pp. 670–680; July, 1947.) Recordings of field strength on 2800, 474, and 45.1 Mc. over a period of 13 months were made at distances of 42 miles (within the horizon) and 70 miles (beyond the horizon) from the transmitter. Maximum values 3 or 4 times the free-space values were obtained at the two higher frequencies. Variations at 474 and 2800 Mc. were greater than those at 45.1 Mc.; variations at 70 miles were greater than those at 42 miles. Refraction was found to be greater in summer; superrefraction only occurred when the wind velocity was less than about 13 m.p.h. Simultaneous meteorological observations were made.

621.396.41.029.64 3629
Calculation of Multiplex U.H.F. Radio-Telephone Links—H. Chireix. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 271–272; May, 1947.) Discussion on 1559 of June.

621.396.81 3630
V.H.F. Propagation Surveys for Mobile Services—R. G. Peters. (*Communications*, vol. 27, pp. 20 and 45; June, 1947.)

621.396.812.029.62 3631
Propagation on Five—E. J. Williams and D. W. Heightman. (*Short Wave Mag.*, vol. 4, pp. 749–751; February, 1947.) Criticism of 1561 of June; for Russell's reply see 3632 below.

621.396.812.029.62 3632
More About V.H.F. Propagation—O. J. Russell. (*Short Wave Mag.*, vol. 5, pp. 46–48; March, 1947.) A reply to criticism in 3631 above of 1561 of June.

621.396.812.029.64 3633
Research in England on the Propagation of Ultra-Short Waves—Bras. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 270–271; May, 1947.) Discussion on 1563 of June.

621.396.812.4.029.62 3634
Tropospheric Reception—G. W. Pickard and H. T. Stetson. (*Tele-tech*, vol. 6, p. 54; May, 1947.) Summary of an Institute of Radio Engineers' paper. Daily records of field strength of W2XMN f.m. transmissions on 42.8 Mc. show variations dependent upon the passage of warm and cold fronts across the transmission path. Reception at 167-mile range was, on the average, three to four times stronger in summer than in winter.

RECEPTION

621.396.621 3635
Modernizing the Old Receiver—W. L. North. (*QST*, vol. 31, pp. 54–55, 130; April, 1947.) Details of alterations to an RME-69 receiver resulting in considerable improvement in both gain and image rejection.

621.396.621 3636
Criteria for Diversity Receiver Design—W. Lyons. (*RCA Rev.*, vol. 8, pp. 373–378; June, 1947.) Discussion limited to receivers incorporating diode switching of the common diode load variety.

621.396.621:621.395.623.66 3637
The Pocket Ear—J. L. Hathaway and W. Hotine. (*RCA Rev.*, vol. 8, pp. 139–146; March, 1947.) The development of a three-tube pocket radio receiver, with a flexible tube for conducting sound to the ear, for maintaining contact between a program producer and a roving announcer.

621.396.621:621.396.619.11 3638
The Synchrodyne—F. M. Apthorpe, and D. G. Tucker. (*Elec. Eng.*, vol. 19, p. 238; July, 1947.) Comment on 2364 of September. Comparison is made with the "Homodyne" (F. M. Colebrook, *Wireless World and Radio Rev.*, vol. 13, pp. 645–648; 1924.) Apthorpe considers distortion of a signal subject to selective fading by reference to vector diagrams and discusses the advantages of harmonic synchronization, and methods for avoiding the howl when off tune. Tucker stresses the essential difference between the homodyne and the synchrodyne, pointing out that whereas the homodyne gives selectivity in preference to quality, in the synchrodyne selectivity and quality are quite independent and both may be excellent.

621.396.621.001.4:621.396.82 3639
Static for Radio Receiver Tests—J. C. R. Licklider and E. B. Newman. (*Electronics*, vol. 20, pp. 98–101; June, 1947.) Apparatus for artificial production of "atmospherics."

621.396/.397.621.004.67 3640
The Servicing of Radio and Television Receivers—R. C. G. Williams. (*Jour. I.E.E.* (London), Part 1, vol. 94, pp. 156–158; March, 1947.) Summary of 2224 of August.

621.396.621.5.029.62 3641
R.F./Mixer Design for V.H.F.—W. J. Crawley. (*Short Wave Mag.*, vol. 5, pp. 44–46; March, 1947.) Describes the development of a circuit for 58 Mc. with two high-gain r.f. stages; it uses a low-noise h.f. pentode and split-stator tuning capacitors.

621.396.621.54 3642
The Inversion of the Autodyne Principle—Saic. (*See* 3472.)

621.396.622.71 3643
The Ratio Detector—S. W. Seeley and J. Avine. (*RCA Rev.*, vol. 8, pp. 201–236; June, 1947.) "In this circuit two frequency-sensitive voltages are applied to diodes and the sum of the rectified voltages held constant. The difference voltage then constitutes the desired a.f. signal. This means of operation makes the output insensitive to amplitude variations.

" . . . The ratio between the primary and secondary components of the frequency-sensitive voltages in a phase-shift type of ratio detector is a function of the instantaneous signal amplitude. The a.m. rejection properties, however, are shown to depend upon the mean ratio between these voltages. An expression which is developed for this ratio in terms of the circuit parameters provides the basis for arriving at an optimum design. The measurements necessary in the design of a ratio detector and in checking its performance are described."

621.396.812.4.029.62 3644
Tropospheric Reception—Pickard and Stetson. (*See* 3634.)

621.396.822:621.396.621 3645
On the Theory of Noise in Radio Receivers with Square Law Detectors—M. Kac and A. J. F. Siegert. (*Jour. Appl. Phys.*, vol. 18, pp. 383–397; April, 1947.) "For the video output V of a receiver, consisting of an i.f. stage, a quadratic detector, and a video amplifier, the probability density $P(V)$ has been obtained for noise alone and for noise and signal. The results are expressed in terms of eigenvalues and eigenfunctions of the integral equation

$$\int_0^\infty K(t)\rho(s-t)f(t)dt = \lambda f(s),$$

where $\rho(\tau)$ is the i.f. correlation function (i.e., the Fourier transform of the i.f. power spectrum) and $K(t)$ is the response function of the video amplifier (i.e., the Fourier transform of the video amplitude spectrum). Two special cases are discussed in which the integral equation can be solved explicitly. Approximations for general amplifiers are given in the limiting cases of wide and narrow videos." Summary abstracted in 1199 of May.

621.396.822.029.6:621.385.2 3646
A Coaxial-Line Diode Noise Source for U.H.F.—Johnson. (*See* 3722.)

621.396.828 3647
A New Noise-Reducing System for C. W. Reception—D. L. Hings. (*QST*, vol. 31, pp. 21–23, 134; June, 1947.) Full details of a practical circuit for application to the second detector and a.f. end of a communicatinos receiver. See also 1576 of April and 3649 below.

621.396.828 3648
A Method for Preventing Impulse Interference with Radio Reception—A. N. Shchukin. (*Bull. Acad. Sci.* (U.R.S.S.), sér. phys., vol. 10, no. 1, pp. 49–56; 1946. In Russian.) The receiver is assumed to consist of a wide-band unit, followed by an amplitude limiter which in turn is followed by a narrow-band unit. The operation of the system is considered when one or more impulses are received in the presence or absence of the desired signal. Formulas are derived determining the ratio of the interference voltage at the output of the receiver to the corresponding useful signal voltage if there were no interference. The interference from another radio station operating at a frequency lying within the wide but outside the narrow band is also considered.

621.396.828:621.394.141 3649
Noise-Free Code Reception—D. L. Hings. (*Electronics*, vol. 20, pp. 125–127; June, 1947.) A method for discriminating between the time constants of signal and noise, allowing c.w. signals to trigger an a.f. generator feeding a loudspeaker. See also 1576 of June and 3647 above.

STATIONS AND COMMUNICATION SYSTEMS

621.391.64 3650
Infrared Communications—M. C. Beese. (*Tele-Tech*, vol. 6, p. 53; May, 1947.) Summary of an Institute of Radio Engineers' paper.

621.394/.395].7(68.01) 3651
Communications Network of the Union of South Africa—D. P. J. Retief. (*Trans. S. Afr. I.E.E.*, vol. 38, part 3, pp. 84–112; March, 1947.) The development since the introduction of voice-frequency amplifiers in 1922 is described, with particular reference to recent expansions.

621.395.5:621.396.5 3652
Wire or Wireless?—T. Roddam. (*Wireless World*, vol. 53, pp. 236–238; July, 1947.) Outlines the future possibilities of wide-band f.m. v.h.f. links for the trunk communications at present handled by telephone lines.

621.396.1 3653
F.C.C. Makes Allocations for Short-Distance Communications—(*Electronics*, vol. 20, p. 152; June, 1947.) Brief survey of the allocation of frequencies in the 152-to-162 Mc. band.

621.396.24:551.510.535 3654
Application of the Theories of Indirect Propagation to the Calculation of Links Using Decametre Waves—R. Aubert. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 265–270; May, 1947.) Discussion on 1210 of May.

621.396.3:621.396.933 3655
International Commercial Aviation Radioteletype Systems—F. V. Long. (*Communications*, vol. 27, pp. 24–26 and 43; June, 1947.)

621.396.41:621.396.619.16 3656
Multiplex Broadcasting—A. M. Levine. (*Tele-Tech*, vol. 6, p. 55; May, 1947.) Summary of an Institute of Radio Engineers' paper. A system using time division multiplexing and pulse-time modulation for eight programs each of bandwidth 9.5 kc., on a single 930-Mc. carrier. See also 1213 of May (Grieg and Levine) and 3657 below.

621.396.41:621.396.619.16:621.396.97 3657
Multiplex Broadcasting—F. Altman and J. H. Dyer. (*Elec. Eng.*, vol. 66, pp. 372–380; April, 1947.) Multiplex operation and methods of modulation are briefly discussed with special reference to pulse-time modulation. An 8-channel system working on 930 Mc. and incorporating the cyclophon is described. The cyclophon, a special c.r. tube with rotating electron beam which acts as a cyclic switch and modulator or demodulator, is discussed in detail. The advantages of multiplex operation are summarized. See also 239 of February, 1213 of May (Grieg and Levine) and 3656 above.

621.396.41.029.64 3658
Calculation of Multiplex U.H.F. Radio-Telephone Links—Chireix. (*See* 3629.)

621.396.5 3659
Experimental Rural Radiotelephony—J. H. Moore, P. K. Seyler and S. B. Wright. (*Elec. Eng.*, vol. 66, pp. 346–348; April, 1947.) A description of an experimental radio installation on 44 to 50 Mc., worked on the party line system, for isolated rural communities having no telephone facilities.

621.396.619.11/.13 3660
Comparison of Amplitude and Frequency Modulation—M. G. Nicholson. (*Wireless Eng.*, vol. 24, pp. 197–208; July, 1947.) Comparisons of performance have previously been made between f.m. on 40 to 50 Mc. with a channel width of about 200 kc. and a.m. at a signal frequency of 1 Mc. or less with a channel width of about 10 kc. The present comparison is made under conditions of frequency stability, channel width and receiver bandwidth normally realized in the v.h.f. band.

It is concluded that f.m. is superior to a.m. only where fluctuation noise is the limiting factor. As regards interference, a.m. is superior to f.m. even if the selectivity of the a.m. receiver is identical with that of the f.m. receiver. A.m. has better discrimination against impulse noise, is less adversely affected by imperfect tuning and is superior to f.m. in "satellite" station operation. See also 3661 below (G.W.O.H.).

621.396.619.11/.13 3661
Amplitude and Frequency Modulation—G.W.O.H. (*Wireless Eng.*, vol. 24, p. 191; July, 1947.) Refers to Nicholson's paper (3660 above) and stresses that the merits of the two modulation systems must be compared under similar conditions.

621.396.619.13:518.3 3662
Radiation Chart for F.M. Stations—C. F. Guthrie. (*Communications*, vol. 27, pp. 34–36; May, 1947.) For determining the effective radiated power, a parameter required in Everett's range prediction chart (736 of 1946).

621.396.931 3663
V.H.F. Railroad Communications in Tunnels—J. P. Shanklin. (*Communications*, vol. 27, pp. 16–19; June, 1947.) Preliminary field strength measurements made in a disused water tunnel were followed by the erection of a train communication system in a railway tunnel 2760 feet long. 152-to 162-Mc. signals were fed into transmission lines in the crown of the tunnel from an external rhombic aerial. Reflecting wires were placed above the transmission lines to reduce signal loss.

621.396.931 3664
V.H.F. Radio Equipment Speeds up Railroad Operation—L. G. Sands. (*Tele-Tech*, vol. 6, pp. 38–41 and 111; May, 1947.) A description of modern two-way f.m. equipment used on United States railways.

621.396.931.029.62 3665
Mobile F.M. Communications Equipment for 30 to 44 Mc—R. B. Hoffman and E. W. Markow. (*Communications*, vol. 27, pp. 28–29 and 41, and 34–35; June and August, 1947.) The transmitter uses a crystal-controlled master oscillator, phase-shift variable transconductance modulation, and 4 frequency-multiplying and amplifying stages. The receiving selective-calling system uses a two-tube Wien-bridge oscillator circuit.

621.396.932 3666
Radio for Merchant Ships [Book Notice]—H. M. Stationery Office. 1s. (*Govt. Publ.* (London), p. 12; April, 1947.) Performance specifications.

621.396.932:[620.178+620.193 3667
Radio and Radar for Merchant Ships [Book Notice]—H.M. Stationery Office, 2d. (*Govt. Publ.* (London), p. 12; April, 1947.) A performance specification for the climatic and durability testing of marine radio and radar equipment.

SUBSIDIARY APPARATUS

621.313.2–9 3668
Sub-Miniature D.C. Motors—(*Electrician*, vol. 138, pp. 1157–1159; May 2, 1947.) For another account see 2943 of October.

621.314.632:621.315.59 3669
Contact Potential Difference in Silicon Crystal Rectifiers—W. E. Meyerhof. (*Phys. Rev.*, vol. 71, pp. 727–735; May 15, 1947.) Measurements show no correlation between the work function differences and the contact potential difference, which is practically independent of the metal used and also of the structure of the silicon surface.

621.314.65+537.525.5 3670
On the Mechanism of Dielectric Ignition and Resistance Ignition in Mercury Arc Rectifiers [Thesis]—N. Warmoltz. (*Philips Res. Rep.*, vol. 1, p. 379; November, 1946.) Summary only. A short survey based on the field theory of the low-pressure mercury arc.

621.316.53.029.5/.6 3671
A Design of Heavy-Current Contact, Particularly for Radio-Frequency Use—A. J. Maddock. (*Jour. I.E.E.* (London), vol. 94, p. 233; May, 1947.) Summary of 2249 of August.

621.318.44 3672
Toroidal Coils. Improved Winding Machine—E. R. Brooke. (*Elec. Rev.* (London), vol. 141, pp. 319–320; August 29, 1947.) Some details of a machine for quantity production of toroidal coils for transformers and chokes. Two rings are threaded on the core, both rings having detachable segments. One ring is channeled to carry enough wire for one winding, while the driving ring carries a wire feed pulley. The method of use for winding an 8-segment coil is described.

621.318.5 3673
Telephone Relays and Their Use in Electronic Circuits: Part 2—A. A. Chubb. (*Electronic Eng.* (London), vol. 19, pp. 211–213; July, 1947.) Various a.c. and d.c. circuits for operating small telephone relays are given, and a complete circuit for remote control of a 1-kw. transmitter and its associated receiver is described. For part 1 see 3294 of November.

621.396.68:621.397.5 3674
Television High Voltage R.F. Supplies—R. S. Mautner and O. H. Schade. (*RCA Rev.*, vol. 8, pp. 43–81; March, 1947.) A detailed consideration of the design of h.v. supply units using r.f. oscillators and voltage multiplier circuits. Sample calculations for a 75-w. 90-kv. supply and a 10-w. 30-kv. supply are included to illustrate the progressive steps in designing and calculating the circuit elements and operating conditions for a specified performance. For earlier work see 2169 of 1943 (Schade).

621.396.682 3675
A Special-Purpose Power Supply—P. W. Howells. (*Gen. Elec. Rev.*, vol. 50, pp. 34–39; June, 1947.) A stabilized power pack with output continuously variable between 160 and 1500 v. at 0.125 a. Characteristics include a low-ripple output voltage and a low output impedance to minimize the possibility of undesired coupling between load circuits through the power supply.

TELEVISION AND PHOTOTELEGRAPHY

621.396/.397].62:621.396.67.029.62 3676
Aerials for Ultra-Short Waves: Part 1—A Double Dipole for Television and F.M.—Maurice. (*See* 3433.)

621.396/.397].621.004.67 3677
The Servicing of Radio and Television Receivers—R. C. G. Williams. (*Jour. I.E.E.* (London), vol. 94, pp. 156–158; March, 1947.) Summary of 2224 of August.

621.397.2 3678
Developments in Picture Transmission—J. J. E. Aspin. (*Jour. I.E.E.* (London), vol. 94, p. 134; March, 1947.) Abstract of chairman's address to the South Midland Radio Group. An historical survey.

621.397.335 3679
New Techniques in Synchronizing-Signal Generators—Schoenfeld, Brown, and Milwitt. (*See* 3489.)

621.397.5:621.396.619.083 3680
A Method of Measuring the Degree of Modulation of a Television Signal—Buzalski. (*See* 3593).

621.397.5:621.396.68 3681
Television High Voltage R.F. Supplies—Mautner and Schade. (*See* 3674.)

621.397.6 3682
Portable Camera Chain for Field Use—L. Mautner. (*Tele-Tech*, vol. 6, pp. 26–31, 109; May, 1947.) Wartime developments have permitted redesign of portable television cameras and associated control equipment for outside broadcase use. An image-orthicon type of pick-up tube was chosen because of the wide range of sensitivity required and lack of shading available. A block diagram of a four-camera control system, and some circuit and construction details of camera-blanking methods, cable-delay compensation, and a camera control and monitor system are given.

621.397.6.001.8 3683
Simplified Television for Industry—R. E. Barrett and M. M. Goodman. (*Electronics*, vol. 20, pp. 120–124; June, 1947.) Complete circuit details for a 250-line 60-frame television system in which a new iconoscope simplifies the circuit and permits reproduction comparable to newspaper half-tones.

621.397.61 3684
The Paris Television Transmitting Centre—H. Delaby. (*Jour. Telev. Soc.*, vol. 4, pp. 307–313; December, 1946.) Translation of a French article abstracted in 1606 of June.

621.397.61–182.3 3685
Television O.B. [Outside Broadcast] Vehicle—(*Wireless World*, vol. 53, p. 241; July, 1947.) A 660-Mc. transmitter complete with iconoscope cameras is housed in a car and obtains power from a 3.5-kw. generator driven from the vehicle engine. The sound channel is conveyed by width-modulated pulses inserted in the line synchronization pulses and the 50-w. output is fed to a beamed horizontal dipole at the top of a 40-ft. telescopic mast.

621.397.62 3686
Television Receivers in Mass Production—D.G.F. (*Electronics*, vol. 20, pp. 86–91; June, 1947.) Design features of the RCA Victor postwar seven-inch, ten-inch, and projection models. Summary of an Institute of Radio Engineer's paper by A. Wright and E. Clark.

621.397.62 3687
Television Receiver Construction: Part 5—(*Wireless World*, vol. 53, pp. 251–257; July, 1947.) Line time-base and high-voltage supply for the c.r.t. For earlier parts see 2595 of September and back references.

621.397.62 3688
Television Receivers—A. Wright. (*RCA Rev.*, vol. 8, pp. 5–28; March, 1947.) A detailed survey of RCA direct viewing and projection type receivers with photographs and circuit diagrams. General circuit principles are considered. The r.f. tuner uses push-pull neutralized triode amplification, a push-pull triode frequency changer and switched coil tuning. The i.f. amplifier has staggered tuned circuits with rejection circuits tuned to adjacent channels. An unusual circuit for line synchronization, which is immune from interference, uses a stable sinusoidal oscillator whose phase is controlled by the line synchronization pulses. The magnetically focused cathode-ray tube has an ion trap to prevent ion bombardment of the screen from causing discoloration.

621.397.62 3689
Television Receiving Equipment [Book Review]—W. T. Cocking. Iliffe and Sons, London, 2nd edn., 1947. 339 pp., 12s. 6d. (PROC. I.R.E., vol. 35, p. 706; July, 1947.) For another review see 2966 of October.

TRANSMISSION

621.317.76 3690
WWV—World Standard Frequency Generator—(*Tele-Tech*, vol. 6, pp. 42–43; May, 1947.) Photographs of some of the equipment used in the standard frequency transmissions.

621.391.63:621.325.53 3691
The Concentrated-Arc Lamp as a Source of Modulated Radiation—W. D. Buckingham and C. R. Deibert. (*Jour. Soc. Mot. Pic. Eng.*, vol. 48, pp. 324–340; April, 1947.) Discussion, pp. 340–342.) A lamp using as radiation source a thin layer of molten zirconium maintained as an incandescent pool by intense argon ion bombardment. The radiation can be modulated at a.f. by modulating the lamp current. The use of suitable modulator circuits, with optical filters to select the best spectral region, enables the output to follow the current modulation with good fidelity.

621.396.1 3692
Types of Emission—(*R.S.G.B. Bull.*, vol. 22, p. 124; February, 1947.) Recommendations accepted by the G.P.O. for the bands allotted to British amateurs are: 1.75, 3.5, 7, and 14 Mc.; c.w., a.m., 28 and 58.5 Mc.; c.w., m.c.w., a.m., f.m., 2300 to 2450 Mc.; any type of emission, including television but excluding pulse transmission.

621.396.61 3693
The Radio Mike—J. L. Hathaway and R. Kennedy. (*RCA Rev.*, vol. 8, pp. 251–258; June, 1947.) A smaller, lighter, and more efficient transmitter to replace the N.B.C. "beermug." The design, uses, and tests applied are described.

621.396.61:621.396.712 3694
Placing a 3-KW. F.M. Broadcast Transmitter in Operation—R. G. Soule, Jr. (*Communications*, vol. 27, pp. 16–18, 46; May, 1947.) Preliminary tests of the area were made with a 50-w. unit which is described. The transmitter itself has a four-bay circular aerial and provides 8.5 kw. radiated power.

621.396.61.029.62 3695
A Low-Cost 2-Meter Transmitter—E. P. Tilton. (*QST*, vol. 31, pp. 26–29, 122; April, 1947.) Circuit and constructional details of a stabilized modulated oscillator with an output of about 3.5 w.

621.396.61.029.62 3696
BC-625 on 144 Mc/s.—L. W. May, Jr. (*Radio Craft*, vol. 18, pp. 35–36, 75; April, 1947.) Complete circuit details of the modifications necessary to convert the transmitter of the Army SCR-522 set for amateur use.

621.396.611.21:621.316.726.078.3:621.396.712 3697
Improvements in Synchronisation of B.B.C. Transmitters: 1938–1946—W. E. C. Varley. (*B.B.C. Quart.*, vol. 2, pp. 51–58; April, 1947.) A review of the development of frequency control of broadcasting transmitters from the pre-1938 tuning fork drive to the present crystal drive. The performance of various crytal-controlled oscillators is given and the technique of frequency comparison described with circuit details.

621.396.615.141.2 3698
Modulated Magnetrons—L. P. Smith, J. Kurshan, and J. S. Donal. (*Tele-Tech*, vol. 6, p. 57; May, 1947.) Summary of an Institute of Radio Engineers' paper. Picture or audio signals control the electron guns and cause the magnetron to generate a f.m. wave without a.m. Under the influence of a static magnetic field and a r.f. electric field, an electron beam follows a spiral path within the cavity and applies f.m. to the natural resonant frequency of the magnetron. For general discussion of this technique and two specific applications, see 3699, 3730, and 3731 below.

621.396.615.141.2:621.316.726 3699
Frequency Modulation and Control by Electron Beams—L. P. Smith and C. I. Shulman. (Proc. I.R.E., vol. 35, pp. 644–657; July, 1947.) General formulas for the effect of electron beams on resonant systems in terms of frequency shift and change in Q are derived both from the point of view of lumped circuits and also from a general electromagnetic field standpoint. Check measurements of the frequency shift produced by such a beam in a multivane magnetron are described. It is shown that this method of frequency control is ideal for frequency modulation or automatic frequency stabilization of magnetrons and that for the former purpose the amplitude and phase distortions are negligible.

621.396.619.11/.13 3700
Generalized Theory of Multitone Amplitude and Frequency Modulation—L. J. Giacoletto. (Proc. I.R.E., vol. 35, pp. 680–693; July, 1947.) The frequency spectra produced by single-tone, two-tone, and multitone modulating signals in the case of a.m., f.m., and combined a.m. and f.m. are studied. Computations of the frequency spectra for typical cases are made and compared with actual spectra obtained by means of a spectrum analyzer.

621.396.619.11 3701
Overmodulation Splatter Suppression—O. G. Villard, Jr. (*QST*, vol. 31, pp. 13–20; June, 1947.) A method of filling in the overmodulation gaps in the carrier and so preventing the generation of spurious sidebands.

621.396.619.15:621.396.3 3702
Relative Amplitude of Side Frequencies in On-Off and Frequency-Shift Telegraph Keying—G. S. Wickizer. (*RCA Rev.*, vol. 8, pp. 158–168; March, 1947.) Measurements and calculations on the frequency spread of the sidebands indicate that frequency-shift keying requires less bandwidth than on-off keying as the characters may be shaped by a low-pass filter.

621.396.619.23 3703
A 40-Watt Modulator with Cathode-Coupled Driver—W. J. Lattin. (*QST*, vol. 31, pp. 42–44; April, 1947.) Circuit details of a unit with built-in power supply and four stages terminating in a 6L6G push-pull class AB2 output stage.

621.396.645 3704
Design of Linear Amplifiers for Single Side Band Transmitters—E. Green. (*Marconi Rev.*, vol. 10, pp. 11–16; January and March, 1947.) Distortion of a modulated carrier in a transmitter due to varying input impedance of the power amplifier is avoided by using screen grid driving tubes with an impedance transforming network.

621.396.65.029.63 3705
An Experimental Transmitter for Ultra-Short-Wave Radio-Telephony with Frequency Modulation—A. van Weel. (*Philips Tech. Rev.*, vol. 8, pp. 121–128; April, 1946.) For another account see 2606 of September.

VACUUM TUBES AND THERMIONICS

621.314.6.032.212 3706
A Cold Cathode Rectifier—W. H. Bennett. (*Jour. Appl. Phys.*, vol. 18, pp. 479–482; May, 1947.) Corona discharge is used at atmospheric and higher pressures in H and N free from electron-attaching impurities. Such rectifiers have definite advantages where current requirements are small.

621.314.67 3707
Determination of Current and Dissipation Values for High-Vacuum Rectifier Tubes—A. P. Kauzmann. (*RCA Rev.*, vol. 8, pp. 82–97; March, 1947.) "Rectifier data are shown graphically with generalized parameters from which it is possible to determine the peak steady-state current, the maximum possible hot-switching current, and the dissipations in the diode and in any added series resistors. The paper covers capacitive-input filters with large capacitors, and includes half-wave, full-wave, and voltage-doubler circuits. A table of operating conditions and efficiency for a group of typical rectifiers is included."

621.314.671:621.386.1:616-073.75 3708
High-Voltage Rectifier Valves for X-Ray Diagnostics—J. H. van der Tuuk. (*Philips Tech. Rev.*, vol. 8, pp. 199–205; July, 1946.) Relative merits of gas-filled and vacuum tubes, and construction of new vacuum tubes with thoriated tungsten cathodes.

621.383.4:535.215 3709
Lead Sulphide Photoconductive Cells—L. Sosnowski, J. Starkiewicz, and O. Simpson. (*Nature* (London), vol. 159, pp. 818–819; June 14, 1947.) The method of production, developed at the Admiralty Research Laboratory, is described in detail. Maximum sensitivity is assured when both lead and oxygen impurity centers are present in sufficient quantity and with relative concentration such that minimum conductivity and zero thermoelectric power are obtained. Theory is presented which is in general quantitative agreement with experiment as regards sensitivity, rectifying effect, and time of response.

621.383.5 3710
Fatigue in Selenium Barrier Layer Photocells—R. A. Houstoun. (*Phil. Mag.*, vol. 37, pp. 13–17; January, 1946.) See also 3433 of 1941.

621.385+621.396.694 3711
Tube Registry—(*Electronics*, vol. 20, pp. 244, 247; June, 1947.) Characteristics of iconoscope Type 5527, triode power amplifiers and oscillators (Types 195 and 196) and c.r. tube, Type 3MP1. See also 2976 of October and 2288 of August.

621.385:518.5 3712
Electrostatic Storage—J. Rajachman. (*Tele-Tech*, vol. 6, p. 61; May, 1947.) Summary of an Institute of Radio Engineers' paper. Describes a vacuum-tube "memory" for electronic computers. A multicellular anode stores up to 4096 impulses separately. Storing time is indefinite and reading follows the reading call by only a few microseconds and can be repeated indefinitely.

621.385:537.533.8 3713
Transit-Angle Suppression in Microwave Tubes—J. H. Owen Harries. (*Electronics*, vol. 20, pp. 132–134; June, 1947.) Details of research into the control of the phase of the u.h.f. field near copper-target anodes, for the suppression of secondary emission. A transverse modulated electron beam was passed through the aperture in a subanode, then traversed a distance d to the surface of the target anode. A resonant cavity in the output circuit was tuned to the modulation frequency f and power transfer was recorded by a diode. The transit angle $\phi = 10^3 d\pi/\lambda V_b^{\frac{1}{2}}$, where V_b is the target and subanode voltage and λ is the wavelength corresponding to f, was varied by altering d and/or V_b. Tests were carried out for λ 40 cm. Three types of copper target were used: (a) polished, (b) roughened and carbonized, and (c) slotted and carbonized. Plots of ϕ versus power-output efficiency show the slotted targets to be the most efficient, with values comparable with theory for $\phi > 0.3\pi$. The theory of suppression is illustrated by graphs in which the target and subanode currents are plotted against ωt for values of ϕ from $\pi/6$ to π.

621.385.029.63/.64]+621.396.615.14 3714
On Some Modern Constructions and Some Recent Designs of Ultra-Short-Wave Receiving and Transmitting Valves—R. Warnecke. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 81–94; February, 1947.) Technical details obtained by the author, during visits to Britain and the United States, of the resnatron, klystrons, and other high-power velocity-modulation tubes, and traveling-wave tubes. The prionotron designed by the author is also described.

621.385.029.63/.64 3715
Helical-Wave Properties—C. C. Cutler. (*Tele-Tech*, vol. 6, p. 56; May, 1947.) Summary of an Institute of Radio Engineers' paper. Probe measurements in a traveling-wave tube show that the longitudinal field component along the axis is greater than that predicted by theory.

[621.385.029.63/.64:621.392+537.291] 3716
On the Theory of Progressive-Wave Amplifiers—Blanc-Lapierre Lapostolle, Voge, and Wallauschek. (*See* 3421.)

621.385.1+621.396.694 3717
A New Range of Glass-Based Valves—(*Electronic Eng.*, vol. 19, p. 231; July, 1947.) Type numbers and brief descriptions are given of the new spigotless miniature tubes with B8A base. The heater current in the a.c./d.c. range is 0.1-ampere and the bulb size approximately 20 mm. An a.c. range with the B8A base and 6.3-volts heaters is also to be introduced, together with a high-gain screened h.f. pentode and a triode especially designed for television reception. Location of these tubes in their holders is effected by a small boss on the side of the base.

621.385.1+621.396.694]:389.6 3718
B8A Valve Base—(*Electronic Eng.*, vol. 19, p. 235; July, 1947.) For details of the base, see 957 and 980 of April. A spigotless version is here announced, the ultimate aim being to standardize a range of tubes to fit both bases.

621.385.1 3719
The Control of the Current Distribution in Electron Tubes—J. L. H. Jonker. (*Philips

Res. Rep., vol. 1, pp. 331–338; November, 1946. Characteristics for control by a negative grid are calculated and shown to agree with experiment.

621.385.1 3720
Miniature Tubes in War and Peace—N. H. Green. (*RCA Rev.*, vol. 8, pp. 331–341; June, 1947.) " . . . describes the design features which account for the versatility and lower cost of the miniature tube, and cites several varied applications of miniatures in both military and commercial equipment. A table showing typical present-day applications for miniature tubes is included."

621.385.1:621.396.694.012.8 3721
Valve Equivalent Circuit—A. W. Keen: B. Salzberg. (*Wireless Eng.*, vol. 24, pp. 217–218; July, 1947.) The use of "equivalent" circuits with a voltage or a current generator is optional and a matter of convenience for external performance, but in general neither gives the correct value of internal power dissipation. See also 2622 of September (Salzberg) and back references.

621.385.2:621.396.822.029.6 3722
A Coaxial-Line Diode Noise Source for U.H.F.—H. Johnson. (*RCA Rev.*, vol. 8, pp. 169–185; March, 1947.) The diode has a single-turn helical filament coaxial with and connected to the inner conductor, the outer conductor being the anode. The tube is connected on one side to a lossy line to give the correct impedance load and on the other side to the 50-ohm input line of the receiver under test. A diode current of 100 milliamperes may be obtained corresponding to a noise factor of 20 db. The effect of the filament capacitance in producing standing-wave errors is considered and the transit-time loss is calculated and is shown to be about 3 db at 3000 Mc. A comparison with signal generator measurements at 750 and 1500 Mc. gave a maximum discrepancy of 0.4 db.

621.385.2.032.216 3723
Effect of the Saturation Current on the Space-Charge Current in Valves Using Oxide Cathodes—R. Champeix. (*Compt. Rend. Acad. Sci.* (Paris), vol. 224, pp. 1626–1628; June 9, 1947.) In a diode with oxide cathode the space-charge current may increase, remain unchanged or sometimes even decrease when the saturation current increases. These anomalous results are explained.

621.385.3/.5:621.317.336.1 3724
The Measurement of Dynamic Mutual Conductance of Valves Using the Grounded-Grid Triode Mode of Operation—Gutmann. (*See* 3576.)

621.385.4.029.63 3725
Tetrodes vs. Triodes—W. G. Wagener. (*Tele-Tech*, vol. 6, p. 54; May, 1947.) Summary of an Institute of Radio Engineers' paper. Neutralized tetrodes offer higher gain and greater circuit stability than neutralized triodes in the region of 500 Mc.

621.385.831:621.317.382 3726
Power Measurement of Class B Audio Amplifier Tubes—Heacock. (*See* 3578.)

621.396.615.14 3727
The Excitation of Resonant Circuits by Electron Currents in the Transit-Time Domain—Gundlach. (*See* 3468.)

621.396.615.141.2.032.21 3728
Coaxial Tantalum Cylinder Cathode for Continuous-Wave Magnetrons—R. L. Jepsen. (*RCA Rev.*, vol. 8, pp. 301–311; June, 1947.) The use of a cathode with a tungsten inner and tantalum outer conductor eliminates many of the drawbacks of normal cathodes.

621.396.615.141.2 3729
On Electron Oscillations in a Magnetron—V. I. Kalinin and I. I. Wassermann. (*Bull. Acad. Sci.* (U.R.S.S.), sér. phys., vol. 10, no. 1, pp. 103–110; 1946. In Russian.) The electron oscillations in a split-anode magnetron are studied from the standpoint of spatial irregularities in the electron beam. The oscillations of the first order, which appear in all magnetrons under conditions close to the critical régime, are due to a certain radial irregularity. These conditions can be reduced to a system with a retarding field. In considering oscillations in a split-anode magnetron a conception of a tangential irregularity which takes place in the "ring current" in close proximity to the anode, is introduced and the frequency of the oscillations determined. Results of experiments with multisegment (from 4 to 20 segments) magnetrons are in satisfactory agreement with the theoretical conclusions. For "ring current" see 64 of 1937 (Möller).

621.396.615.141.2 3730
A Frequency-Modulated Magnetron for Super-High Frequencies—G. R. Kilgore, C. I. Shulman, and J. Kurshan. (Proc. I.R.E., vol. 35, pp. 657–664; July, 1947.) The development of a 25-watt 4000-Mc. c.w. magnetron capable of a frequency deviation of 2.5 Mc. without a.m. is described. F.m. is accomplished by the introduction of electron beams in two of the twelve cavities (see 3699 above). Design details and performance data are given.

621.396.615.141.2 3731
A 1-Kilowatt Frequency-Modulated Magnetron for 900 Megacycles—J. S. Donal, Jr., R. R. Bush, C. L. Cuccia, and H. R. Hegbar. (Proc. I.R.E., vol. 35, pp. 664–669; July, 1947.) The design and performance of the magnetron are described. F.m. is accomplished by the introduction of electron beams in nine of the twelve cavities (see 3699 above) and a deviation of 3.5 Mc. is obtained.

621.396.615.141.2:513.732.6 3732
A Flux Plotting Method for Obtaining Fields Satisfying Maxwell's Equations, with Applications to the Magnetron—Crout. (*See* 3560.)

621.396.615.141.2:537.533.8 3733
The Secondary Emission in Magnetron Oscillators—S. Ya. Braude and I. E. Ostrovski. (*Bull. Acad. Sci.* (U.R.S.S.), sér. phys., vol. 10, no. 1, pp. 65–73; 1946. In Russian.) It has been observed that under certain conditions the anode current of a magnetron with grid control can be from 5 to 7 times the normal. A detailed investigation, both theoretical and experimental, shows that this phenomenon is due not to the ionization of the residual gases in the magnetron, as was supposed by some investigators, but to secondary emission from the grid of electrons traveling in the grid-anode space. It is also shown that the grid secondary emission can be greatly increased if oscillating potentials are present in the magnetron.

621.396.615.141.2:621.316.726 3734
Frequency Modulation and Control by Electron Beams—Smith and Shulman. (*See* 3699.)

621.396.615.141.2:621.365.92 3735
A Magnetron Oscillator for Dielectric Heating—R. B. Nelson. (*Jour. Appl. Phys.*, vol. 18, pp. 356–361; April, 1947.) Design and performance of a magnetron having 5 kw. continuous output at 1050 Mc.

621.396.615.141.2:621.396.933.2 3736
Stabilized Magnetron for Beacon Service: Part 1—Development of Unstabilized Tube—J. S. Donal, Jr., C. L. Cuccia, and B. B. Brown. (*RCA Rev.*, vol. 8, pp. 352–361; June, 1947.) The design of the tube is unconventional inasmuch as all the parts are supported on a header to which the envelope is welded. The inserts in the magnetic circuit are at cathode potential. The tube is designed for a pulsed input power of 2.5 kw. The unstabilized peak output is approximately 1 kw. at 2500 volts anode potential and a frequency of 9310 Mc.

621.396.615.141.2:621.396.933.2 3737
Stabilized Magnetron for Beacon Service: Part 2—Engineering of Tube and Stabilizer—C. P. Vogel and W. J. Dodds. (*RCA Rev.*, vol. 8, pp. 361–372; June, 1947.) The frequency-stabilization device includes an invar tunable cavity using a plunger supported by a spindle of higher-expansion steel and is filled with dry nitrogen. The wave-guide coupling to the load contains adjustable screw tuners. The stabilization process consists in the proper adjustment of these screws. The stability is improved by a factor of 10. For part 1 see 3736 above.

621.396.615.142 3738
The Principles of a General Theory of the Generation of Electron Oscillations at Ultra-High Frequencies—Kalinin. (*See* 3470.)

621.396.615.142.2 3739
The Theory of a Single-Circuit Klystron—L. N. Loshakov and S. D. Gvozdover. (*Bull. Acad. Sci.* (U.R.S.S.) sér. phys., vol. 10, no. 1, pp. 79–86; 1946. In Russian.) In a reflex klystron the buncher and collector voltages coincide, with the result that its efficiency is lower than that of a klystron using two coupled resonators. To simplify the construction of the latter type, a klystron using a single resonator but with separate buncher and collector voltages is proposed (Fig. 1). An approximate theory of this klystron is given together with some preliminary experimental results.

621.396.615.142.2 3740
The Self-Excitation of a Reflex Klystron—S. D. Gvozdover. (*Bull. Acad. Sci.* (U.R.S.S.) sér. phys., vol. 10, no. 1, pp. 75–78; 1946. In Russian.) The theory of the reflex klystron is discussed and the condition (5) necessary for self-excitation is established for the case when the potential of the reflecting electrode is equal to that of the cathode. Equation (6) determining the frequency of self-oscillations is also obtained.

621.396.615.142.2:621.396.645.029.64 3741
On U.H.F. Amplification and on the Resonance Method for Suppressing Noise in a Klystron—Katsman. (*See* 3481.)

MISCELLANEOUS

061.6"1947"I.R.E.:621.396 3742
I.R.E. Reveals Engineering Advances—(*Tele-Tech.*, vol. 6, pp. 52–61; May, 1947.) A Report of the 1947 Institute of Radio Engineers National Convention, with abstracts of 43 of the 125 technical addresses presented. For abstracts of selected individual papers, see other sections.

061.6(73) 3743
Science Advancing—The Future of Testing—E. U. Condon. (*ASTM Bull.*, no. 146, pp. 53–58; May, 1947.) A review of some of the wartime activities of the National Bureau of Standards and brief discussion of future developments.

621.3 3744
British Industries Fair—(*Elect. Rev.*, (London), vol. 140, pp. 697–721; May 2, 1947.) A guide to the electrical exhibits at Castle Bromwich, Birmingham, and Olympia and Earls Court, London.

621.38/.39+539.17 3745
Nucleonics and Electronics—K. Henney. (*Electronics*, vol. 20, pp. 80–81; June, 1947.) Nucleonics, a generic name for atomic energy and related subjects and intimately related to electronics.

621.385.1+621.396.694]:389.6 3746
B8A Valve Base—(*See* 3718.)

PROCEEDINGS OF THE I.R.E.

(Including WAVES AND ELECTRONS Section)

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Index to Volume 35—1947

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Input	27.2 μufd
Output	7.6 μufd
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D-C Screen Voltage	1000 Max. Volts
D-C Grid Voltage	—500 Max. Volts
D-C Plate Current	700 Max. ma
Plate Dissipation	1000 Max. Watts
Screen Dissipation	75 Max. Watts
Grid Dissipation	25 Max. Watts

TYPICAL OPERATION
(Frequencies below 40 Mc.)

D-C Plate Voltage	6000 Volts
D-C Screen Voltage	500 Volts
D-C Grid Voltage	—200 Volts
D-C Plate Current	681 ma
D-C Screen Current	141 ma
D-C Grid Current	41 ma
Screen Dissipation	71 Watts
Grid Dissipation	6.1 Watts
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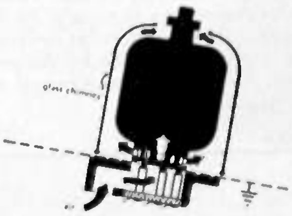

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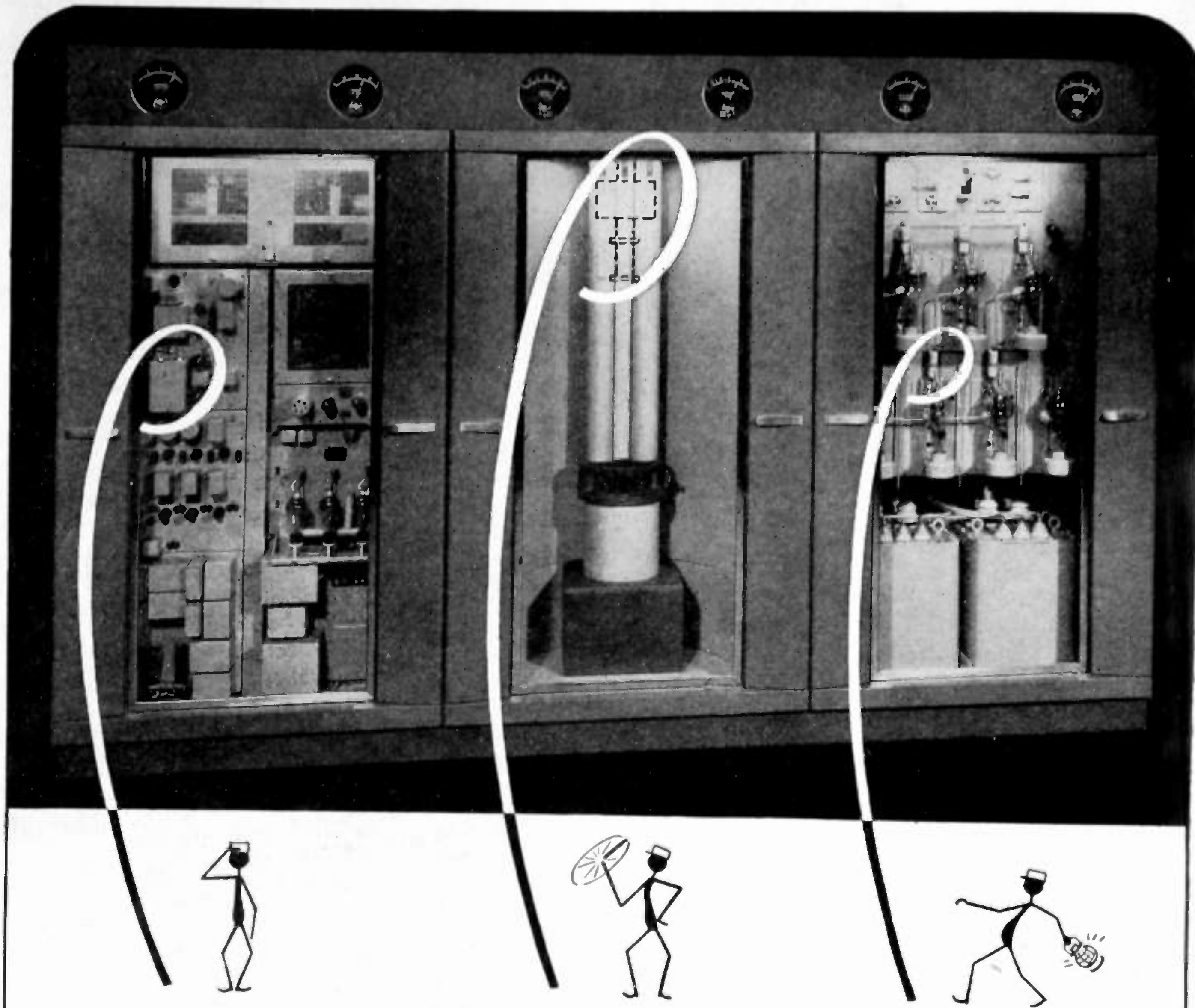

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"Induction Heating Equipment and Applications," by E. S. Winlund, Radio Corporation of America; October 21, 1947.

"Experiences with Dielectric Heating," by D. P. Whitacre, Sr., Vetric, Inc.; October 21, 1947.

"Electronic Methods of Gauging," by R. L. Sink, Consolidated Eng. Corporation; October 21, 1947.

LOUISVILLE

Election of Officers, September 5, 1947.

"New Instruments for Radio and Electrical Measurements," by I. G. Easton, General Radio Company; October 29, 1947.

MONTREAL

"Radar and Microwaves," by J. O. Perrine, American Telephone and Telegraph Company; October 1, 1947.

"Stormy Weather," by J. S. Marshall, McGill University; October 8, 1947.

NORTH CAROLINA-VIRGINIA

"A New Radio-Frequency Bridge for F. M." by I. G. Easton, General Radio Company; October 17, 1947.

OTTAWA

"Problems Involved in High-Fidelity Reproduction of Music," by J. E. Breeze, National Research Council; October 9, 1947.

"Measurement of Time," by J. P. Henderson, Dominion Observatory; October 30, 1947.

PHILADELPHIA

"Viewing Screens for Projection Television Receivers," by W. E. Bradley, Philco Corporation; October 2, 1947.

PITTSBURGH

"Stabilizers and Servo Mechanisms," by C. R. Hanna, Westinghouse Research Laboratories; September 8, 1947.

PORTLAND

"Television and Microwave Research," by J. W. McRae, Bell Telephone Laboratories, Inc.; Ocober 2, 1947.

PRINCETON

"Surmises on Atomic Energy Development," by A. N. Goldsmith, Consulting Engineer and Editor, the PROCEEDINGS OF THE I.R.E.; October 9, 1947.

ROCHESTER

"Instrumentation for Bikini Bomb Tests," by K. Swartzell, Cornell Aeronautical Laboratories; October 16, 1947.

SACRAMENTO

"Frequency-Modulation Receivers," by W. E. Evans, Jr., McClatchy Broadcasting Company; October 20, 1947.

ST. LOUIS

"Proximity Fuze," by F. W. Bubb, Jr., Washington University; October 23, 1947.

SAN DIEGO

"U.H.F. Measurement Techniques," by R. A. Soderman, General Radio Company; October 1, 1947.

SYRACUSE

"Highway Mobile Telephone Service," by D. Dewire, New York Telephone Company; October 3, 1947.

(Continued on page 38A)

SECTION MEETINGS

(Continued from page 36A)

TORONTO

"Canadian National Exhibition Sound System," by J. R. Bain, Northern Electric Company; October 6, 1947.

"Tasks for the Radio Engineer in Communication for the Army in the Field," by Col. A. E. Wrinch, Royal Canadian Corps Signals; October 27, 1947.

TWIN CITIES

"Solar Radiation and its Effect Upon Power Transmission and Radio Communication," by J. T. Wilson, Allis-Chalmers Company; October 9, 1947.

WASHINGTON

"Cosmic Radio Noise," by J. W. Herbstreit, National Bureau of Standards; October 13, 1947.

WILLIAMSPORT

"Considerations in High-Fidelity Transformers and Amplifier Design," by L. Walsh, Dinion Coil Company, Inc.; October 8, 1947.

Election of new Chairman; October 8, 1947.

SUBSECTIONS

HAMILTON

"Frequency Modulation," by B. Graham, Sparton of Canada; September 22, 1947.

"I.R.E. Activities," by F. R. Pounsett, Stromberg Carlson; October 20, 1947.

URBANA

"Registration of Professional Engineers," by T. C. Shedd, State of Illinois; April 24, 1947.

"High-Frequency Welding—A New Tool—A New Use for Electronics," by W. N. Parker, Radio Corporation of America; May 6, 1947.

STUDENT BRANCH MEETINGS

UNIVERSITY OF ALBERTA, I.R.E. BRANCH

"Magic of Flourescence," "Railroading," moving pictures; Reading of Model Constitution; October 22, 1947.

UNIVERSITY OF CALIFORNIA, I.R.E.—A.I.E.E. BRANCH

"The Engineer and his Professional Society," by B. E. Shackelford, President-Elect of I.R.E.; September 25, 1947.

KANSAS STATE COLLEGE, I.R.E. BRANCH

Election of Officers and Adopting of Constitution; October 2, 1947.

UNIVERSITY OF MICHIGAN, I.R.E.—A.I.E.E. BRANCH

"Magnetic Recording," by J. S. Kemp, Armour Research Foundation; Contest for Best Student Papers Announced; Demonstration of Wire and Tape Recorders, by R. Hammer, student; October 15, 1947.

COLLEGE OF THE CITY OF NEW YORK, I.R.E. BRANCH

"Student Membership in the I.R.E.," by F. B. Llewellyn, Junior Past President of I.R.E.; October 7, 1947.

(Continued on page 40A)

STUDENT BRANCH MEETINGS

(Continued from page 38A)

"Practical Coil Design," by Ben Yelsay, President of Coil Winders, Inc.; October 14, 1947.

"Electroencephalographic Technique," by W. G. Egan, formerly, Walter Reed General Hospital; Discussion of Work Project on War Surplus Electronic Equipment in Conjunction with Electrical Engineering Department; October 21, 1947.

New York University, I.R.E. Branch

Election of Officers; October 17, 1947.

North Carolina State College, I.R.E. Branch

"Carrier Circuits," by B. O. Jenkins, Communications Engineer, Carolina Power and Light Company; October 15, 1947.

Adoption of Constitution; October 29, 1947.

Rutgers University, I.R.E.—A.I.E.E. Branch

"Advantages of Joining Engineering Societies," by E. C. Plant, Public Service Electric Company; October 7, 1947.

Adoption of Constitution and Announcement of Program Committee Members; October 14, 1947.

Stanford University, I.R.E.—A.I.E.E. Branch

"Metal Locating Devices," by G. R. Fisher, Head of Fisher Research Laboratory; Election of Corresponding Secretaries; October 22, 1947.

University of Texas, I.R.E.—A.I.E.E. Branch

"The National and Local Organizations of I.R.E.," by W. E. Gordan, Associate Director of E.E.R.L.; "Engineer and the A.I.E.E.," by Sam Friedsam, Chairman of South Texas Section of A.I.E.E.; "The local A.I.E.E.," by W. R. Warren, Member of A.I.E.E. Committee on Student Branches; September 30, 1947.

University of Utah, I.R.E.—A.I.E.E. Branch

"Student Branch Organization and Business," by G. M. Waterfall, Chairman; Appointment of Committee Chairmen; October 21, 1947.

MEMBERSHIP

The following transfers and admissions were approved on November 11, 1947; to be effective as of December 1, 1947:

Transfer to Senior Member

Baracket, A. J., 49 Bell St., Bloomfield, N. J.
Berger, U. S., Bell Telephone Laboratories, Inc., Whippany, N. J.
Bitter, A. R., 4292 Monroe St., Toledo 6, Ohio
Carter, W. H., Jr., 1309 Marshall Ave., Houston 6, Tex.
Chinski, G. R. 1511 Marshall Ave., Houston 6, Tex.
Frear, W., 108 Roseland Ave., Fox Chase, Philadelphia, Pa.
Gieringer, C. K., The Leibel-Flarsheim Co., 303 W. Third St., Cincinnati 2, Ohio
Goldstine, H. E., RCA Laboratories Division, Rocky Point, N. Y.
Guzzi, P. N., Cangallo 1286, Buenos Aires, Argentina
Johnson, H. W., RCA International Division, 745 Fifth Ave., New York 22, N.Y.
Lewinstein, M., 148-21—90 Ave., Jamaica 2, L. I., N. Y.
Martin, D. W., Bendix Radio Division, Baltimore 4, Md.

MEMBERSHIP

Reed. O. W. B., Jr., 970 National Press Building, Washington 4, D. C.
Richards, A. P., 4224 Alton Place, N. W., Washington 16, D. C.
Shaw, R. F., 3504 Baring St., Philadelphia 4, Pa.
Shotliff, L. A., 25 W. 90 St., New York 24, N. Y.
Somerville, T., B.B.C. Research Department Nightingale Square, London, S.W. 12, England
Stec. C. L., 2720 Terrace Road, S. E., Washington 20, D. C.
Taber, H. E., c/o Dir. Electrical and Communications Department, Department of National Defense, Ottawa, Ont., Canada
Taylor, A. H., Box 532, Alamosa, Col.
Taylor, A. R., Joint Research and Development Board, 3171 New War Department Bldg., 21 & Virginia Ave., N. W., Washington, D. C.

Admission to Senior Member

Aronson, C. A., 960 Glencove Rd., Syracuse 6, N. Y.
Eliasen, E., 5/7 St. Helens Place, London E.C. 3, England
Hovey, B. K., 208 Thaw Hall, University of Pittsburgh, Pittsburgh 13, Pa.
Kunz, W. J., Jr., 66 Demarest Ave., Westwood, N. J.
McClurg, G. H., 31 N. Granada St., Arlington, Va.
Rector, R. V., 307 Demong Dr., Syracuse 3, N. Y.
Tyack, L. C., Box 112, Lovett Ave., Little Silver, N. J.

Transfer to Member

Altomare, R. E., Capitol Radio Engineering Institute, 3224—16 St., N. W., Washington 10, D. C.
Brown, F. M., 4635 S. E. Hawthorne Blvd,. Portland 15, Ore.
Castle, C. X., 1469 Church St., N. W., Washington 5, D. C.
Cozzens, W. D., 2255 Madison Ave., Odgen, Utah
Crick, W. R., 22 Avenue Rd., Chelmsford, Essex, England
Devaney, R. G., 631 S. 60 St., Philadelphia 43, Pa.
Eaton, J. E., 132 N. Terrace Ave., Mount Vernon, N. Y.
Forrest, G. E., Specialities, Inc., Skunks Misery Rd., Syosset, L. I., N. Y.
Gilbert, C. W., 52 Hathaway Ct., Pittsburgh 21, Pa.
Greene, W. E., Office of Naval Research, Navy Department, Washington 25, D. C.
Greenwood, I. A., Jr., General Presicion Laboratory, Pleasantville, N. Y.
Henrich, W. H., 16 Laurel Ave., E. Norwalk, Conn.
Hill, J. S., Box 298, Hudson, Ohio
McConnell, I. R., 1422 Tenth Ave., Neptune, N. J.
Olson, C. P., Jr., 1251 S. Maryland Ave., Glendale 5, Calif.
Petkovsek, J., Jr., 565 Walnut, Beaumont, Tex.
Pile, D. H., 80 Paterson Village Dr., Dayton 9, Ohio
Rodriguez, M. F., Puente Peredo No. 5, Depto. 3, Mexico, D. F., Mexico
Simons, K. A., 5201 W. 77 Terrace, Overland Park, Kan.
Stoops, C. W., 1615 Kenyon St., N. W., Apt. 55, Washington 10, D. C.
Tooley, M. D., 28 Queen's Rd., Colchester, Essex, England
Tuttle, D. F., Jr., 1429 Cambridge St., Cambridge 39, Mass.
Wedel, J. J., Jr., Box 223, Altadena, Calif.
Weinschel, B. O., 365 E. 21 St., Brooklyn 26, N. Y.
Wheeler, L. B., 5707 Wyngate Dr., Bethesda 14, Md.

Admission to Member

Alne, L. A., Electronics Test, U.S. Naval Air Test Center, Patuxent River, Md.

(Continued on page 42A)

MEMBERSHIP

(Continued from page 41A)

Anderson, E. R., Box 35, Hoffmans, N. Y.
Bruland, R. V., c/o Pan American-Grace Airways, Inc., Lima, Peru
Cronin, W. J., 1921 Kalorama Rd., N. W., Washington 9, D. C.
Ehret, R. J., Brown Instrument Co., Wayne and Roberts Ave., Philadelphia 44, Pa.
Frevert A. W., Box 255A, R.F.D. 2, Neptune, N. J.
Giannetto, C., Box 267, Station E, Cincinnati 19, Ohio
Gross, A. S., 53 W. 175 St., New York 53, N. Y.
Gruen, W. J., c/o J. Siwiec, 655 W. Taft Ave., Bridgeport, Conn.
Gunter, D. L., 2500 Que St., N. W., Washington 7, D. C.
Hafstrom, W. F., 601 Seventh, Liverpool, N. Y.
Miller, C. F., Research Department National Union Radio Corp. 350 Scotland Rd., Orange, N. J.
Monkhouse, J. R., 2535 Dunstan, Houston 5, Tex.
Negroni, J. A., Puerto Rico Water Resources Authority, San Juan, Puerto Rico
Otto, E. D., 215 C St., S. E., Washington 3, D. C.
Paynter, A. E., B.P. Radio, 754 Finchley Rd., Golders Green, London N.W. 11, England
Pike, W. S., RCA Laboratories, Princeton, N. J.
Reynolds, J. C., 22 Elm St., Calais, Me.
Ringger, W. A., Jr., Electric Operations Department, Memphis Light, Gas & Water Division, Beale & Walnut St., Memphis, Tenn.
Rolf, W. R., 41 E. Fourth, Emporium, Pa.
Rubin, I., Box 250, Red Bank, N. J.
Schnedorf, L. J., 610 Monroe Ave., River Forest, Ill.
Shepherd, M. F., Box 836, Sweetwater, Tex.
Stanton, L., 5525 Miriam Rd., Philadelphia 24, Pa.
Sweets, F. M., 1633 Beechwood Ave., Louisville 4, Ky.
Telang, V. S., 180 State St., Brooklyn 2, N. Y.
Toman, W. J., V., 1746 Fourth St., Cuyahoga Falls, Ohio
Van Horn, I. H., Jr., 1102 Roslyn Ave., Akron 2, Ohio
Waldvogel, J., 568 Fouse Ave., Akron 10, Ohio
Warren, H. R., 4631 S. Third St., Louisville 8, Ky.
Woelz, A. L., Rua Franca 83, (Jardim Europa), Sao Paulo, Brazil

The following admissions to Associate were approved on November 11, 1947; to be effective as of December 1, 1947:

Allen, T., 540 St. Johns Pl., Brooklyn 16, N. Y.
Angella, C. P., 970 S. 53 St., Richmond, Calif.
Anthes, J., 920 N. 15 St., Milwaukee 3, Wis.
Berler, R. M., 39 W. 85 St., New York 24, N. Y.
Body, H. T., c/o Siemens Brothers (Canada), Ltd., 281 McDermot Ave., Winnipeg, Man., Canada
Boff, A. F., 48 Vicars Close, Enfield, Middlesex, England
Bowman, W. B., 1630 N. Clybourn Ave., Chicago 14, Ill.
Brenner, W. F., 323 S. 43 St., Philedlphia 4, Pa.
Brown, H. W., 4506 W. Keefe Ave., Milwaukee 10, Wis.
Bryan, W. C., Box 1204, Cedar Rapids, Iowa
Buehler, E. R., Valparaiso Technical Institute, Valparaiso, Ind.
Busch, C. R., c/o Westinghouse Electric Corp. 40 Wall St., New York, N. Y.

Caballero, C. A., Obrero Mundial 207, Col. del Valle, Mexico, D. F., Mexico
Caloni, J. C., 4016 N. Kostner, Chicago 41, Ill.
Canepa, J. A., Servicio Communicaciones Navales, Darsena Norte, Buenos Aires, Argentina
Carney, W. L., Box 3, McKinley, Me.
Chambers, R. M., 4660 W. Florissant Ave., St. Louis, Mo.
Chmielewski, E. A., 200 E. Mill St., Angola, Ind.
Clayton, J. W., Electronics Material School, Treasure Island, San Francisco, Calif.

MEMBERSHIP

Concelman, C. W., Brookfield, Conn.
Couper, T. B., Jr., 11 Pleasant Court, Mayview Gardens, Maywood, N. J.
Dail, E. E., 121 Halstead Ave., Norfolk 2, Va.
Datta, A. K., Indian Engineering Institute, Connaught Place, New Delhi, India
Dillenburg, H. F., 3909 N. Seeley Ave., Chicago 18, Ill.
Di Loreto, A. A., 42 Mayhew St., Dorchester 25, Mass.
Dorf, R. H., 255 W. 84 St., New York 24, N. Y.
Edwards, G. C., 1658 Indiana Ave., Louisville 4, Ky.
Fahringer, W. H., 5436 N. Newcastle, Chicago 31, Ill.
Farber, E., c/o Novick, 4311—15 St., N. W. Washington, D. C.
Farley, J., Room 2400, 20 Exchange Pl., New York 5, N. Y.
Finley, A. D., 22 Hovey St., Watertown 72, Mass.
Foy, R. H., 5 E. Furnace Rd., Linthicum Hts., Md.
Ghose, A., Government Engineering College, Jubbulpore, India
Gleason, M. J., 7097 S. Throop St., Chicago 20, Ill.
Goldstein, S. E., 2558 Grnad Concourse, New York 58, N. Y.
Goodale, W. D., Jr., Bell Telephone Laboratories, Inc., Murray Hill, N. J.
Goslin, T. P., 506 S. Darling St., Angola, Ind.
Graham, W., 31 Evelyn Place, Nutley, N. J.
Grahame, F. W., Jr., 29 Copley Terrace, Pittsfield, Mass.
Grassia, C. C., 1114 S. Hawthorne Rd., Winston-Salem 7, N. C.
Hall, R. C., 167—10 Crocheron Ave., Flushing, L. I. N. Y.
Hanus, F. J., 1111 Niagara, Houston, Tex.
Harden, R. D., 1940 Union St., Columbus, Ind.
Hathaway, S. D., Electrical Engineering Department, Virginia Polytechnic Institute, Blacksburg, Va.
Heckart, H. A., Unionville, Iowa
Hild, A. H., 23 E. Livingston St., Valhalla, N. Y.
Hirashima, H. K., 809—13 Ave., S. E., Minneapolis 14, Minn.
Hoeflich, A. F., 626—16 Ave., San Francisco 18, Calif.
Hoff, D. L., 918 Eastwood Ave., Chicago 40, Ill.
Hollich, F. J., 1131 W. Oakdale Ave., Chicago 13, Ill.
Holmes, W. S., 90 Edgewood Ave., Kenmore 17, N. Y.
Hornyak, E. A., RCA Laboratories, Princeton, N. J.
Howell, F. N., 111 W. Main St., Middletown, Del.
Huerta, A. F., 57 Conde Penalver, Madrid, Spain
Ignatowsky, P., 218 E. 11 St., New York 3, N. Y.
Ilowite, R., 570 Westminster Rd., Brooklyn 30, N. Y.
Johnson, W. C., 717 W. Prospect St., Angola, Ind.
Kahn, H., 2053—72 St. Brooklyn 4, N. Y.
Keller, C. N., 22 Richland Ave., Dayton 5, Ohio
Kesler, C. K., 2818 Winburn Ave., Dayton 10, Ohio
Kieselbach, R., 115 W. Maple Ave., Bound Brook, N. J.
Koenreich, J. L., 215 W. 23 St., New York 11, N. Y.
Kowitz, V. M., 2259 Montrose Ave., Chicago 18, Ill.
Kuehn, R. L., 1295 Third Ave., New York 21, N. Y.
Kunik, I. J., 75 Pearl St., Hartford 3, Conn.
Langfelder, G., 31 N. Third St., Meriden, Conn.
Lavalee, M. L., 115 Central St., Manville, R. I.
LeGrand, C., 1951 Marengo Ave., South Pasadena, Calif.
Lewis, R. E., 2432 N. Lincoln Ave., Chicago 14, Ill.
Lindell, R. L., 405 Raleigh Ave., Norfolk 7, Va.
Linville, E. H., 432 W. Santa Barbara Ave., Los Angeles, Calif.
Martin, R. L., 3448 Pennsylvania St., Kansas City, Mo.
McColl, H. M., 432 S. Humphrey Ave., Oak Park, Ill.
McCullough, J. S., 502 Bradley Ave., San Jose 10, Calif.
McLean, R. J., 107½ E. South St., Angola, Ind.

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Membership

(Continued from page 43A)

Morales B., M. G., Albine Gorzo 53, Gustave A. Madero, Mexico, D. F., Mexico
Moe, W. J., 640 Holly Ave., St. Paul 5, Minn.
Montes, G. R., Oriente 53, 223 Col. Iztaccihuatl, Mexico, D. F., Mexico
Moore, C. J., 407 Oakview Ave., Joliet, Ill.
Morones, L., Concepcion Beistegui 804, Col. del. Valle, Mexico, D. F., Mexico
Most, N. S., 310 S. Wayne St., Angola, Ind.
Murray, G. E., 514 Greenleaf Ave., Willamette, Ill.
Museus, H. U., 4122 Fremont Ave., N., Minneapolis 12, Minn.
Myers, G. E., 3309 Sims Drive, Des Moines, Iowa
Nichols, L., Jr., 733 N. LaSalle St., Chicago 10, Ill.
Padula Pintos, V. H., Centro Naval, Florida 801, Buenos Aires, Argentina
Page, R. S., Box 23, Angola, Ind.
Perkins, D. W., 530 Allen St., Syracuse 10, N. Y.
Perkins, G., Jr., Department of Chemistry, Louisiana State University, Baton Rouge 3, La.
Platt, J. H., 73 England St., Valley Falls, R. I.
Porter, T. J., Rose Valley, Moylan, Pa.
Price, L. A., 576 E. Archwood Ave., Akron 1, Ohio
Prior, T. K., 166 W. Palisades Ave., Englewood, N. J.
Reed, J. C., 253 Pleasant St., Arlington, Mass.
Reeves, R. M., 6256 Bakman Ave., N. Hollywood, Calif.
Regis, R. J., 1325 B Ave., N. E., Cedar Rapids, Iowa
Reynolds, J. L., 1528 Milwaukee Ave., Chicago 22, Ill.
Rightmyer, J. M., 15825 Hesperian Blvd., San Lorenzo, Calif.
Roman, A. D., 1049 Ainslie, Chicago, Ill.
Rosen, J., Bldg. 7, Sec. K, 308 Harbor Rd., San Francisco 24, Calif.
Rosenberg, N. W., 47 Washington Village, Asbury Park, N. J.
Russell, N. E., Tri-State College, Angola, Ind.
Schneider, S., 4032 W. Melrose St., Chicago 41, Ill.
Simonsen, J., Radio Station CFRA, Ottawa, Ont., Canada
Singleton, C. M., c/o Pan American-Grace Airways, Inc., Guayaquil, Ecuador
Skitek, G. G., 401 E. 11 St., Rolla, Mo.
Stark, A. J., 1126 Spring St., Michigan City, Ind.
Strapp, J. P., 1847 W. 11 Place, Los Angeles 6, Calif.
Style, N. V., 334 W. 73 St., Los Angeles 3, Calif.
Sunderland, H. L., Electronics Section, AGF Board No. 4, Fort Bliss, Tex.
Symanovich, T., 194-F E. 99 Drive, Brooklyn 12, N. Y.
Taylor, G. H. R., 52 Holtspur Top Lane, Beaconsfield, Bucks, England
Teal, E. E., 3031 Cason, Houston 5, Tex.
Teslik, J. A., 4527 N. Malden, Chicago 40, Ill.
Thomas, H. G., American Television Laboratory, Inc., 5050 N. Broadway, Chicago 40, Ill.
Thomas, J. N., Engineering Department, George Washington University, Washington 6, D. C.
Thomas, R., 7750 Essex, Chicago 49, Ill.
Thrasher, P. M., Jr., 207 Dewitt St., Syracuse, N. Y.
Tognazzini, M. O., 443 Ramsell St., San Francisco 12, Cal.
Turnow, W. F., 95-12—125 St., Richmond Hill 19, L. I. N. Y.
Ule, L. A., 5216 N. Winthrop Ave., Chicago 40, Ill.
Vega, E. J., Dr. Martinez del Rio 93, Int. 19, Mexico, D. F., Mexico
Vick, E. K., 2739—First Ave., S. E., Cedar Rapids, Iowa
Vinyard, E. L., Box 42, Highland Park, Ill.
Watkins, D. A., 1608½ Washington Ave., S. E., Cedar Rapids, Iowa
Welch, F. R., 106 Pleasant Ave., Dayton 3, Ohio
Wescott, J. C., Box 481, Valparaiso, Ind.
White, M. F., Electronic Process Heat Treating Co 465 W. 162 St., New York 32, N. Y.
Wholey, C. W., 1408—19 Ave., San Francisco 22, Calif.
Wilson, L. F., 226 Scarsdale Rd., Yonkers, N. Y.
Wong, B. G., 522 W. Third St., Fort Wayne, Ind.
Zimmerman, W. L., Electronics Material School, Treasure Island, San Francisco, Calif.
Zwarra, H. J., Wisconsin Telephone Co., 722 N. Broadway, Milwaukee 2, Wis.

News—New Products

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

Type RC-108 Capacitor

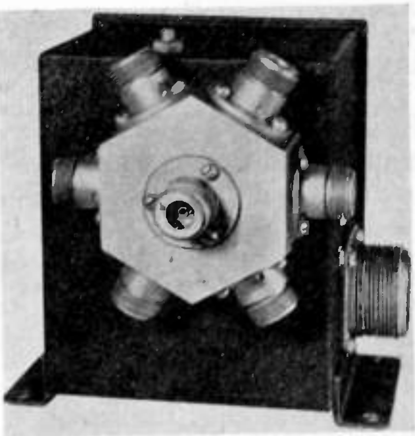
Recently introduced by **Cornell-Dubilier Electric Corp.**, So. Plainfield, N. J., as an addition to their line of television capacitors, the new type RC-108 capacitor with a rating of 0.05 μfd., 3500 volts d.c., is made with a very high safety factor.

The Type RC-108 capacitor is built in a cylindrical metal container $1\frac{1}{2}$ inches in diameter, 3 inches long, with screw-type terminals mounted on ceramic insulators protruding $\frac{3}{4}$ inch from each end of the case. A wax-impregnated cardboard sleeve provides insulation.

It is claimed that Dykanol C impregnation and the metal hermetical seal assure efficient operation under all atmospheric conditions; and that this new capacitor has very stable characteristics assuring long life at elevated temperatures.

New Coaxial Switch

Designers for Industry, Inc., 2915 Detroit Ave., Cleveland 13, Ohio, are now supplying a series of new coxial switches for v.h.f. and u.h.f. applications.



Pictured here is the type D, a six-position switch for use with RG-8/U r.f. cable. Voltage-standing-wave ratio is less than 1.5 to 1; adjacent-channel attenuation is better than 50 db.

These switches are remotely controlled and operated from 117-volt 60-cycle a.c. Provision is made for the operation of position-indicating pilot lights at the central point.

The manufacturer announces that other switches are available for use with the larger sizes of coxial cable.

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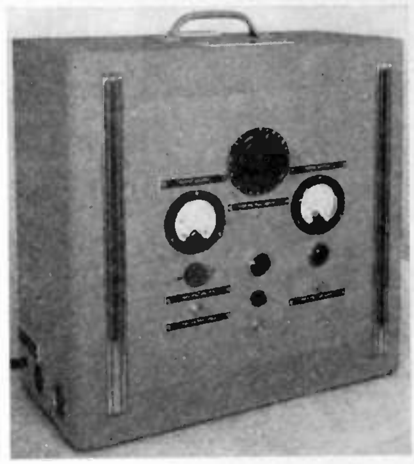
News—New Products

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)

Portable Power Supply



The Model 201 portable 0–30 kv. d.c. power supply, illustrated above, finds applications in television work, cathode-ray oscillography, flash photography, electronic precipitation, high-voltage insulation, or wherever compactness, portability and adjustability are desired. Availability of this unit was recently announced by the manufacturer, **Beta Electronics Co.,** 1762 Third Ave., New York 29, N. Y.

The maximum load current of this instrument at 30 kv. is 300 microamperes, dropping off to about 23 kv. at 1.5 milliamperes. Currents up to 2 milliamperes may be drawn. Output ripple at 30 kv. is claimed to be less than 2 per cent. Approximately 200 volt-amperes are drawn from a 115-volt 60-cycle line, at maximum output voltage.

With a cabinet measuring 16×16×8 inches, the power supply includes an output kilovoltmeter, a Variac current meter, filament and power-on indicating lamps, and relay switching circuits for permitting either manual or automatic "high-voltage-on" control.

A current-limiting resistor is included in the output circuit to limit the surge current to a safe value, in case of flashover of the load. The sustained short-circuit current is limited to a safe value by inherent circuit regulation.

Selenium Rectifier

Especially designed for relays and low-current control applications where space is limited, a new selenium rectifier, known as Model SE-8M20F, has been announced by **Bradley Laboratories, Inc.,** 82 Meadow St., New Haven, Conn.

Rated at 110 volts a.c., 80 volts d.c., 10 ma. d.c., the unit can be modified to meet different electrical specifications. The rectifier has a tubular bakelite case $\frac{3}{8}$ inch in diameter by $1\frac{1}{2}$ inches long, with four 2-inch tinned leads. It mounts on two standard screws and is completely sealed against moisture and corrosive atmosphere.

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News—New Products

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

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PM Speakers

Designed for the best radio and general-purpose loudspeaker applications, the **William J. Murdock Co.**, 158 Carter St., Chelsea 50, Mass, announce that L301, (3-inch) and L401 (4-inch) pulse-modulation speakers are available for original or replacement use.

These speakers use 1.47-oz. Alnico V magnets, an unusually sturdy frame, and a preformed speaker cone and curves verify the fine response characteristics. The standard speakers are of RMA 3.2-ohm impedance. Special impedances are available.

Recent Catalogs

• • • On a pocket-size signal generator producing r.f., i.f., and a.f. signals simultaneously from approximately 2500 c.p.s. through 20 Mc., known as the "Signalette", by **Clippard Instrument Laboratory, Inc.**, 1125–33 Bank St., Cincinnati 14, Ohio.

• • • On radio wire products including electronic and intercommunicating wires and cables, illustrated catalog No. 55, by **Cornish Wire Co., Inc.**, 15 Park Row, New York 7, N. Y.

• • • On technical data, including a table showing impedance vs. decibel loss with values calculated for impedance mismatch, minimum tee loss, and bridging-pad loss, published by **The Daven Co.**, 191 Central Ave., Newark 4, N. J.

• • • On special ceramic materials, giving a description of the various types of ceramic bodies, their physical properties, and their applications in the different fields of engineering, by **General Ceramics and Steatite Corp.**, Keasbey, N. J.

• • • On high-voltage resistors, a four-page technical data bulletin giving specifications and characteristics of Type MV high-voltage resistors, by **International Resistance Co.**, 401 No. Broad St., Philadelphia 8, Pa. Ask for Bulletin G-1.

News—New Products

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

Recent Catalogs

• • • On GCA (ground control approach), by Philco Service Division, **Philco Corp.**, Tioga and C Sts., Philadelphia 34, Pa. Ask for Brochure PR-1473.

• • • On crystals, a four-page illustrated bulletin giving technical data covering twelve widely used crystal types, issued by **Premier Crystal Laboratories, Inc.**, 57-67 Park Row, New York 7, N. Y. Write for Bulletin No. 201.

• • • On self-generating photoelectric cells, a twelve-page illustrated brochure giving standard specifications, characteristics, applications and design factors, by **Selenium Corp. of America,** 2160 East Imperial Highway, El Segundo, Calif.

• • • On electronic relay Type EA.2 (List EA.10) and adjustable time-delay switch units, Types TYE and TYD (List TD.10), and other electronic control instruments, by **Sunvic Controls, Ltd.**, 10 Essex St., Strand, London WC.2, England.

• • • On mercury-contact relays, a twelve-page illustrated catalog giving operating characteristics and other technical data regarding Type 275 and Type 276 relays for use in such devices as computing machines, signalling devices, servomechanisms, high-speed keying relays, relay amplifiers, and vibrator power supplies. Issued by **Western Electric Co.**, 195 Broadway, New York 7, N. Y. These relays are distributed in the U. S. A. by Graybar Electric Co.

Vacuum Phototube

The RCA-5653 (glass-octal type) is a new vacuum phototube intended especially for light-operated relay and other applications where there is always plenty of incident light and where a wider-than-usual range of luminous sensitivity may be tolerated.

The availability of this new vacuum phototube has recently been announced by the Tube Department, **Radio Corporation of America**, Harrison, N. J.

Having S-4 response, the 5653 is particularly sensitive to blue radiation, but has good response to light from an incandescent lamp.

For applications requiring more critical performance, the 929 or the 1P39 is recommended.

(Continued on page 54A)

WANTED PHYSICISTS ENGINEERS

Engineering laboratory of precision instrument manufacturer has interesting opportunities for graduate engineers with research, design and/or development experience on radio communications systems, electronic & mechanical aeronautical navigation instruments and ultra-high frequency & microwave technique.

WRITE FULL DETAILS

TO

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Lake Success, L.I.

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PhD—Physicist, Math. or Electronic, Major.
MSME—Designer, small electro-mechanisms.
MSEE—Radiation Lab. or Equivalent.
BSEE—Electronic Engr.; Servos.
BSEE—Electronic Engr.; pulse circuits.

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   Opportunity to have job fit YOU.
2. **Quicker delegation of responsibility.**
   Opportunity for YOU to develop.
3. **Greater variation in problems.**
   Opportunity for YOU to choose.
4. **Quicker advancement.**
   Opportunity to increase YOUR income.

If you are alert to these advantages, please send résumé of qualifications to Box 498.

The Institute of Radio Engineers
1 East 79th Street New York 21, N.Y.

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.

ENGINEERING PROFESSORSHIPS

Outstanding technical school in Chicago has openings in radio, industrial electronics, and electric power engineering fields. Unusual opportunities can be offered men possessing desirable industrial, research or teaching experience. Write giving field of interest and outline experience. Box 488.

ENGINEERS

Microwave engineers wanted. Laboratory experience essential (industry or government). Positions of junior engineers, engineers, senior engineers. Permanent. Salary relatively high. Video men also wanted. You are invited to visit our modern plant and talk to our engineers, or write us your job history and education. Motorola, Inc., 4545 W. Augusta Blvd., Chicago 51, Illinois. Att: Mr. E. Dyke.

ENGINEERS, PHYSICISTS, MATHEMATICIANS

To fill 10 positions on seismograph field parties throughout the Rocky Mountains, mid continent and gulf coast states. Duties consist of operating seismic recording instruments, or computing seismic data, or alidade surveying seismic locations. Nature of work requires several changes of address per year; part of it is outdoors and part indoors; certain operations performed under standard procedure, others require ingenuity and initiative; salary \$200-\$300 per month to begin, with excellent opportunity to advance for those with practical ability. To apply, write giving scholastic and employment background, age, nationality and family status to Box 490.

JUNIOR ENGINEERS

Microwave research and other advanced radio work, requiring college degree and natural aptitude. Opportunity for valuable experience and advancement in a small growing organization. Suburban location on Long Island near New York City. Send personal record to Harold A. Wheeler, Wheeler Laboratories, Inc., Great Neck, N.Y.

PHYSICISTS, RESEARCH ENGINEERS, TECHNICIANS

Growing research and manufacturing concern in suburban Philadelphia, specializing in multi-gun cathode ray tubes, has attractive openings, particularly for those experienced in vacuum tubes, photo surfaces and electron optics. Electronic Tube Corporation, 1200 E. Mermaid Avenue, Chestnut Hill, Philadelphia 18, Pa.

PRODUCTION DESIGN ENGINEER

Engineer, preferably with radio-phonograph mechanical design background, capable of producing practical, low cost, mass production designs starting from performance specifications. The work involves specification for purchase of components, establishing of inspection and quality standards, coordination of appearance styling, and follow-up of initial production. Reply giving a brief résumé of personal data, educational background, and details of type of product worked on, and extent of responsibility therefore, over the past ten years. Box 492.

ELECTRONIC THEORIST

Our New York laboratory is seeking an Electrical Engineer or Physicist to carry on theoretical investigations of problems associated with vacuum tubes, thermionics and microwave equipment and to interpret theoretical developments in terms of experimental results. MS or equivalent in experience in field of thermionics and microwave engineering desired. Send résumé outlining age, education, experience, salary requirements to: Supervisor of Employment, Industrial Relations Department, Sylvania Electric Products, Inc., 40-22 Lawrence Street, Flushing, N.Y.

MECHANICAL DESIGN ENGINEER

Having experience in quantity production of small metal stampings and component assemblies. Pleasant working conditions with electronics equipment manufacturer in small Minnesota town. Box 493.

ELECTRONIC ENGINEER—PHYSICIST

A major oil company in the southwest requires services of competent Physicists and Electronic Engineers as permanent

ELECTRONIC DESIGN ENGINEERS

Openings exist for engineers and physicists with college degrees or equivalent to work on electronic apparatus of various types, including microwave communications, radar, and electronic control equipment. Experience desirable. Salaries commensurate with education and experience. Communicate with Westinghouse Electric Corporation, Industrial Relations Department, 2519 Wilkens Avenue, Baltimore 3, Maryland.

POSITIONS OPEN

research members. Positions available for project engineers and group leaders. Prefer men with Ph.D. or equivalent. Work involves research in field of physics, physical chemistry, and geophysical exploration, development of analytical instruments and equipment. Applicant should have training and experience along theoretical and experimental lines. These positions are permanent and offer unusual opportunities for the right men. Give complete details—personal history, education, experience, and salary desired. Applications treated confidentially. Box 494.

MASS SPECTROMETRY

Engineer with advanced degree and experience in electronics, ion-optics, and high-vacua techniques to take charge of long term program in development and research in field of mass spectrometry at an eastern university. Salary $5000-$8000. Box 495.

TELEVISION TRAINEES

Opportunity with National Broadcasting Company for graduate engineers major in communications. 20 to 30 years of age. 18 months intensive training prior to placement on regular staff. Apply to Personnel Dept., National Broadcasting Co., 30 Rockefeller Plaza, New York 20, N.Y. by letter only. No interviews in person.

ELECTRONIC CIRCUIT ENGINEERS

For design, construction and test of electronic circuit components and systems in forms suitable for field operation of a complete electronic field installation. Ingenuity, imagination and capable theoretical inclinations suitable for research laboratory work are desired. Send résumé outlining age, education, experience and salary requirements to: Supervisor of Employment, Industrial Relations Department, Sylvania Electric Products, Inc., 40-22 Lawrence Street, Flushing, New York.

ELECTRICAL ENGINEER (ELECTRONICS OPTION)

Young recent graduate with E.E. degree to design communication equipment, special electronic instruments, electronically controlled automatic equipment involving servo-mechanisms for pipe line company affiliated with large oil company. 1st class commercial radio license desirable. Some travel involved in summer months. Winter months in laboratory. Location Eastern Pennsylvania. Appearance, personality and ability to work with small group of engineers important. State age, experience and salary requirements. Box 497.

ELECTRONIC ENGINEERS

Unusual opportunities for senior engineers experienced in: Recording sound on film and magnetic tape recording; Microwave—antennae, wave guides, mixers, cavity resonators; Receiver Engineer—design experience in broadband receivers. Radar preferred. Two or three years experience in U.H.F. work. Outstanding opportunity for top flight men with a small aggressive company. Write Melpar, Inc., Employment Section, 452 Swann Avenue, Alexandria, Virginia.

(Continued on page 52A)

POSITIONS OPEN

(Continued from page 51A)

ENGINEER OR PHYSICIST

Engineer or physicist for mathematical research work on vacuum tubes. Should have a good knowledge of microwave tubes and electron optics. Apply Director of Research, 350 Scotland Road, Orange, New Jersey, National Union Radio Corporation.

DEVELOPMENT AND PRODUCTION ENGINEER—RADIO CONTROLS

Requires considerable experience in receivers and transmitter (V.H.F.) Development and production to specialize on radio controls. Requires an E.E. degree or equivalent (minimum of 6 years experience in development and production engineering of radio and/or electronic equipment) at least 4 years in competent industry. Mr. A. F. Malmquist, Personnel Director, Pacific Division, Bendix Aviation Corp., 11600 Sherman Way, North Hollywood, California.

DEVELOPMENT AND PRODUCTION ENGINEER—SONAR EQUIPMENT

Requires E.E. degree or equivalent, minimum of 6 years experience in development and production engineering of which at least 4 years has been with competent industry, preferably with marine and sonar experience. Requires ability to handle mechanical design related to electronic design. Apply Mr. A. F. Malmquist, Personnel Director, Pacific Division, Bendix Aviation Corp., 11600 Sherman Way, North Hollywood, California.

SCIENTISTS AND ENGINEERS

Wanted for research and advanced development work in the fields of microwaves, radar circuits, gyroscope systems and general electronics. Scientific or engineering degrees required. Salary commensurate with experience and ability. Inquiries should be directed to: Mgr.—Engineering Personnel, Bell Aircraft Corp., Buffalo 5, N.Y.

JUNIOR ENGINEER—Editor and Writer

Excellent opportunity for experienced writer in radio and electronic fields to edit technical publication and handle articles for electronic, broadcast, aviation and amateur radio press. Congenial surroundings in attractive midwest city. Please give full particulars as to background, experience, age and salary in first letter. Collins Radio Company, Cedar Rapids, Iowa.

EXECUTIVE ENGINEER

This notice is inserted by a large manufacturer of radio and television receivers. We have an opening for the one right top executive engineer who really belongs in such an organization. The position is that of heading up all phases of design, development, and research engineering. It is probably the toughest engineering chief job in the industry. It pays $25,000. Do not waste your time and ours unless you are unqualifiedly sure that you are completely ready for this assignment. Our own key personnel are aware of this announcement. Write to Box 499.

★ ★ ★ ★

Positions Wanted By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants, and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The Institute publishes free of charge notices of positions wanted by I.R.E. members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The Institute necessarily reserves the right to decline any announcement without assignment of reason.

ELECTRONIC ENGINEER

B.S.E.E. Northeastern university in September 1947. Age 23. 1½ years experience with all types of Naval Airborne radio and radar equipment. Hold 1st class radio-telephone license. Member Tau Beta Pi. Desires position as Junior Engineer in electronic design research or development. Further details on request. Box 113W.

ENGINEER

Schools—N.C.E., Harvard and M.I.T. Flying Air Corps officer. Presently engineer in development laboratory. Familiar with radio, radar, G.M., microwave techniques. Desires industrial engineering position in laboratory or plant. Box 114W.

JUNIOR ENGINEER

B.E.E. 1947, Polytechnic Institute of Brooklyn. Age 30. Married. One child. Two years Army Radar officer, Harvard —M.I.T. radar school. Eta Kappa Nu. Desires position as a junior engineer in electronic design, development. Anywhere in U.S. Box 118W.

ENGINEER (CANADIAN)

B.S.E.E. 1939. Six years experience in maintenance and installation of Naval radar and radio equipment. Last 3 years in administration and supervision. Present rank Lieutenant Commander (Electrical). Licensed amateur since 1932. Age 30. Married. One child. Interested in engineering, sales or representative position particularly in maritime provinces or Newfoundland. Box 120W.

ELECTRONICS ENGINEER

B.E.E. Drexel Institute, 1936. 11 years experience in radio transmitter design, electronic control circuits, bridge, oscillator, amplifier and null detector design. Investigation of foreign electronic equipment. Also new short wave therapy circuits. Box 123W.

JUNIOR ENGINEER

B.S.E.E., B.S.M.E., 1939, University of Paris. Age 28. Completing graduate work E.E. electronics at B.P.I. New York. 2½ years Army Signal Corps experience on RDF and communications equipment. Seeks position as junior engineer, physicist or instructor in New York City area. Box 130W.

(Continued on page 54A)

Positions Wanted

(Continued from page 53A)

TECHNICAL EDITOR

Former AAF officer with four years experience in writing and editing technical project reports and summaries, budget defenses, press releases, technical papers, etc. Assigned during this period to Radiation Laboratory, M.I.T. and Aircraft Radio Laboratory, Wright Field. Additional experience as sales engineer (2 years), radio and radar technician (2 years), and two years of college credits towards E.E. degree. Box 131W.

ELECTRONIC ENGINEER

B.E.E. Ohio State University, June 1948. Two years as electronics Technician, U. S. Navy. Would like position in research or development. Vicinity New York or Cleveland. Box 136W.

RADIO ENGINEER

B.E.E. from N.Y.U. 1947; Graduate work B.S. in physics from C.C.N.Y. 1943. Signal Corps radio and repeater man telephone experience. Desire design and development work in radio, industrial electronics or communications. Box 137W.

ELECTRICAL ENGINEER

B.S. in E.E. 1944, University of New Brunswick (Canada). M.S. 1947, University of Western Ontario. Age 23. Single. Former officer Royal Canadian Signals. Limited experience in telephone work. Desires electrical engineering employment abroad, preferably in tropics. Box 138W.

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)

Type 125 Capacitance Bridge

A capacitance bridge suitable for measurement of capacitance in multi-electrode systems has been announced by the Electronics Division, **Sylvania Electric Products, Inc.,** 500 Fifth Ave., New York 18, N. Y. The instrument, particularly useful in measuring interelectrode capacitances in vacuum tubes, provides a range of 0 to 100 $\mu\mu$fd. through the use of five multipliers and measurement at 465 kc. Direct-capacitance accuracy of 1 per cent and direct-conductance accuracy of 10 per cent are provided when calibrated with standards of commensurate accuracy.

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

The bridge consists of three separate sections including r.f. signal generator and power supply; r.f. amplifier, detector and vacuum-tube voltmeter; and associated switches, controls and, 500-μa. meter indicating bridge balance. Ground to lead or jig capacitance may be tuned out when combined values do not exceed 25 $\mu\mu$fd. Special tube-tester adapters are available for measurements with or without base shield or metal shell connected to the tube element or ground.

Type 125 is rated at 45 watts, 110-120 volts, 50-60 cycle a.c.; weighs 50 pounds; and measures 19 inches long by 12½ inches high.

Mega Match

Recently announced by **Kay Electric** Co., 34 Marshall St., Newark 2, N. J., a new electronic, basic laboratory instrument for measuring reflected energy is known as the "Mega Match."

This instrument measures reflected energy over a wide frequency band, 10 to 250 Mc. and up. It presents a visual display of reflected energy over any band up to 30 Mc. This visual display eliminates tedious tabulation work, saving hours of engineering time.

The manufacturer states that it is possible to instantly observe and measure mismatches. There are no slotted lines, moving parts, directional couplers, or other frequency-sensitive devices. A precision frequency meter is incorporated in the unit.

Applications include measuring antennas, transmission lines, and input and output impedances. It is claimed that no other commercial instrument gives a complete visual display of reflected energy from the above-mentioned devices.

New Enterprise

• • • The **Precision Manufacturing Co.**, of Bergholz, Ohio, a new division of the Alliance Manufacturing Co., is now producing phonograph turntables on a large scale, according to an announcement by the Progressive Welder Company of Detroit which supplied the battery welders used in the manufacture of these turntables.

(Continued on page 56A)

News—New Products

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

(Continued from page 55A)

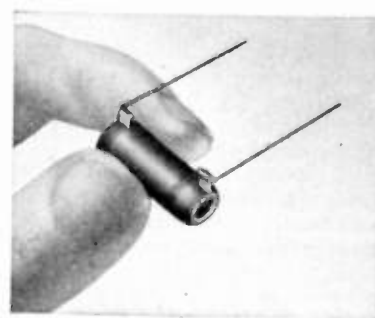
Light-Weight Distortion Meter

A new distortion meter, Model 400, recently introduced by **Barker & Williamson, Inc., 237 Fairfield Ave., Upper Darby, Pa.,** is an ideal unit for either laboratory or field use because of its small size and light weight. Dimensions $13\frac{3}{4}\times7\frac{1}{2}\times9\frac{1}{2}$ inches. Weight, $11\frac{1}{2}$ pounds.

This meter is suited for measuring low-level audio voltages and determining their noise and harmonic content, and may also be used in measuring frequency and gain characteristics of audio amplifiers, or for any other application where a vacuum-tube voltmeter is required in the audio range. The variable-frequency selective filter provides a single-frequency suppression circuit for the frequency range of 50 to 15,000 cycles. Frequency range as voltmeter and db. meter is from 30 to 30,000 cycles.

Wire-Wound Resistors

To supplement their *Brown Devil* line of 10- and 20-watt enameled wire-wound resistors for radio, electronic, and industrial circuits, **Ohmite Manufacturing Company, 4952 Flournoy St., Chicago, Ill.,** has added a compact 5-watt size.

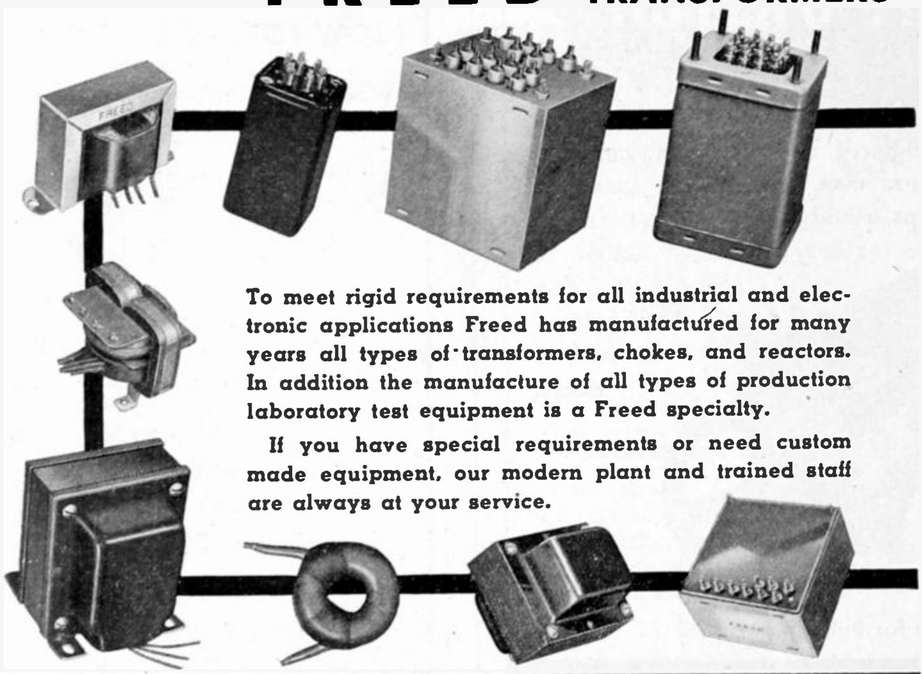


Although this small unit has been available on special order, the 5-watt size ($\frac{5}{16}$ by 1 inch) may now be obtained from regular stock in resistance values from 1 to 10,000 ohms. Standard tolerance is ±10 per cent. The new 5-watt resistors are of all-welded construction and have $1\frac{1}{2}$-inch copper-wire leads. Write to the manufacturer for Bulletin 132.

(Continued on page 59A)

COMMUNICATIONS—ELECTRONICS—TELEVISION

for DEPENDABLE performance FREED TRANSFORMERS



To meet rigid requirements for all industrial and electronic applications Freed has manufactured for many years all types of transformers, chokes, and reactors. In addition the manufacture of all types of production laboratory test equipment is a Freed specialty.

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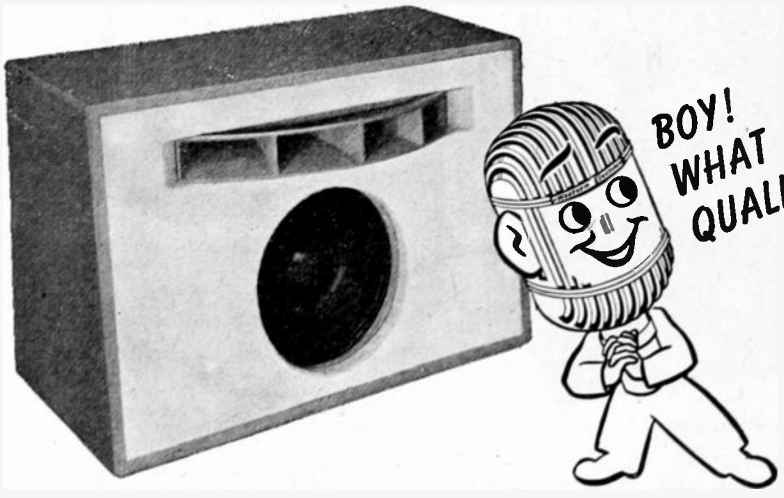

Western Electric

757A LOUDSPEAKER

With uniform response from 60 right up to 15,000 cycles—a 90 degree coverage angle—power handling capacity of 30 watts—this is *THE* speaker where highest quality in sound reproduction is a must!

The 757A is just one of the complete line of new high quality speakers—from 8 to 120 watts—developed by Bell Laboratories and made by Western Electric.

For full details, write today to Graybar Electric Co., 420 Lexington Ave., New York 17, N. Y.—or...

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BOLAND & BOYCE INC., PUBLISHERS

News—New Products

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

(Continued from page 56A)

Interesting Abstracts

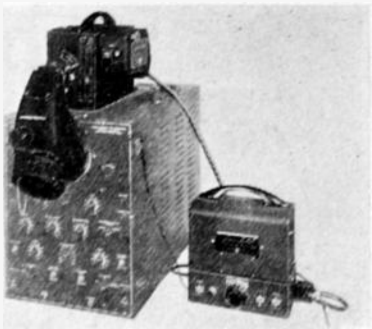
• • • The "Frequency Standard, Tuning-Fork Type" which was described on page 44-A of the October, 1947, issue of the PROCEEDINGS, covers any frequency in the range from 200 to 1000 cycles (not 200 to 100 cycles, as stated in the description). This unit is manufactured by **American Time Products, Inc.,** 580 Fifth Ave., New York 19, N. Y.

• • • As announced by **John F. Rider,** publisher, 404 Fourth Ave., New York 16, N. Y., a new book on "F.M. Transmission and Reception" includes pictorial representations of f.m. as well as phase modulation, with the fundamental theory simplified. It contains more than 300 pages and is available in two bindings at $1.80 and $2.70.

• • • A recent announcement from **Sound Apparatus Co.** 233 Broadway, New York 7, N. Y., manufacturer of graphic recorders and vacuum-tube voltmeters, indicates an expansion of their facilities in the establishment of a Canadian representative. Harris Pound, 2235 Addington Ave., Montreal 28, P.Q., Canada, will be in charge of Canadian distribution of their products.

Oscillo-Record Camera

Operated on electronic principles, the new oscillo-record camera illustrated below was designed by **Fairchild Camera and Instrument Corp.,** 88-06 Van Wyck Blvd., Jamaica 1, N.Y., for general-purpose use in recording oscilloscope traces.



Equipped for mounting atop standard laboratory oscilloscopes, this 35-mm. camera, which makes still or continuously moving film records, photographs high-speed phenomena as well as very low-speed phenomena (too low for visual continuity). It may also be used for quantitative studies of oscilloscope traces, for record purposes, and for tests using new multiple-beam tubes.

This instrument operates from 20 seconds at maximum speed to 20 hours at minimum speed, with 100-foot rolls; and from 3½ minutes to 200 hours, with the 1,000-foot magazine. A footage indicator shows the number of feet of film exposed.

(Continued on page 60A)

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

(Continued from page 59A)

New Cathode-Ray Tube

A newly developed NORELCO cathode-ray tube, Type 3QP1, for oscilloscope use is very short, has a flat face, and provides improved electron-optical characteristics, particularly at the screen edge. The tube has improved cross-talk characteristics between deflection-plate pairs, and is suited to the design of very small, light-weight service equipment needed in television installation and maintenance work, according to an announcement by **North American Philips Co., Inc.**, 100 East 42 St. New York 17, N.Y.

Over-all length of the 3QP1 is only $6\frac{3}{8}$ inches, and the face diameter is $2\frac{3}{4}$ inches. The tube utilizes P1 (green) phospor and has electrostatic focus and deflection. Rated heater drain is low; 0.3 amperes at 6.3 volts. Capacitance between terminals varies between 2 and 9 $\mu\mu$fd.

Miniature Receiving Tubes

Three new nine-pin miniature tubes, Types 6T8, 19T8, and 12AT7, have been developed especially for use in f.m. and television receivers by the Tube Division, Electronics Dept., **General Electric Co.**, at Schenectady, N.Y.

The 12AT7 is a miniature-type twin triode designed for use as a grounded-grid r.f. amplifier or as a frequency converter at frequencies below approximately 300 Mc. A center-tapped heater permits operation of the tube from either a 6.3-volt or a 12.6-volt heater supply. Both the triple-diode-triode 6T8 and 19T8 contain three high-perveance diodes and a high-μ triode in the same envelope. One of the diodes has a separate cathode connection. The tubes are designed for use as combined a.n. and f.m. detectors and a.f. amplifiers. The heater of the 6T8 is designed for 6.3-volt operation at 450 ma. while the 19T8's heater is built for 18.9-volt operation at 150 ma.

These tubes are $\frac{7}{8}$ inch wide by $2\frac{1}{16}$ inches high.

(Continued on page 62A)

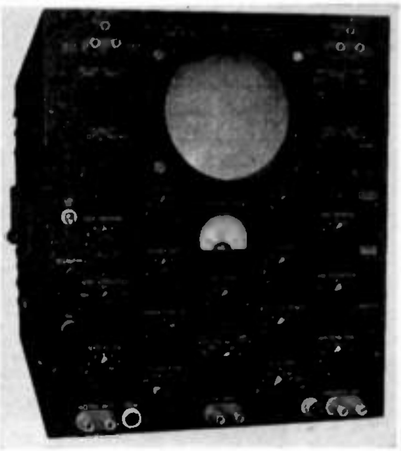
News—New Products

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)

General-Purpose Oscilloscope

Incorporating the latest circuits, a new omnipurpose 5-inch oscilloscope, known as Model OL-15A, has been designed by **Browning Laboratories, Inc.,** 750 Main St., Winchester, Mass.



Among the features of this instrument is the response curve of the vertical amplifier, which is linear and without positive slope from 10 cycles to 4 Mc. Thus, the transient response is such that a 100-kc. square wave which rises or falls in the order of 500 volts per microsecond is faithfully reproduced. The horizontal amplifier response extends linearly from 10 cycles to 1 Mc. to accommodate any type of externally generated sweep voltage which one may wish to employ. The sawtooth sweep range is from 5 cycles to 500 kc. with synchronizing sensitivity permitting syncing and viewing 10Mc. r.f. sine waves.

Because of the versatility of the OL-15A oscilloscope, the manufacturer will undertake to advise those interested in the adaptation of the instrument to their particular measurement or viewing problem. Its dimensions are 15¾ 12¾ 19¾ inches.

High-Speed Recording Paper

Developed for use with galvanometer and cathode-ray oscillographs, a new photographic recording paper has been announced by **Eastman Kodak Co.,** Rochester, N.Y.

Known as Kodak Linagraph 1127 Paper it is claimed to be the fastest of its type now made. It is more than twice as sensitive to blue light as existing high-speed recording papers and between one-third and one-half as sensitive as various recording films.

(Continued on page 66A)

Do Radio Engineers Know What You Make?

• **They Need Specifications**—for Radio and Electronic Engineers control the technical buying of a two-billion-dollar industry. These men alone are competent to set specifications for, and authorize the purchasing of complicated equipment, instruments, tubes, materials and components that only a trained and experienced electronic engineer understands. These men are the key to your sales—and need the product specifications your advertising provides.

• **The Market**—is over 17,000 qualified radio engineers, working in 3000 manufacturing firms, radio and communication stations, engineering research laboratories, government bureaus and services and in production control divisions of large plants. They are the members of The Institute of Radio Engineers, selected both by stiff membership requirements at the beginning, and high enough dues through the years to insure active and interested readers.

• **Why Engineers will use this Directory!** The true value of reference advertising in a directory is "constant use and service." This is a source-reference-book edited by engineers for engineers and it lists engineers as well as firms and products. I.R.E. members will find themselves and their friends listed both alphabetically, and geographically in "The I.R.E. Yearbook." Personal interest is the key to keeping and using The I.R.E. Yearbook.

• **3 Directories in 1**—Not only is it the only published personnel list of radio and electronic engineers, but combined in the same covers and always at hand is an alphabetical list of nearly 3000 firms supplying the industry, with code-keys to their products. In addition there is a product index for advertisers only. This classifies in 100 fundamental and understandable groups set up by engineers, the products, instruments and materials they need.

RATES			For "Proceedings of I.R.E. Contract Advertisers"	
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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 3,000 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.

Check off the appropriate items and send this questionnaire today: Industry Research Div., Proceedings of the I.R.E. Room 707, 303 West 42nd St., New York 18, N.Y.

Please read carefully and fill out completely.

PERSONNEL INFORMATION

In addition to data regarding products and/or services you render the radio-electronic field, please supply information requested below.

Firm Name .... Chief Engineer ....

Street .... Town .... Zone .... State ....

The proper person to receive information and announcements—

On Product Data .... Title ....

On Advertising .... Title ....

Classifications to Be Checked by Non-Manufacturing Firms Rendering Services to the Radio-Electronic Field

( ) 501. Books & Book Publishers.
( ) 502. Broadcasting Stations & Communication Cos.
( ) 503. Consulting Engineers:
  ( ) a. Acoustical.
  ( ) b. Electrical.
  ( ) c. Mechanical.
  ( ) d. Radio.
( ) 504. Distributors & Jobbers of Radio & Electronic Equipment & Supplies.
( ) 505. Frequency Measuring & Monitoring Services.
( ) 506. Laboratories & Custom Builders of Equipment.
( ) 507. Recording Studios & Services.
( ) 508. Technical Schools & Institutions.
( ) 509. Transcription Libraries.

Wholesale Radio Dealers, see Distributors & Jobbers.

Classification If Not Listed

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Products to Be Checked by Radio-Electronic Manufacturers

( ) 1. Aircraft & Airport Radio Equipment.
( ) 2. Amplifiers, Audio Frequency.
( ) 3. Antennas:
  ( ) a. AM Broadcast Transmitting.
  ( ) b. Dummy.
  ( ) c. FM Broadcast Transmitting.
  ( ) Miscellaneous Types.
  ( ) e. Receiving, all services.
  ( ) f. Television Transmitting.
( ) 4. Antenna Phasing Equipment & Accessories:
  ( ) a. Feeder Lines & Accs.
  ( ) b. Phasing Equipment.
  ( ) c. Tower Lighting Equipment.
( ) 5. Attenuators.
( ) 6. Batteries:
  ( ) a. Dry "A".
  ( ) b. Dry "B".
  ( ) c. Dry Miniature.
  ( ) d. Wet Primary.
  ( ) e. Wet Secondary.
( ) 7. Blowers & Cooling Fans.
Bridges, see Testing Equip.
( ) 8. Cabinets, Wooden.
( ) 9. Cables:
  ( ) a. Coaxial.
  ( ) b. Microphone.
  ( ) c. Pre-Formed Harnesses.
  ( ) d. Shielded.
  ( ) e. Ultra-High Freq.
( ) 10. Capacitors, Fixed:
  ( ) a. Ceramic.
  ( ) b. Electrolytic.
  ( ) c. Mica.
  ( ) d. Oil Filled.
  ( ) e. Paper.
  ( ) f. Pressurized Gas.
  ( ) g. Vacuum.
( ) 11. Capacitors, Variable:
  ( ) a. Neutralizing.
  ( ) b. Precision.
  ( ) c. Temperature-Freq. Compensating.
  ( ) d. Trimmer.
  ( ) e. Tuning.
  ( ) f. Vacuum.
( ) 12. Ceramics:
  ( ) a. Coil Forms.
  ( ) b. Ground Mica.
  ( ) c. Rods.
  ( ) d. Sheets.
( ) 13. Chassis & Relay Rack Cabinets, Metal.
Coil Forms, see Ceramics.
( ) 14. Coils:
  ( ) a. Chokes, AF & RF.
  ( ) b. Miscellaneous Types.
  ( ) c. Toroids.
  ( ) d. Transformer Coils.
  ( ) e. Tuning.
Condensers, see Capacitors.
( ) 15. Connectors.
Consoles, see Amplifiers.
( ) 16. Converters:
  ( ) a. Frequency.
  b. *Rotary, see Motor Generators.*
  ( ) c. Vibrator.
( ) 17. Core Materials:
  ( ) a. Complete Cores.
  ( ) b. Laminations.
  ( ) c. Powdered Iron.
( ) 18. Crystals:
  ( ) a. Oscillating Quartz.
  ( ) b. Piezo-Electric.
  ( ) c. Rectifier.
Discs, Recording, see Recording Equipment.
( ) 19. Drafting Equipment & Supplies.
Dynamotors, see Motor Generators.
( ) 20. Electronic Control Equip.:
  ( ) a. Air Conditioning Controls.
  ( ) b. Burglar Alarm & Protection Devices.
  ( ) c. Combustion & Smoke Control Equip.
  ( ) d. Fire Prevention Eq.
  ( ) e. Photo-Electric Control Devices.
  ( ) f. Production Controls, Counting & Sorting Equipment.
  ( ) g. Variable Speed Motor Controlling Eq.
( ) 21. Equalizers.

Products to Be Checked by Radio-Electronic Manufacturers

( ) 22. **Facsimile Equipment.**
( ) 23. **Filters:**
- ( ) a. Band Pass.
- ( ) b. Noise Elimination.
- ( ) c. Sound Effect.

( ) 24. **Frequency Meas. Equip.:**
- ( ) a. Audio Frequency.
- ( ) b. Primary Standards.
- ( ) c. Radio Frequency.
- ( ) d. Secondary Standards.

( ) 25. **Fuses & Fuse Holders:**
Generators:
- a. *Power, see Motor.*
- b. *Signal, see Frequency Meas. Equip., also Testing Equip.*

( ) 26. **Graphic Recorders.**
( ) 27. **Hardware:**
- ( ) a. Binding Posts.
- ( ) b. Bushings.
- ( ) c. Dials & Tuning Controls.
- ( ) d. Flexible Shafts.
- ( ) e. Lugs.
- ( ) f. Screws.
- ( ) g. Springs.

( ) 28. **Induction Heating Equip.**
( ) 29. **Inductors.**
( ) 30. **Insulation:**
(Also see Ceramics)
- ( ) a. Cloth.
- ( ) b. Glass Seals.
- ( ) c. Mica.
- ( ) d. Paper.
- ( ) e. Varnished Cambric.

( ) 31. **Jacks & Jack Fields.**
( ) 32. **Keys:**
- ( ) a. Switching.
- ( ) b. Telegraph.

Knobs, see Moulded Prods.

( ) 33. **Lacquers:**
- ( ) a. Finishing.
- ( ) b. Fungus Proofing.
- ( ) c. Protecting.
- ( ) d. Waterproofing.

( ) 34. **Loudspeakers & Headphones.**
( ) 35. **Machinery, Fixtures, & Tools for Radio-Electronic Mfg.**
( ) 36. **Magnets:**
- ( ) a. Electro.
- ( ) b. Permanent.

Measuring Equipment, see Testing Equipment.

( ) 37. **Metals:**
- ( ) a. Copper.
- ( ) b. Ferrous.
- ( ) c. Non-Ferrous.
- ( ) d. Precious & Rare.

( ) 38. **Meters:**
- ( ) a. Ammeters.
- ( ) b. Frequency Indicating.
- ( ) c. Power Level.
- ( ) d. Vacuum Tube Voltmeters.
- ( ) e. Voltmeters.
- ( ) f. Wattmeters.

Mica, see Insulation.

( ) 39. **Microphones.**
( ) 40. **Monitoring Equipment:**
- ( ) a. Frequency.
- ( ) b. Modulation.

( ) 41. **Motor Generators:**
- ( ) a. Dynamotors.
- ( ) b. Motor Generators.
- ( ) c. Rotary Converters.

( ) 42. **Motors, Very Small.**
( ) 43. **Moulded Products & Moulding Services:**
- ( ) a. Cabinets.
- ( ) b. Insulators.
- ( ) c. Knobs, etc.
- ( ) d. Plastic Parts.
- ( ) e. Phenolic Sheet.

( ) 44. **Optical Systems, Mirrors, Screens & Accessories.**
( ) 45. **Oscillators:**
- ( ) a. Audio Frequency.
- ( ) b. Radio Frequency.
- ( ) c. Square Wave Generators.

( ) 46. **Oscillographs & Accessories.**
( ) 47. **Panels.**
( ) 48. **Phonograph & Transcription Pickups:**
- ( ) a. Crystal Cartridges.
- ( ) b. Magnetic Cartridges.
- ( ) c. Playback Arms, only.
- d. *Preamplifiers, see Amplifiers.*

( ) 49. **Pilot Lights & Assemblies.**
( ) 50. **Plastics:**
- ( ) a. Raw Materials.
- ( ) b. Rods.
- ( ) c. Sheets.

( ) 51. **Plugs.**
( ) 52. **Power Supplies.**
( ) 53. **Pumps, Vacuum.**
Racks, see Chassis.
( ) 54. **Radar Equipment & Associated Apparatus. *(Also see Aircraft & Airport Eq.)***
( ) 55. **Radio Receivers:**
- ( ) a. Broadcast.
- ( ) b. Communications.
- ( ) c. Fixed Frequency.
- ( ) d. Freq. Modulation.
- ( ) e. Special Purpose.
- ( ) f. Television.

( ) 56. **Record Changers.**
( ) 57. **Recording Equip. & Supp.:**
- ( ) a. Blanks.
- ( ) b. Cutting Heads.
- ( ) c. Magnetic Wire Recorders.
- ( ) d. Needles.
- ( ) e. Turntables & Machs.

( ) 58. **Rectifiers:**
- ( ) a. Metallic.
- ( ) b. Meter Rectifiers.
- ( ) c. Vacuum Tube. *(Also see Power Supp.)*

Regulators, Voltage, see Voltage Regulators.

( ) 59. **Relays:**
- ( ) a. Keying.
- ( ) b. Power.
- ( ) c. Stepping.
- ( ) d. Telephone Types.
- ( ) e. Time Delay.
- ( ) f. Vacuum Enclosed.

( ) 60. **Remote Controlling Equip.:**
- ( ) a. Automatic Tuning Mechanisms.
- ( ) b. Remote Switching Mechanisms.

( ) 61. **Resistors:**
- ( ) a. Fixed.
- ( ) b. Precision.
- ( ) c. Vacuum Sealed.
- ( ) d. Variable.
- ( ) e. Wire Wound.

( ) 62. **Sockets:**
- ( ) a. Receiving Types.
- ( ) b. Transmitting Types.

( ) 63. **Solder:**
- ( ) a. Cored.
- ( ) b. Plain.

Speakers, see Loudspeakers.

( ) 64. **Switches:**
- ( ) a. Circuit Breaking.
- ( ) b. Key.
- ( ) c. Power.
- ( ) d. Receiver Wave Band Changing.
- ( ) e. Rotary.
- ( ) f. Time Delay.
- ( ) g. Transmitter Wave Band Changing.

( ) 65. **Testing & Measuring Equip.:**
- ( ) a. Bridges.
- ( ) b. Capacitor Testing.
- ( ) c. Inductance & "Q" Testing.
- ( ) d. Resistance Testing.
- ( ) e. Vacuum Tube Testing.
- ( ) f. Wave Form Analyzers & Distortion Testing.

( ) 66. **Transformers:**
- ( ) a. Audio Frequency.
- ( ) b. Hermetically Sealed Types.
- ( ) c. High Fidelity Audio Types.
- ( ) d. Power Components.
- ( ) e. Pulse Generating Types.
- ( ) f. Radio Frequency.

( ) 67. **Transmitters:**
- ( ) a. Amplitude Modulation.
- ( ) b. Communication.
- ( ) c. Freq. Modulation.
- ( ) d. Police & Emergency.
- ( ) e. Television.
- ( ) f. Ultra-High Freq.

( ) 68. **Ultra-High Frequency Equipment & Accessories:**
- ( ) a. Antennas & Reflectors.
- ( ) b. Measuring & Testing Equipment.
- ( ) c. Tuning Elements.
- ( ) d. Wave Guides.

( ) 69. **Vacuum Tubes:**
- ( ) a. Cathode Ray.
- ( ) b. Geiger Mueller.
- ( ) c. Industrial Types.
- ( ) d. Klystrons & Magnetrons.
- ( ) e. Receiving Types.
- ( ) f. Rectifiers.
- ( ) g. Special Purpose & Phototubes.
- ( ) h. Television Tubes.
- ( ) i. Transmitting Types.
- ( ) j. Voltage Regulator.

Varnishes, see Lacquers.

( ) 70. **Vibrators, Power Supply.**
( ) 71. **Voltage Regulators:**
- ( ) a. Automatic.
- ( ) b. Manually Controlled.

( ) 72. **Waxes & Sealing Compounds.**
( ) 73. **Wire:**
- ( ) a. Copper.
- ( ) b. Precious Metal.

Products Not Listed Above

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IMPORTANT—*Mail Today To:*
Industry Research Division,
Proceedings of the I.R.E.,
Room 707, 303 W. 42nd St.,
New York 18, New York.

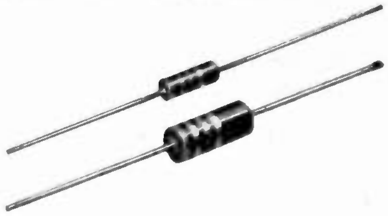
News—New Products

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

(Continued from page 62A)

Composition Resistors

The well-known "Little-Devil" composition resistors, offered by **Ohmite Manufacturing Co.**, 4862 Fluornoy St., Chicago 44, Ill., are now available in the ½- and 1-watt sizes with a tolerance of ±5 per cent. This is in addition to the standard line of 10 per cent units.



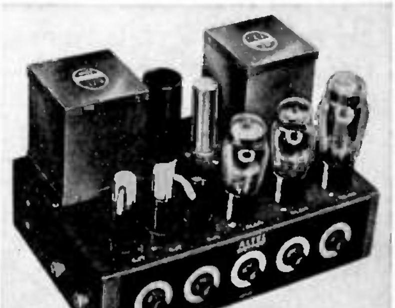
The size of the ½-watt resistor is only ⅜-inch long and 9/64-inch diameter; the 1-watt unit is 9/16-inch long and 7/32-inch diameter. They are claimed to meet all test requirements for the best quality characteristics of joint Army-Navy specification JAN-R-11, including salt-water immersion cycling and high-humidity tests.

These resistors are completely sealed and insulated by molded plastic. Leads are soft copper wire, hardened immediately adjacent to the resistor body, strongly anchored, and hot-solder coated.

All units are individually marked with the resistance value and wattage for quick identification, and in addition are color-coded. They are available from stock, according to an announcement by the manufacturer, in standard RMA values and 10 ohms to 22 megohms.

New A-323B Amplifier

Altec Lansing Corp., of Hollywood, Calif., announce the availability of a new amplifier, designated as A-323B. It is claimed to be capable of realizing the full resources of the new professional f.m. tuners.



The manufacturer emphasises the fact that this amplifier was designed with a particular view to its use in high-quality music reproducing systems in which the

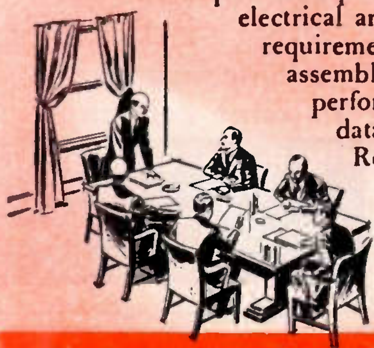
(Continued on page 68A)

The Best Resistors Are Not Enough

SERVICE IS VITAL

The most complete line of high quality resistors is not enough. IRC considers sincere service—cooperative development work, unbiased recommendations, on time deliveries, genuine help in emergencies and friendly follow thru also vital in meeting advancing demands of industry.

The RESISTOR ANALYSIS COUNCIL is a natural development of this concept. Sponsored by IRC, and established to provide experienced technical aid on your resistor problems—electrical and mechanical. Working together on your specific requirements, confidential analysis may disclose ways to cut assembly costs, eliminate expensive "specials" or improve performance. You may obtain this counsel by sending available data on your resistor problem to the RAC at—International Resistance Company, 401 N. Broad St., Philadelphia 8, Pa.



Resistor Analysis Council

A new IRC industry service. Composed of IRC electrical and mechanical engineers plus production specialists, the RAC—Resistar Analysis Council operates as consultant to engineers and designers. Provides confidential analysis of resistor requirements—helps solve electrical, mechanical and cost considerations. RAC's industry knowledge is sufficiently broad that recommendations need nat be canfined to IRC products. Consult the Resistor Analysis Council on your present or anticipated resistor problems.

Complete Line

Only IRC produces such a wide range of resistor types. All your requirements can be readily supplied from one source. Manufacturing all types, IRC's recommendation on the proper resistor for your product is unbiased. For over two decades IRC has concentrated its engineering and manufacturing talent exclusively on resistors. You benefit by this accumulated experience when you specify IRC. Technical Data Bulletins are available on each IRC resistor type.

On Time Deliveries

Purchasing Agents and material control executives rely upon IRC's "on time" deliveries. They know that regardless of a product's high quality, assembly line problems are a natural consequence when delivery schedules aren't met. IRC delivers "on time"—also maintains factory stock piles of most popular resistor types and ranges assuring you of real assistance in emergencies.

Industrial Service Plan

Providing speedy "round-the-corner" deliveries on your small order requirements, IRC's distributor network maintains well-stocked shelves of all standard items. No time lost when you need experimental or maintenance quantities in a hurry. When time means money you profit by competent service from the IRC distributor in your area—write for his name and address.

INTERNATIONAL RESISTANCE COMPANY

IRC

IN CANADA: INTERNATIONAL RESISTANCE COMPANY, LTD., TORONTO, LICENSEE

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

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 66A)

electrical elements—tuner, amplifier, record player and speaker—are either wholly concealed in the interior structure of a room, or partly concealed in furniture already in harmony with the interior scheme.

This amplifier features built-in equalization to operate direct from the new high-quality General Electric variable-reluctance or Pickering magnetic pickup cartridges. It has a hum-balancing potentiometer that eliminates the necessity of careful selection of present-day tubes for quiet, noiseless operation. Another feature is a treble tone control consisting of a true low-pass filter which is adjustable by steps to give a sharp cutoff of noise frequencies and yet allow full reproduction of all usable high frequencies on phonograph records.

The A323B amplifier has two high-impedance inputs, one for phonograph pickup and the other for radio. It carries a nominal rating of 15 watts and will deliver this rated power within 1db from 35 cycles to 12,000 cycles. Its frequency response is flat from 20 to 20,000 cycles.

Address inquiries to Altec Service Corp., 250 West 57th St., New York 19, N.Y.

High-Voltage Ignitron

A new ignitron tube, Type GL-5630, for radio transmitter and power-rectification applications, has been developed by the Tube Division of the General Electric Co., Electronics Dept., Schenectady N. Y. The new tube rectifies and regulates current and provides a one-cycle circuit-breaker action simultaneously.

Suitable for applications which require up to 3000 kilowatts of d.c. power, the new tube is expected to solve an important power-supply problem for broadcast stations, users of induction heating, and laboratories employing cyclotrons and synchrotrons.

A control grid, which times current to a microsecond, gives the tube its voltage-regulating and circuit-breaker qualities. Its handling of high voltages is achieved by a special potential-dividing grid which lowers the voltage gradient between the anode and cathode.

This new tube is of the stainless-steel-jacketed type, and has a peak voltage, forward or reverse, of 20,000 volts. Its peak current is 200 amperes and its average current is 50 amperes.

News—New Products

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

Electronic Microammeter

A high-sensitivity d.c. microammeter, known as Model 301, is now being produced by **Beta Electronics Co., 1762 Third Ave., New York 29, N. Y.**

The manufacturer claims that this instrument cannot be damaged by any degree of overload. Five sensitivity ranges, of from 0.01 microamperes full-scale to 100 microamperes full-scale, are provided, with 40 millivolts full-scale input on all ranges. This unit may also be used as a null-detecting galvanometer, with a sensitivity of approximately 100 millivolts full-scale, or 0.0005 microamperes per division.

The d. c. amplifier is stablized with negative feedback, so that the zero drift after a short warm-up period is unnoticeable, and the zero shift between ranges is negligible. The instrument operates satisfactorily over an input-voltage range of 95–130 volts, 50/60 cycles.

A sloping-panel steel cabinet, measuring 8×8×8 inches, houses the entire unit. There is a minimum number of controls, rendering operation extremely simple. An internal calibrating circuit is included, permitting rapid check of accuracy at any time.

Applications for the instrument may be found in the fields of photoelectricity, ionization-gauge current measurements, high-resistivity measurements, biophysical research, etc., particularly where small currents are to be measured.

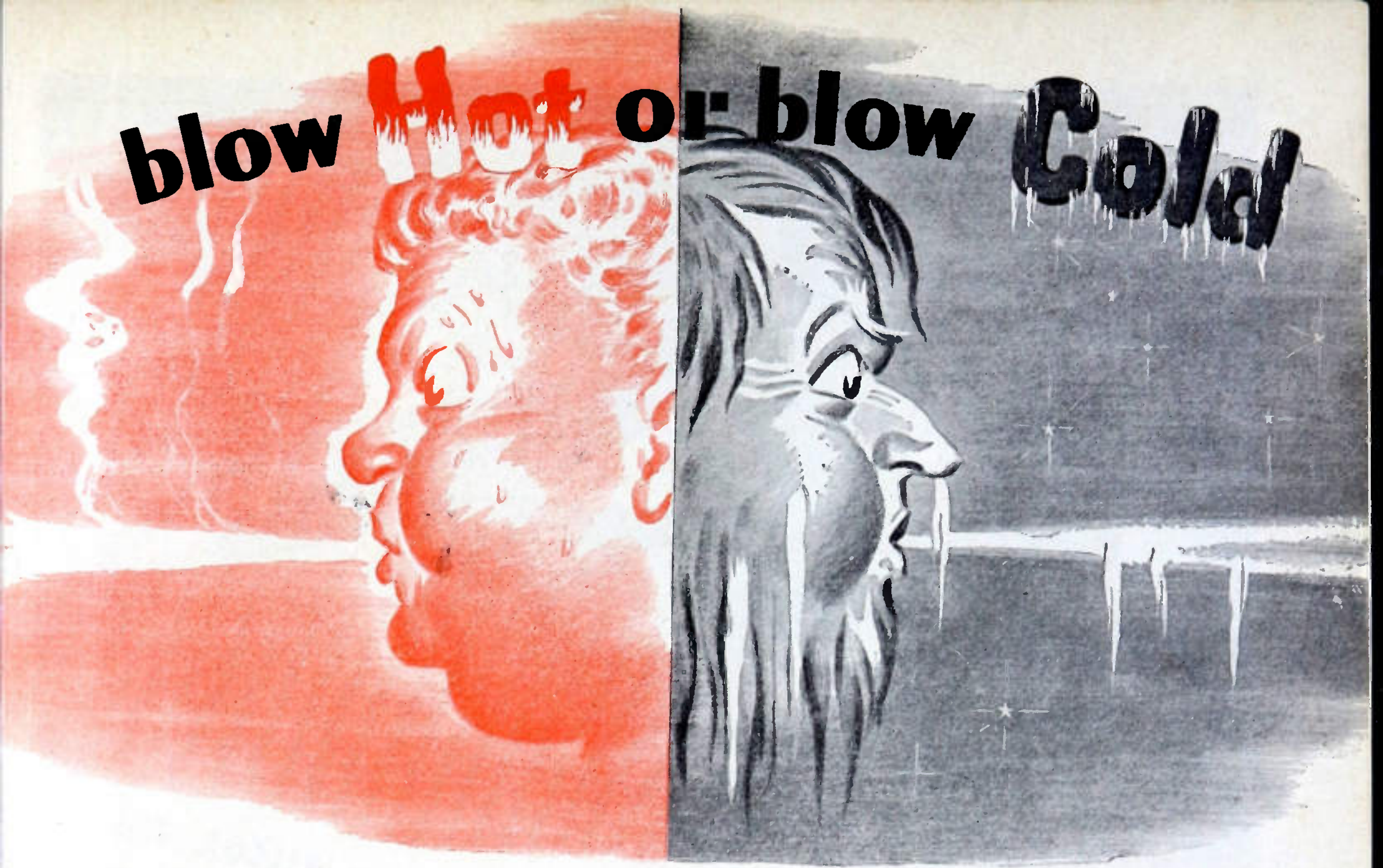
NOTICE

Information for our News and New Products section is warmly welcomed. News releases should be addressed to **Mrs. Harriet P. Watkins, Industry Research Division, Proceedings of I.R.E., Room 707, 303 West 42nd St., New York 18, N. Y.** Photographs, and electrotypes if not over 2" wide, are helpful. Stories should pertain to products of interest specifically to radio engineers.

PROFESSIONAL CARDS

INDEX AND DISPLAY ADVERTISERS

DISPLAY ADVERTISERS



silicone-treated DIEFLEX

VARNISHED TUBING PRODUCTS

...provides superior electrical insulation

Here's a happy combination of materials that will solve your problem. Dieflex tubings and sleevings of braided glass fiber are impregnated with heat-defying Silicone. The result is an insulating product that will stand unusually high temperatures for long periods of time. You can use it confidently where other products have failed.

Dieflex tubings and sleevings in all VTA and ASTM grades are carefully manufactured to help solve your insulation problems. Whether the base is braided cotton or inorganic glass fiber, you're sure of a continuously high standard of quality that insures the best performance in your product.

DIEFLEX PRODUCTS LIST

MADE WITH BRAIDED COTTON SLEEVING BASE—VTA Grade A-1 Magneto Grade Varnished Tubings—VTA Grade B-1 Standard Grade Varnished Tubings—VTA Grades C-1 and C-2 Heavily Coated Saturated Sleevings—VTA Grade C-3 Lightly Coated Saturated Sleevings—Heavy Wall Varnished Tubings and Saturated Sleevings.

MADE WITH BRAIDED GLASS SLEEVING BASE—VTA Grade A-1 Magneto Grade Varnished Fiberglas Tubings—VTA Grade C-1 Extra Heavily Saturated Fiberglas Sleevings—VTA Grade C-2 Heavily Saturated Fiberglas Sleevings—VTA Grade C-3 Lightly Saturated Fiberglas Sleevings—Silicone-Treated Fiberglas Varnished Tubings and Saturated Sleevings.

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THE GREAT SILVER FLEET

We have designed—and have available—many types of C-D Quietones which are equally effective on both Radio and video bands. They meet every requirement of manufacturers' cost and production schedules. One of these standard types may remove your product from the list of Radio interference generators. If not, we're ready and waiting—with a modern and complete laboratory and experienced engineers—to design and build a Quietone to meet your specific needs. Your inquiry is cordially invited. Cornell-Dubilier Electric Corporation, Dept. **M-12** South Plainfield, New Jersey. Other large plants in New Bedford, Worcester and Brookline, Massachusetts, and Providence, Rhode Island.

Make Your Product More Saleable with C-D Quietone Radio Interference Filters and Spark Suppressors.

CORNELL-DUBILIER

WORLD'S LARGEST MANUFACTURER OF

CAPACITORS

* Reg. U.S. Pat. Off.