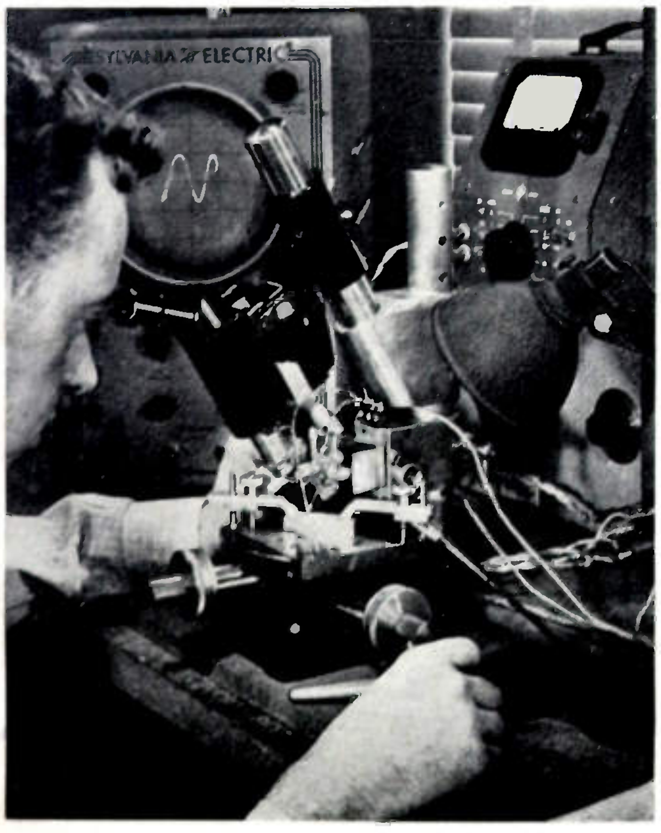
Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

(Including the WAVES AND ELECTRONS Section)

May, 1949

Volume 37 Number 5



Sylvania Electric Products Inc.

PRECISION ADJUSTMENT OF GERMANIUM-TRIODE CONTACT POINTS

Adjustment of the contact points on germanium triodes requires literally microscopic precision. In the laboratory the engineer uses a high-power microscope and micrometer adjusting heads to set the points within a small fraction of a mil.

PROCEEDINGS OF THE I.R.E.

Waves and Electrons Section

Table of Contents Follows Page 32A

The Institute of Radio Engineers


HERMONIC TVTA
AMPEREX 892
if your reserve tube is ready for active service in your THERMONIC equipment . . . it will pay you to put another tube on your shelf . . . the tube designed to our specifications, for your THERMONIC UNIT . . .
because we have been keeping an accurate check on tube performance life . . . and find that even though the manufacturer guarantees a minimum of 1,000 hours of operating life . . .
we can safely extend that guarantee to 1500 HOURS!
at a new LOWER PRICE!
. . . by purchasing your THERMONIC tube replacement through us, you now can be assured of
50% more
for your "efficient operation" dollar.
THER-MONIC
INDUCTION HEATING CORPORATION
181 WYTHE AVENUE • BROOKLYN 11, N. Y.
. . . More People Find That They Get MORE out of AMPEREX Electronic Tubes . . . Whatever the Application — it Will Pay You to . . .
Re-tube with AMPEREX
AMPEREX ELECTRONIC CORP.
25 WASHINGTON STREET, BROOKLYN 1, N. Y.
In Canada and Newfoundland: Rogers Majestic Limited
11-19 Brentcliffe Road, Leaside, Toronto, Ontario, Canada
AMPEREX

PROCEEDINGS OF THE I.R.E. May, 1949, Vol. 37, No. 5. Published monthly in two sections by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price $2.25 per copy. Subscriptions: United States and Canada, $18.00 a year; foreign countries $19.00 a year. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

Table of contents will be found following page 32A

NEW -hp- SIGNAL GENERATOR

FAST DIRECT READINGS

800 mc to 2100 mc

NO CHARTS OR INTERPOLATIONS

-hp- 614A UHF Signal Generator

Direct reading output, accuracy ±1 db...Constant internal impedance, SWR 3 db...Direct frequency control...External modulation 0.5 microseconds pulses to square waves...CW, FM, pulsed output.

This new *-hp-* signal generator will save you hours of time and work in making UHF measurements between 800 and 2100 mc. Its many different modulation and pulsing capabilities mean these man-hour economies can be applied to a wide variety of measurements—receiver sensitivity and alignment, signal-to-noise ratio, conversion gain, standing wave ratios, antenna gain and transmission line characteristics, to name but a few.

Carrier frequency in mc can be set and read directly on the large central tuning dial. R-f output from the klystron oscillator is also directly set and read in microvolts or db. No calibration charts or tedious interpolation are necessary. And thanks to the unique *-hp-* automatic tracking mechanism, no voltage adjustments are needed during operation.

R-f output ranges from 0.1 volt to 0.1 microvolt. Output may be continuous, pulsed, or frequency modulated at power supply frequency. The instrument may be modulated either externally or internally and may be synchronized with positive or negative pulses or sine waves.

Because of its wide range, high stability and versatile usefulness, this new *-hp-* signal generator is adaptable to almost any uhf measuring need. The instrument is available for early delivery. Contact your *-hp-* field representative or write direct to factory for complete details and technical specifications.

HEWLETT-PACKARD CO.
1874-D Page Mill Road, Palo Alto, California
Export Agents: Frazar & Hansen, Ltd.
301 Clay Street • San Francisco, Calif., U.S.A.

-hp- laboratory instruments
FOR SPEED AND ACCURACY

SPECIFICATIONS

FREQUENCY RANGE:
800 to 2100 mc. Selection is made by means of a single directly-calibrated control covering entire range. No charts are necessary.

FREQUENCY CALIBRATION ACCURACY:
±1%.

OUTPUT RANGE:
1 milliwatt or .223 volts to 0.1 microvolt (0 dbm to −127 dbm). Directly calibrated in microvolts and db; continuously monitored.

ATTENUATOR ACCURACY:
Within ±1 db without correction charts. A correction chart is provided when greater accuracy is desired.

OUTPUT IMPEDANCE:
50 ohms. SWR 3 db (VSWR 1.4).

EXTERNAL MODULATION:
By external pulses, positive or negative, peak amplitude 40 to 70v., 0.5 microseconds to square wave.

FM MODULATION:
Oscillator frequency sweeps at power line frequency. Phasing and sweep range controls provided. Maximum deviation approximately ±5 mc.

INTERNAL MODULATION:
Pulse repetition rate variable from 40 to 4000 per second; pulse length variable from 1 to 10 microseconds. Pulse rise and decay approximately 0.1 microseconds.

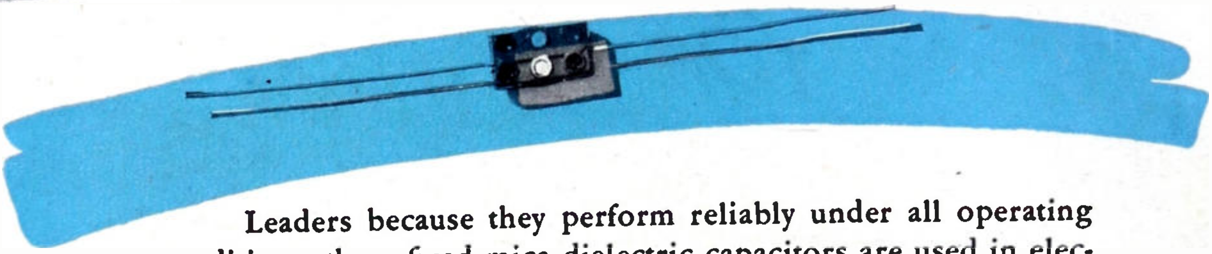
TRIGGER PULSES OUT:
1. Simultaneous with r-f pulse.
2. In advance of r-f pulse, variable 3 to 300 microseconds.
   (Both approximately 1 microsecond rise time, height 10 to 40 volts.)

EXTERNAL SYNC PULSE REQUIRED:
Amplitude from 10 to 50 volts of either positive or negative polarity and 1 to 20 microseconds width. May also be synchronized with sine waves.

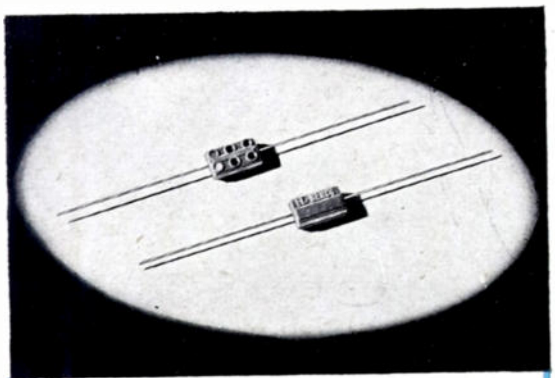
Data subject to change without notice.

EL-MENCO CAPACITORS

Leaders



Leaders because they perform reliably under all operating conditions, these fixed mica dielectric capacitors are used in electronic applications wherever long life and successful performance are demanded.



CM 15 MINIATURE CAPACITOR

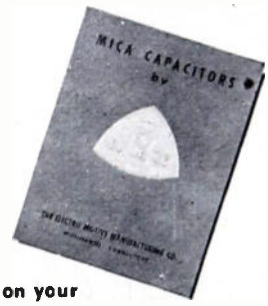
Actual Size $\frac{9}{32}$" x $\frac{1}{2}$" x $\frac{3}{16}$"
For Radio, Television and Other Electronic Applications
2 to 420 mmf. capacity at 500vDCw
2 to 524 mmf. capacity at 300v DCw
Temp. Co-efficient ±50 parts per million per degree C for most capacity values
6-dot standard color coded

Each tiny El-Menco Capacitor must pass life and humidity tests; meet standards set by the United States Army and Navy; pass tests at double their working voltages; prove their dielectric strength, temperature co-efficient and capacitance drift, and have their insulation resistance double-checked. These little leaders are molded in low-loss bakelite and wax-dipped for salt water immersion seal. They're available in a wide range, all impregnated, all precision-made, all JAN, RMA and RCM color-coded.

Why not protect your product's performance with capacitors made under these rigid conditions?

Specify **EL-MENCO**
TESTED ● RELIABLE ● LEADERS!

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



Write on your firm letterhead for Catalog and Samples

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

FOREIGN RADIO AND ELECTRONIC MANUFACTURERS COMMUNICATE DIRECT WITH OUR EXPORT DEPT. AT WILLIMANTIC, CONN.
ARCO ELECTRONICS, INC. 135 Liberty St., New York, N. Y.—Sole Agent for Jobbers and Distributors in U.S. and Canada

You can *depend* on Stackpole controls. Carefully supervised production means that you can depend on each unit to operate satisfactorily after it has been soldered into the circuit—and Stackpole facilities are such that you can depend, too, on *quantity deliveries* to meet *your* needs.

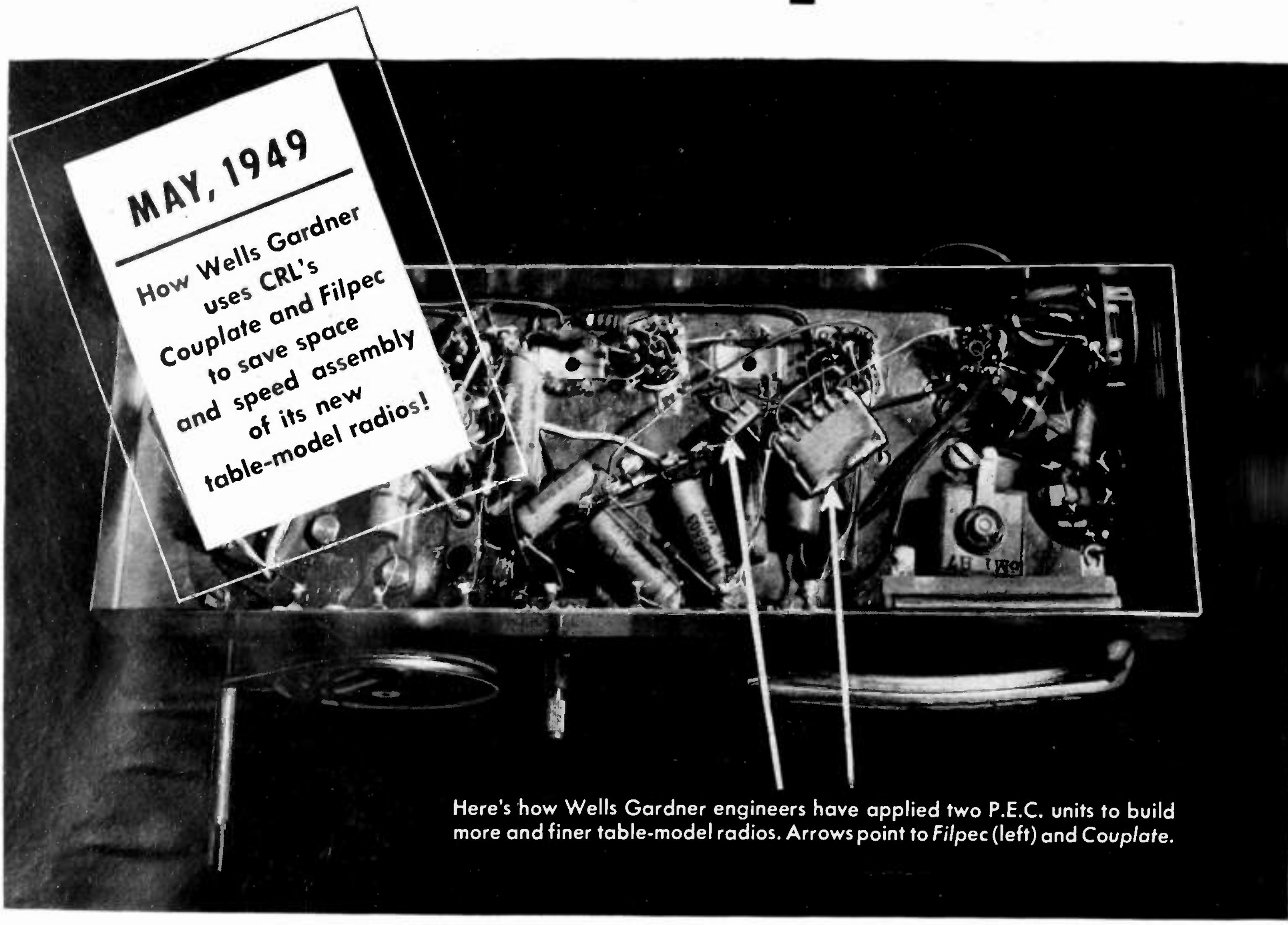
In both fixed and variable resistors, Stackpole is a major supplier to an important segment of the radio and electronic industries. If you are not already checking Stackpole regularly as your production releases and design requirements come up, we welcome the opportunity to cooperate on your next assignment. Write for Stackpole Control Engineering Bulletin RC-7D. The Complete Stackpole Electronic Components Catalog is also available on request.

STACKPOLE CARBON COMPANY · ST. MARYS, PA.

STACKPOLE

RESISTORS · CONTROLS · SWITCHES · IRON CORES

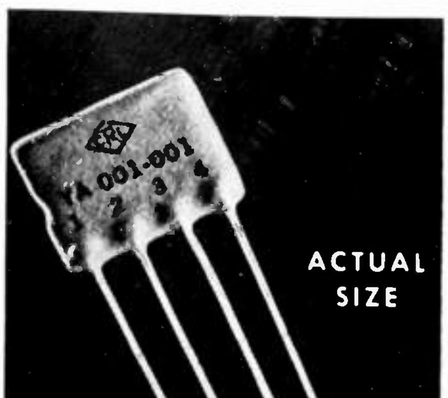
Centralab reports to

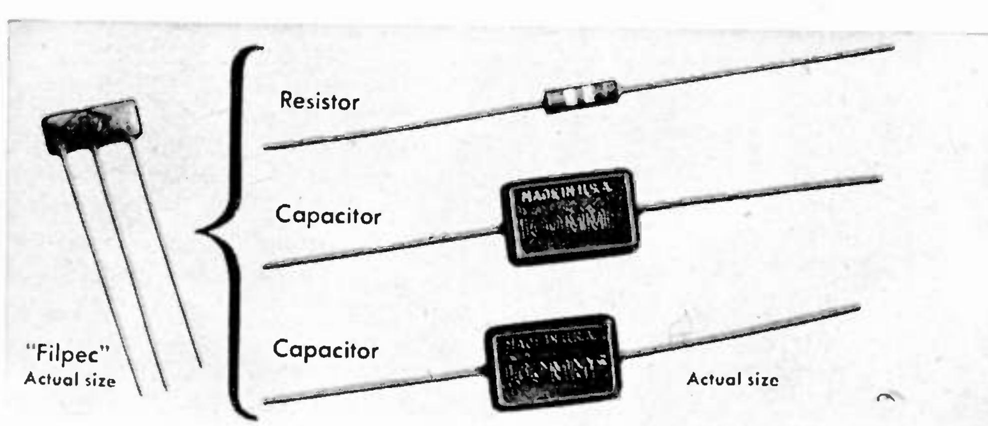


Here's how Wells Gardner engineers have applied two P.E.C. units to build more and finer table-model radios. Arrows point to *Filpec* (left) and *Couplate*.

Chassis courtesy of Wells Gardner & Co.

1 More and more manufacturers are turning to CRL's space-saving *Printed Electronic Circuits* to help them produce finer products, faster. That's how it is with Wells Gardner & Co., Chicago. Two Centralab P. E. C. units — *Couplate* and *Filpec* — are helping this firm cut radio assembling time by reducing the number of components needed and by eliminating many soldering operations. What's more, these same units improve performance by resisting temperature and humidity.

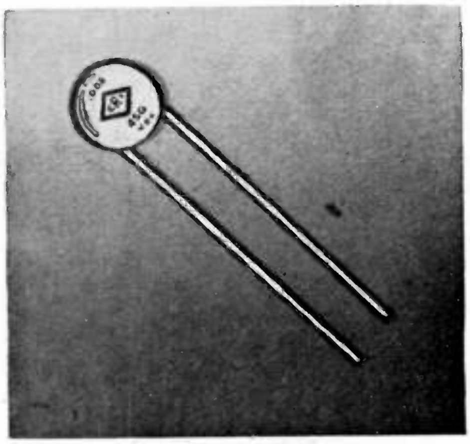


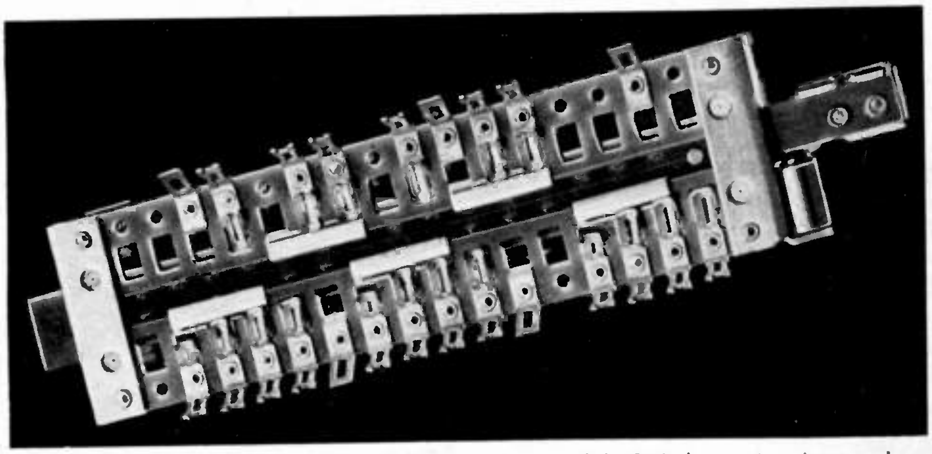


2 CRL's *Couplate* consists of a plate lead resistor, grid resistor, plate by pass capacitor and coupling capacitor. Write for Bulletin 42-6.

3 Centralab's *Filpec* is designed for use as a balanced diode load filter, combines up to three major components into one tiny unit, lighter and smaller than one ordinary capacitor. Capacitor values from 50 to 200 mmf. Resistor values from 5 ohms to 5 megohms. For complete information, write for Bulletin 42-9.

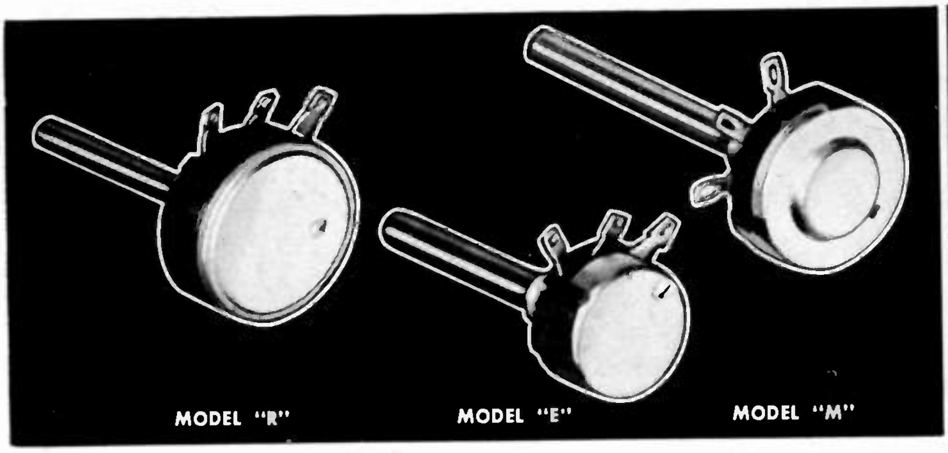
Electronic Industry

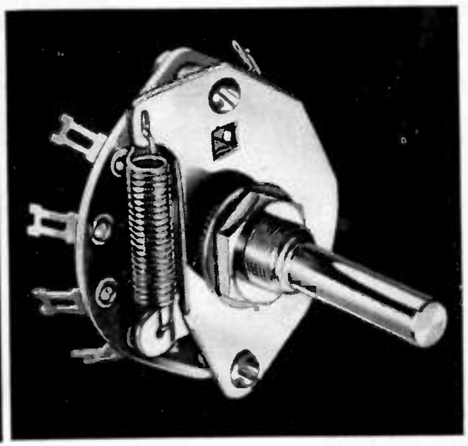




4 For by-pass or coupling applications, check CRL's original line of ceramic disc and tubular *Hi-Kaps*. For full facts, order Bulletins 42-3 and 42-4.

5 Centralab's development of a revolutionary, new *Slide Switch* promises improved AM and FM performance! Flat, horizontal design saves valuable space, allows short leads, convenient location to coils, reduced lead inductances for increased efficiency in low and high frequencies. Rugged, efficient. Write for Bulletin 953.





6 Let Centralab's complete *Radiohm* line take care of your special needs. Wide range of variations: *Model "R"* — wire wound, 3 watts; or composition type, 1 watt. *Model "E"* — composition type, ¼ watt. Direct contact, 6 resistance tapers. *Model "M"* — composition type, ½ watt. Write for Bulletin 697.

7 Great step forward in switching is CRL's New *Rotary Coil and Cam Index Switch*. Its coil spring gives you smoother action, longer life.

LOOK TO CENTRALAB IN 1949! ***First in component research that means lower costs for the electronic industry. If you're planning new equipment, let Centralab's sales and engineering service work with you. Get in touch with Centralab!***

Centralab

CRL ®

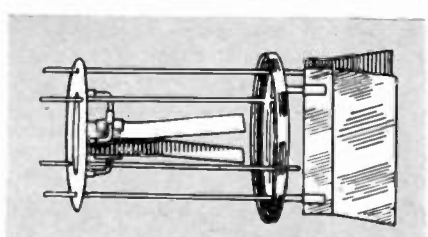
DIVISION OF GLOBE-UNION INC., MILWAUKEE, WIS.

Another Du Mont "First"—the new Du Mont Type 5XP-! A multiple-intensifier design, it features deflection sensitivity never before achieved by a cathode-ray tube in either the low- or high-voltage category. Specifically:

At E_{b2} of 2000 volts and E_{b3} of 4000 volts, only 24 to 36 d-c volts/in. of deflection are required! This is approximately *three* times the sensitivity of a low-voltage tube such as Type 5LP-A. This superlative performance of the vertical plate system is due to the design of the plates and to a slight increase in overall tube length—only ⅞" longer than Type 5LP-A.

Also featured are the high ratios of E_{b3} to E_{b2} voltages—up to 10:1, and high overall accelerating potential—up to 25,500 d-c volts.

Because the usable vertical deflection is a function of the ratio E_{b3}/E_{b2}, the full-screen deflection available at ratio 1:1 is reduced to 2.5" at 2:1, 1.75" at 5:1, and 1.25" at 10:1 ratios, respectively.



Capacitance from D_3 to D_4 held to 1.7 μμf by virtue of this new deflection-plate design, despite longer length and closer spacing required for high sensitivity.

Another feature is the shielding between deflection plates D_1-D_2 and D_3-D_4 to prevent interaction between plate pairs. And for general shielding of the tube, Du Mont mu-metal shield Type 2502 is available.

A choice of phosphors is available, such as the P1, P2, P4, P5, P7 and P11 screens. The flat face makes for ease of visual measurement and photography.

As with all Du Mont tubes, Type 5XP- is available as a separate unit or in combination with a Du Mont oscillograph. Several Du Mont oscillographs already in use, notably Types 280, 256-D, 250-H and 248-A, are readily adaptable to this latest tube.

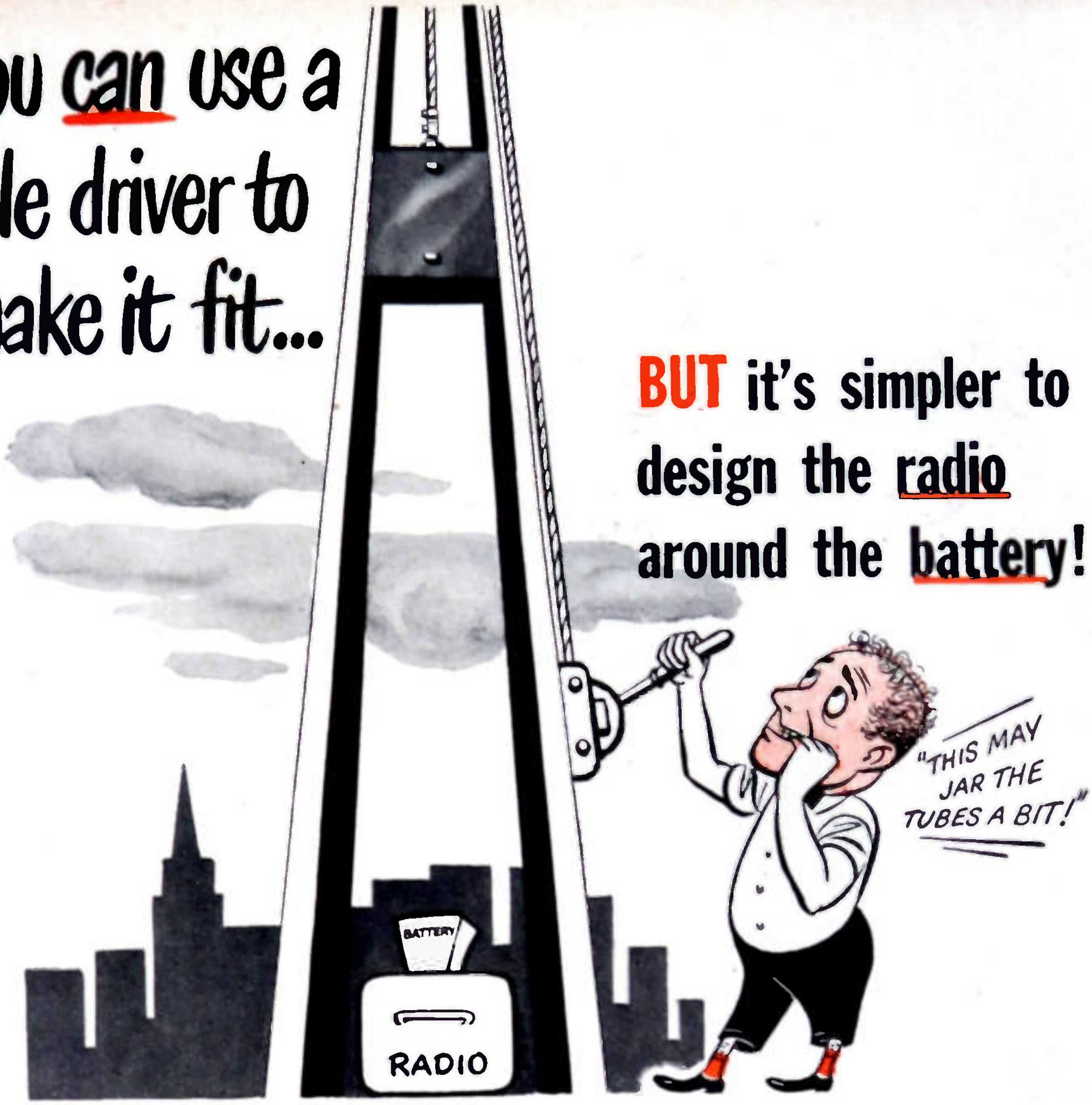
Write for detailed literature on the Type 5XP- tube and how it can be used in your Du Mont oscillograph.

DUMONT *for Oscillography*

ALLEN B. DU MONT LABORATORIES, INC., PASSAIC, N. J.
CABLE ADDRESS: ALBEEDU, NEW YORK, N. Y., U. S. A.

You can use a pile driver to make it fit...

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



● You'll find an "Eveready" radio battery to fit the size and power requirements of any type portable. "Eveready" brand batteries give longer playing life. They are the accepted standard for portable radios. Users can get replacements everywhere.

Call on our Battery Engineering Department for complete data on "Eveready" batteries.

● The No. 753 "Eveready" "A-B" battery pack provides plenty of power for compact "pickup" portables. Send for Battery Engineering Bulletin No. 2 which gives complete details.

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

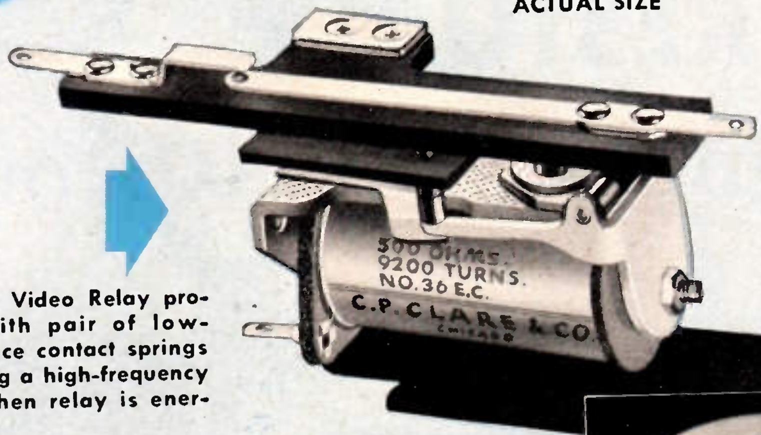
NATIONAL CARBON COMPANY, INC.
Unit of Union Carbide and Carbon Corporation
UCC
30 East 42nd Street, New York 17, N. Y.

EVEREADY
TRADE-MARK
RADIO BATTERIES

Division Sales Offices: Atlanta, Chicago, Dallas, Kansas City, New York, Pittsburgh, San Francisco

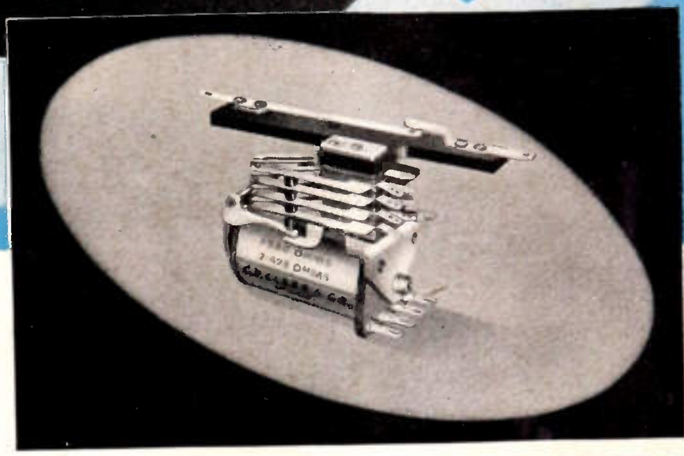
How Big does a Video Relay have to be?

ACTUAL SIZE



Type "J" Video Relay provided with pair of low-capacitance contact springs for closing a high-frequency circuit when relay is energized.

This Type "J" Video Relay shows how a number of auxiliary contact springs may be added for switching circuits which do not carry high-frequency currents.



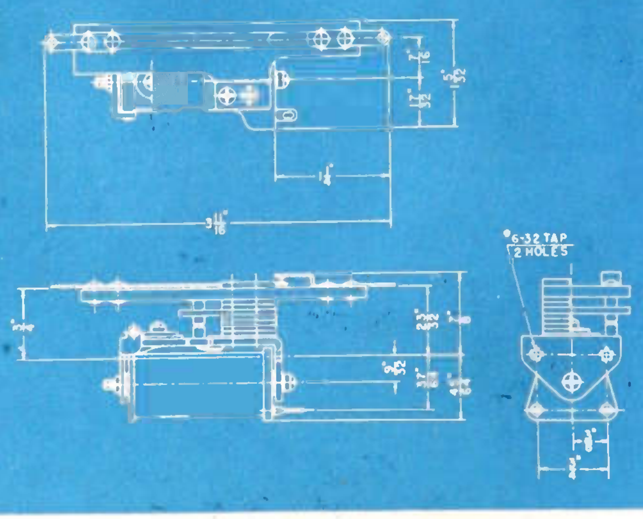
RCA Engineers posed this question to CLARE when available relays for their purpose proved too large, too cumbersome.

The close cooperation between engineers of the two companies which resulted has produced the CLARE Type "J" Video Relay, which meets every requirement for switching these high-frequency currents . . . and occupies but 7 cubic inches.

Success of this cooperation between RCA and CLARE engineers in developing a superior small-size, low-capacitance relay is not only important to this, the world's largest manufacturer of radio and television equipment, but it is of vital interest to every television engineer whose designs are often frustrated by the 17 cubic inches that other typical video relays require.

Clare sales engineers are located in principal cities. They will be glad to give you full information on this new video relay. Their counsel and advice may help you solve other relay problems. More and more, industrial designers bring their problems to Clare, whose long experience in meeting and solving them can save you many hours of tedious and costly experiment. Call your nearest CLARE sales engineer, or write to: C. P. Clare & Co., 4719 West Sunnyside Ave., Chicago 30, Illinois.

Dimensions of CLARE Video Relay



Capacitance of CLARE Video Relay

Tests show that this new CLARE relay with a contact gap of 0.025" has the following capacitances:

Interspring Capacitance, Contact Open

0.5 mmf. at 3 megacycles
0.5 mmf. at 10 megacycles
0.55 mmf. at 20 megacycles

Spring-to-Frame Capacitance, Contact Closed

1.4 mmf. at 3 megacycles
1.45 mmf. at 10 megacycles
1.8 mmf. at 20 megacycles

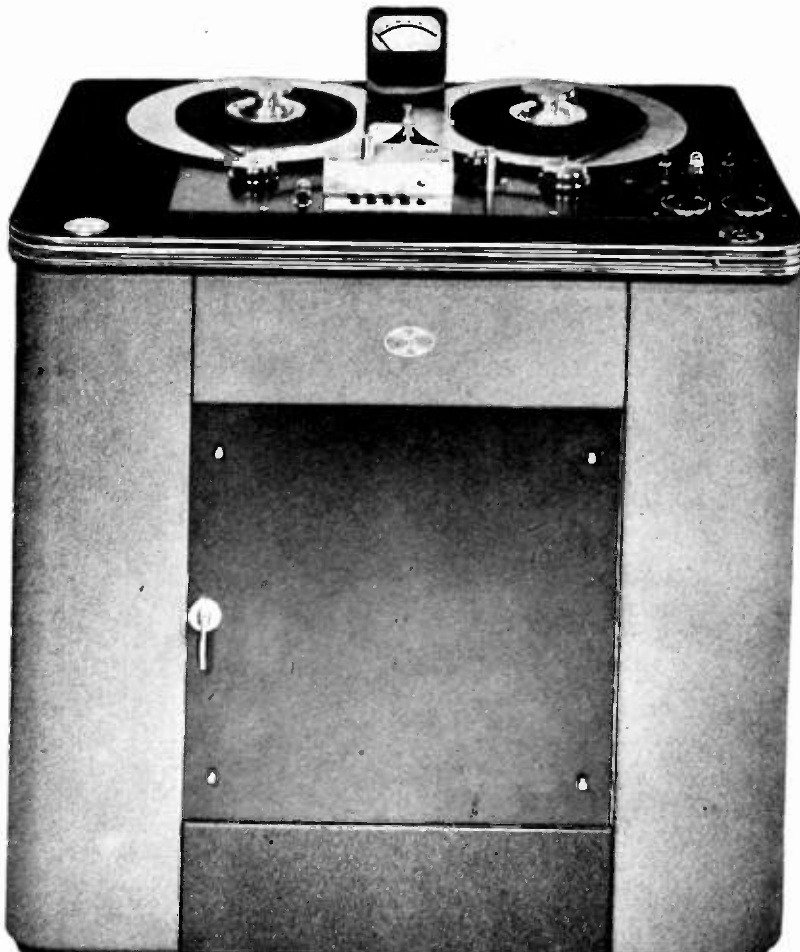
Write for Clare BULLETIN 106

CLARE RELAYS

First in the Industrial Field

Now! A Top Quality Tape Recorder at a Reasonable Price

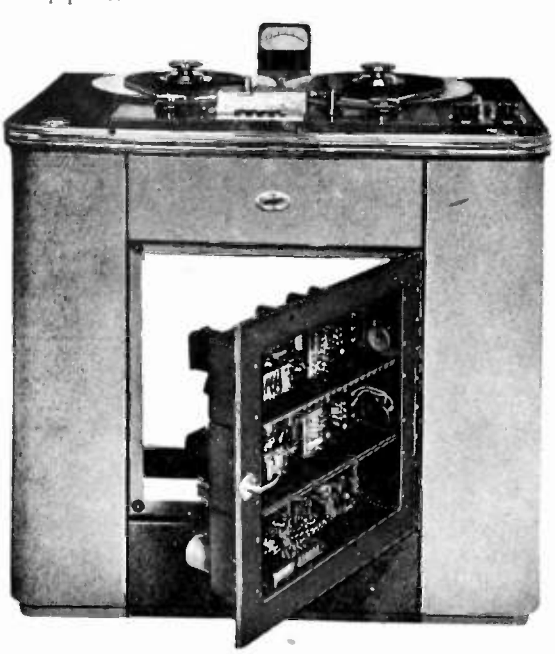
New PRESTO Magnetic Tape Recorder



AT LAST, a magnetic tape recorder that fully measures up to the most exacting requirements of broadcast network operations, independent stations and transcription producers, yet priced to have wide appeal.

Compare these specifications:

- Frequency response: 30 to 15,000 cps ± 1 db.
- Signal to noise ratio: Over 60 db below max. signal.
- Fast speed, 240 ft. per second forward and rewind, instantly reversible.
- Recording speeds 7½" or 15" per second (15" or 30" per second provided on request). Speed selection by special 2-speed motor.
- Reels direct mounted on motor shafts. Uses any type and size of reel up to 14".
- Erasing, recording and playback heads all mounted in separate housing – entire unit connected by plug-in for immediate replacement.
- Full-size illuminated scale V. U. meter on top panel.



Now! Greater Accessibility

Illustration shows how everything mechanical and electrical can be serviced from the front and top. Amplifiers and power supply are in swinging door behind removable panels. Mechanical units are mounted on top panel, hinged at rear so it can be opened upwards.

PRESTO

RECORDING CORPORATION

Paramus, New Jersey

Mailing Address:

P. O. Box 500, Hackensack, N. J.

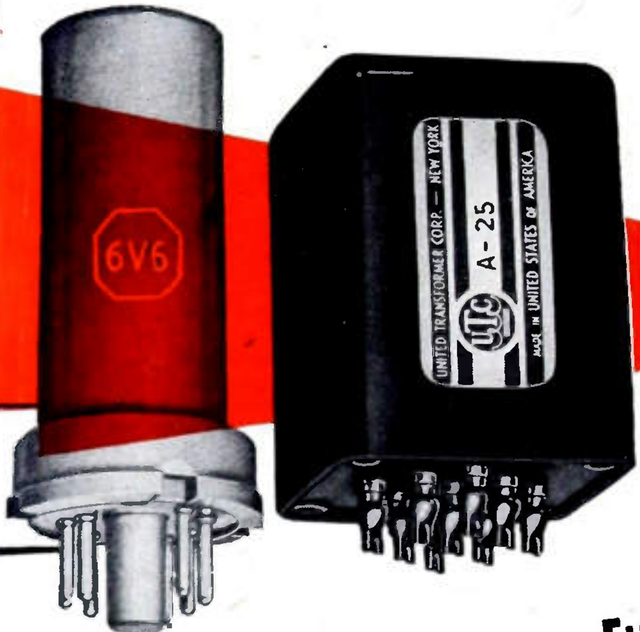
In Canada:

WALTER P. DOWNS, Ltd., Dominion Square Building, Montreal

WORLD'S LARGEST MANUFACTURER OF INSTANTANEOUS SOUND RECORDING EQUIPMENT AND DISCS

For Compact High Fidelity Equipment

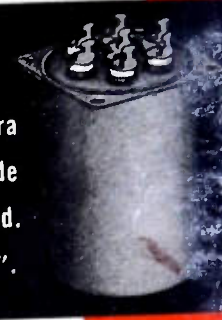
Ultra compact, lightweight, these UTC audio units are ideal for remote control amplifier and similar small equipment. New design methods provide high fidelity in all individual units, the frequency response being ±2DB from 30 to 20,000 cycles. There is no need to resonate one unit in an amplifier to compensate for the drop of another unit. All units, except those carrying DC in Primary, employ a true hum balancing coil structure which, combined with a high conductivity outer case, effects good inductive shielding. Maximum operating level + 10 DB. Weight—8 ounces. Dimensions—1½" wide x 1½" deep x 2" high.



Unit shown is actual size. 6V6 tube shown for comparison only.

HERMETICALLY SEALED

On special order, we can supply any of the Ultra Compacts hermetically sealed per Jan T-27 Grade 1 Class A in our RC 50 case as illustrated. Dimensions: Height 2¼", Base $1\frac{9}{16}$" x $1\frac{9}{16}$".



Ultra Compact High Fidelity Audio Units

Type No.	Application	Primary Impedance	Secondary Impedance	±2 DB from	List Price
A-10	Low impedance mike, pickup, or multiple line to grid	50, 125/150, 200/250, 333, 500/600 ohms	50,000 ohms	30-20,000	$15.00
A-11	Low impedance mike, pickup, or line to 1 or 2 grids. Multiple alloy shielded for extremely low hum pickup	50, 200, 500 ohms	50,000 ohms	50-10,000	16.00
A-12	Low impedance mike, pickup, or multiple line to push pull grids	50, 125/150, 200/250, 333, 500/600 ohms	80,000 ohms overall in two sections	30-20,000	15.00
A-18	Single plate to two grids, split primary	8,000 to 15,000 ohms	80,000 ohms overall, 2.3:1 turn ratio overall	30-20,000	14.00
A-19	Single plate to two grids 8 MA unbalanced D.C.	15,000 ohms	80,000 ohms overall, 2.3:1 turn ratio overall	50-20,000	18.00
A-24	Single plate to multiple line	8,000 to 15,000 ohms	50, 125/150, 200/250, 333, 500/600 ohms	30-20,000	15.00
A-25	Single plate to multiple line 8 MA unbalanced D.C.	8,000 to 15,000 ohms	50, 125/150, 200/250, 333, 500/600 ohms	50-12,000	14.00
A-26	Push pull low level plates to multiple line	8,000 to 15,000 ohms each side	50, 125/150, 200/250, 333, 500/600 ohms	50-20,000	15.00
A-30	Audio choke, 300 henrys @ 2 MA 6000 ohms D.C., 75 henrys @ 4 MA 1500 ohms D.C., inductance with no D.C. 450 henrys				10.00

The above listing includes only a few of the many Ultra Compact Audio Units available . . . Write for catalog PS409

United Transformer Co.

150 VARICK STREET NEW YORK 13, N. Y.

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

BENDIX-SCINTILLA

ELECTRICAL CONNECTORS

the finest money can buy!

these are the features

THAT HAVE MADE IT *the* ELECTRICAL CONNECTOR

- Moisture-proof, Pressure-tight
- Radio Quiet
- Single-piece Inserts
- Vibration-proof
- Light Weight
- High Arc Resistance
- Easy Assembly and Disassembly
- Fewer Parts than any other Connector
- Contacts filled with high-grade solder to insure top performance.

Plus this Important Advantage—

PRACTICALLY NO VOLTAGE DROP!

Contacts that carry maximum currents with a minimum voltage drop are only part of the many new advantages you get with Bendix-Scintilla* Electrical Connectors. The use of "Scinflex" dielectric material, an exclusive new Bendix-Scintilla development of outstanding stability, increases resistance to flash-over and creepage. In temperature extremes, from −67° F. to +300° F., performance is remarkable. Dielectric strength is never less than 300 volts per mil. Bendix-Scintilla Connectors have fewer parts than any other connector on the market—and that means lower maintenance costs and better performance.

*TRADEMARK

Write to our Sales Department for more detailed information.

BENDIX SCINTILLA

SCINTILLA MAGNETO DIVISION OF Bendix AVIATION CORPORATION
SIDNEY, NEW YORK

Export Sales: Bendix International Division, 72 Fifth Avenue, N.Y. 11, N.Y.

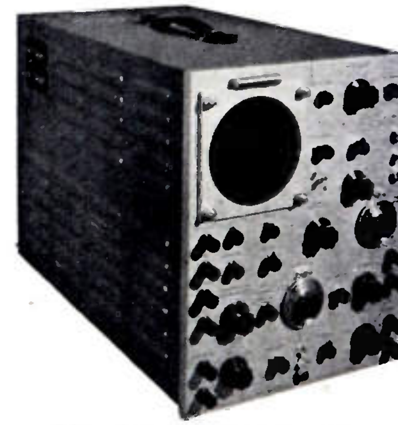
TEKTRONIX PLEDGES...

To serve our customers with products and policies unexcelled in the electronics industry and limited only by the current state of the art.

Tektronix Type 511-AD Oscilloscope
$845 f.o.b. Portland

Wide Band, Fast Sweeps

The **Type 511-AD**, with its 10 mc. amplifier, 0.25 microsecond video delay line and sweeps as fast as .1 microsec./cm. is excellent for the observation of pulses and high speed transient phenomena. Sweeps as slow as .01 sec./cm. enable the 511-AD to perform superlatively as a conventional oscilloscope.



Tektronix Type 512 Oscilloscope
$950 f.o.b. Portland

Direct Coupled, Slow Sweeps

The **Type 512** with a sensitivity of 5 mv./cm. DC and sweeps as slow as .3 sec./cm. solves many problems confronting workers in the fields where comparatively slow phenomena must be observed. Vertical amplifier bandwidth of 2 mc. and sweeps as fast as 3 microsec./cm. make it an excellent general purpose oscilloscope as well.

Both Instruments Feature:

- Direct reading sweep speed dials.
- Single, triggered or recurrent sweeps.
- Amplitude calibration facilities.
- All DC voltages electronically regulated.
- Any 20% of normal sweep may be expanded 5 times.

The Tektronix Field Engineering Representative in your area will be pleased to demonstrate our instruments upon request.

Phone EA 6197 — TEKTRONIX INC. — Cables Tektronix

712 S. E. Hawthorne Blvd.
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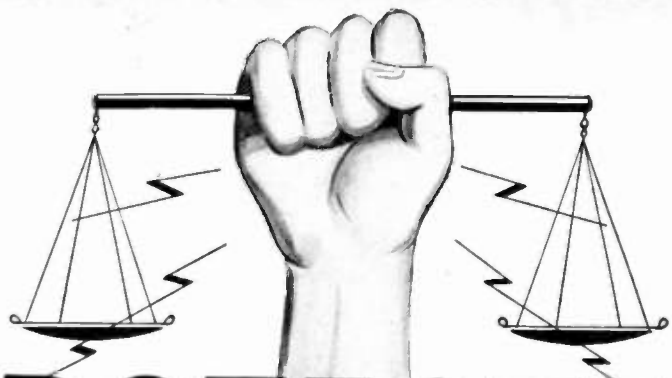
operate in safety



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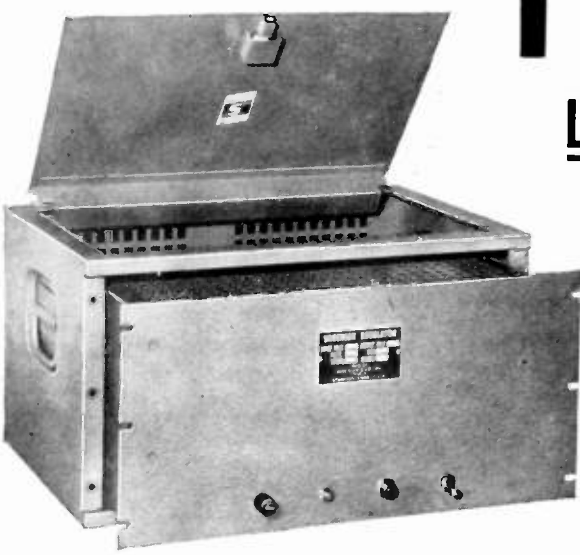

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298A & B

343A

5530

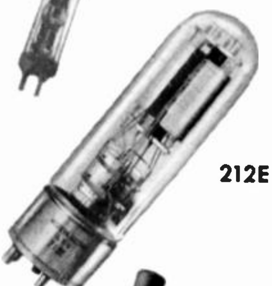
357B

232B

342A

233A

Western Electric Power Tubes for AM and FM



212E

251A

WHETHER your station operates on low power or high power, AM or FM, you'll find the tubes you want in Western Electric's line.

Always known for long service life and top quality performance, these broadcast power tubes and rectifiers—all engineered by Bell Telephone Laboratories—are now being made for Western Electric by Machlett Laboratories, Inc., another pioneer in the development of electron tubes.

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220C

343AA

5541

Western Electric

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212E	Air cooled triode, 275 watts
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222A	Water cooled high vacuum rectifier, 25 kv. inverse voltage
228A	Water cooled triode, 5 kilowatts
232B	Water cooled triode, 25 kilowatts
232BA	Forced-air cooled triode, 8 kilowatts
233A	Water cooled high vacuum rectifier, 50 kv. inverse voltage
236A	Water cooled triode, 20 kilowatts
240B	Water cooled triode, 10 kilowatts
241B	Air-cooled triode, 275 watts
251A	Air-cooled triode, 1000 watts
255B	Mercury vapor rectifier, 20 kv. inverse voltage
270A	Air cooled triode, 350 watts
279A	Air cooled triode, 1200 watts
298A and B	Water cooled triode, 100 kilowatts
308B	Air cooled triode, 250 watts
340A	Water cooled triode, 25 kilowatts
341AA	Forced-air cooled triode, 5 kilowatts
342A	Water cooled triode, 25 kilowatts
343A	Water cooled triode, 10 kilowatts
343AA	Forced-air cooled triode, 5 kilowatts
357B	Air cooled triode vhf, 400 watts
363A	Air cooled pentode, vhf, 350 watts
379A	Air cooled triode, 1200 watts
5530	Forced-air cooled triode, vhf, 3 kilowatts
5541	Forced-air cooled triode, vhf, 10 kilowatts

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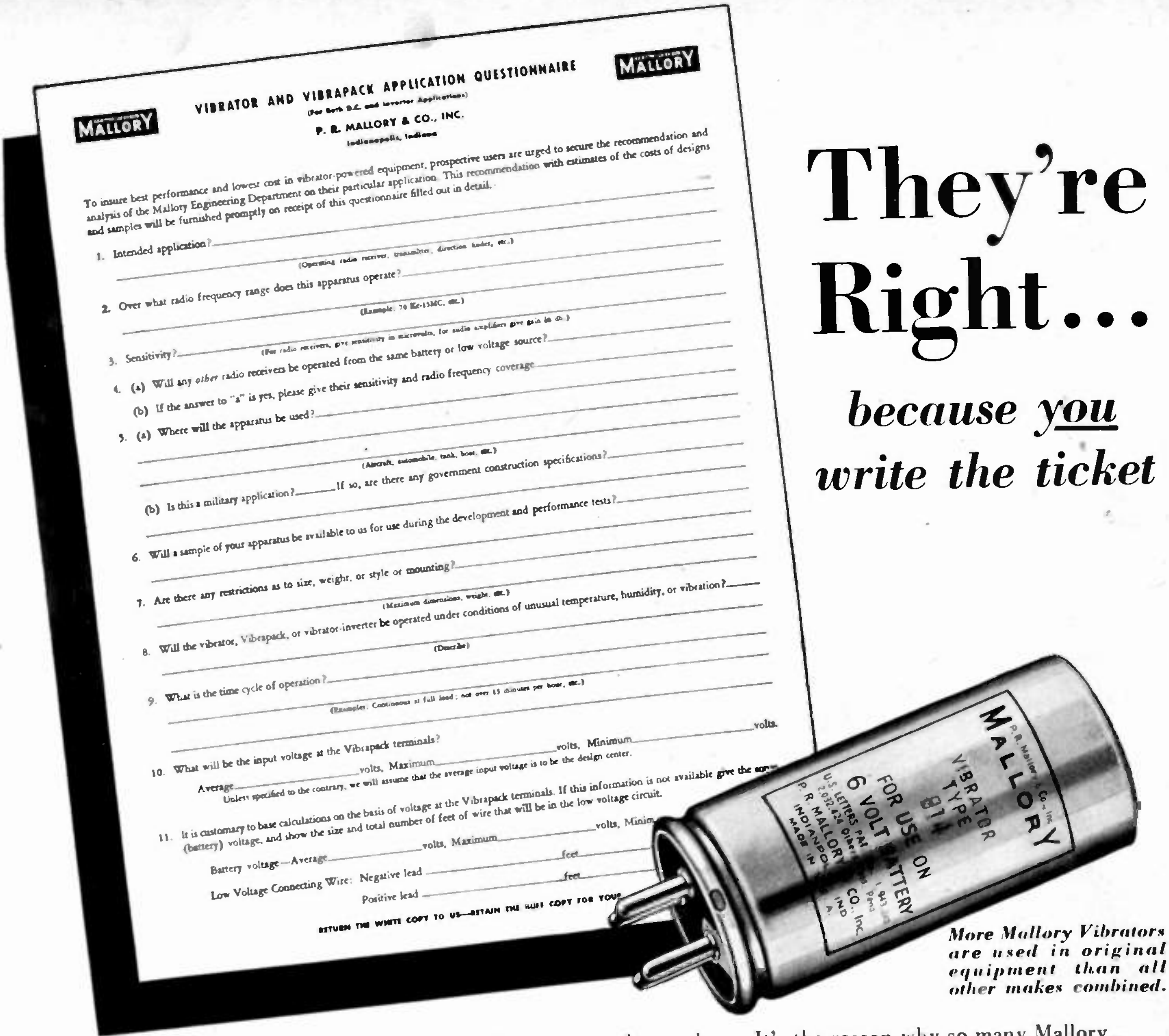

MALLORY VIBRATOR AND VIBRAPACK APPLICATION QUESTIONNAIRE MALLORY
(For Both D.C. and Inverter Applications)
P. R. MALLORY & CO., INC.
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To insure best performance and lowest cost in vibrator-powered equipment, prospective users are urged to secure the recommendation and analysis of the Mallory Engineering Department on their particular application. This recommendation with estimates of the costs of designs and samples will be furnished promptly on receipt of this questionnaire filled out in detail.

1. Intended application? ______ (Operating radio receiver, transmitter, direction finder, etc.)
2. Over what radio frequency range does this apparatus operate? ______ (Example: 70 Kc-15MC, etc.)
3. Sensitivity? ______ (For radio receivers, give sensitivity in microvolts, for audio amplifiers give gain in db.)
4. (a) Will any *other* radio receivers be operated from the same battery or low voltage source? ______
(b) If the answer to "a" is yes, please give their sensitivity and radio frequency coverage ______
5. (a) Where will the apparatus be used? ______ (Aircraft, automobile, tank, boat, etc.)
(b) Is this a military application? ______ If so, are there any government construction specifications? ______
6. Will a sample of your apparatus be available to us for use during the development and performance tests? ______
7. Are there any restrictions as to size, weight, or style or mounting? ______ (Maximum dimensions, weight, etc.)
8. Will the vibrator, Vibrapack, or vibrator-inverter be operated under conditions of unusual temperature, humidity, or vibration? ______ (Describe)
9. What is the time cycle of operation? ______ (Examples: Continuous at full load; not over 15 minutes per hour, etc.)
10. What will be the input voltage at the Vibrapack terminals? Average ______ volts, Maximum ______ volts, Minimum ______ volts.
Unless specified to the contrary, we will assume that the average input voltage is to be the design center.
11. It is customary to base calculations on the basis of voltage at the Vibrapack terminals. If this information is not available give the source (battery) voltage, and show the size and total number of feet of wire that will be in the low voltage circuit.
Battery voltage—Average ______ volts, Maximum ______ volts, Minim...
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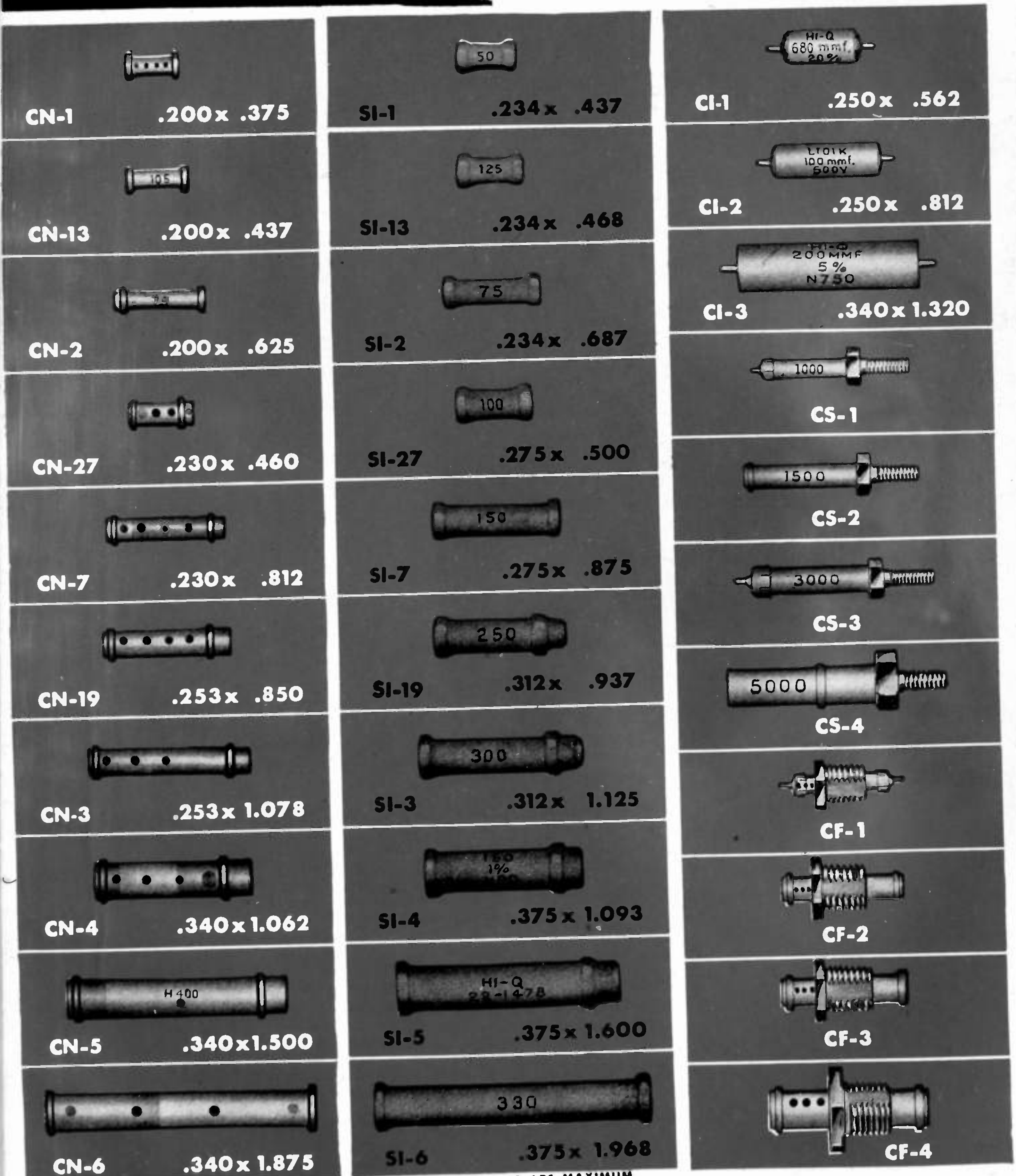

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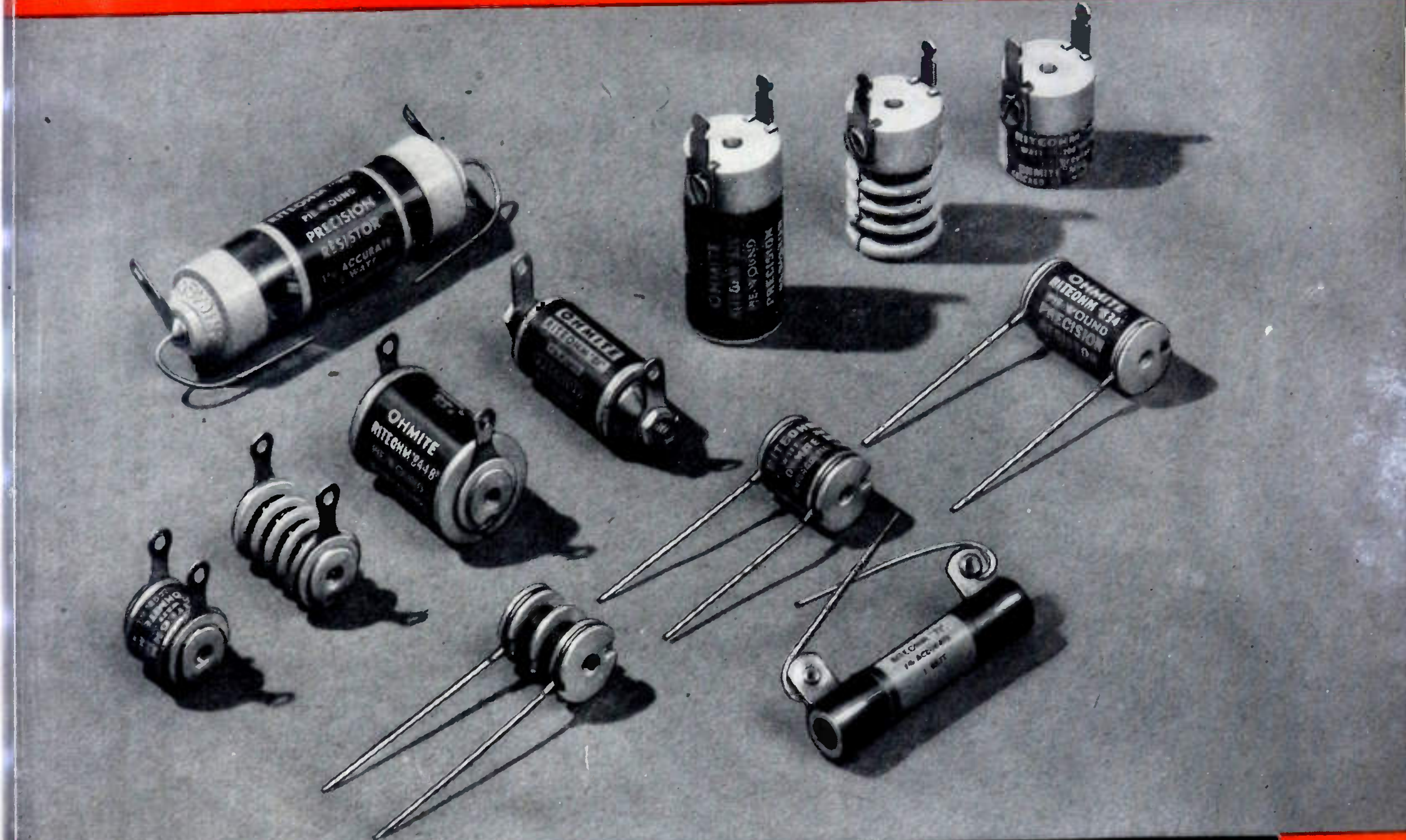

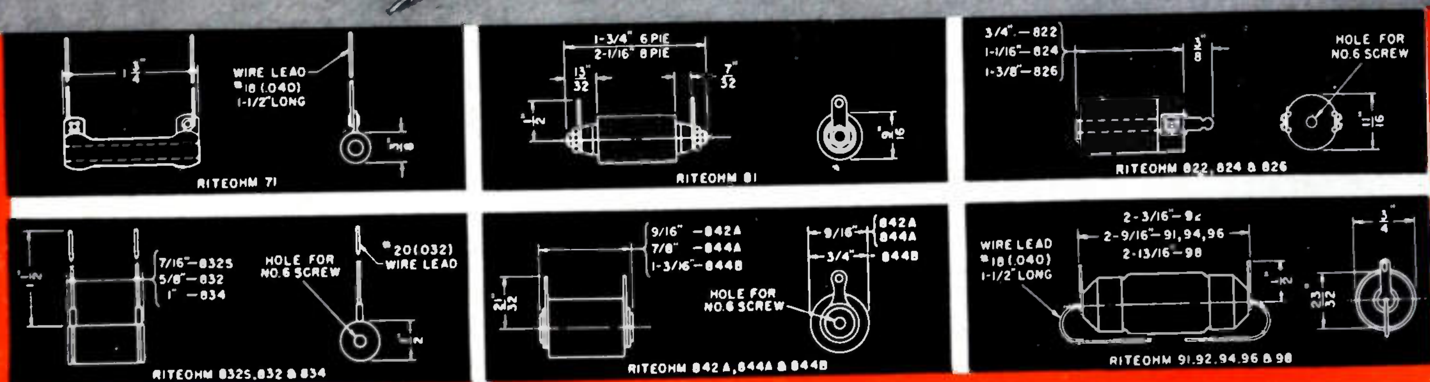

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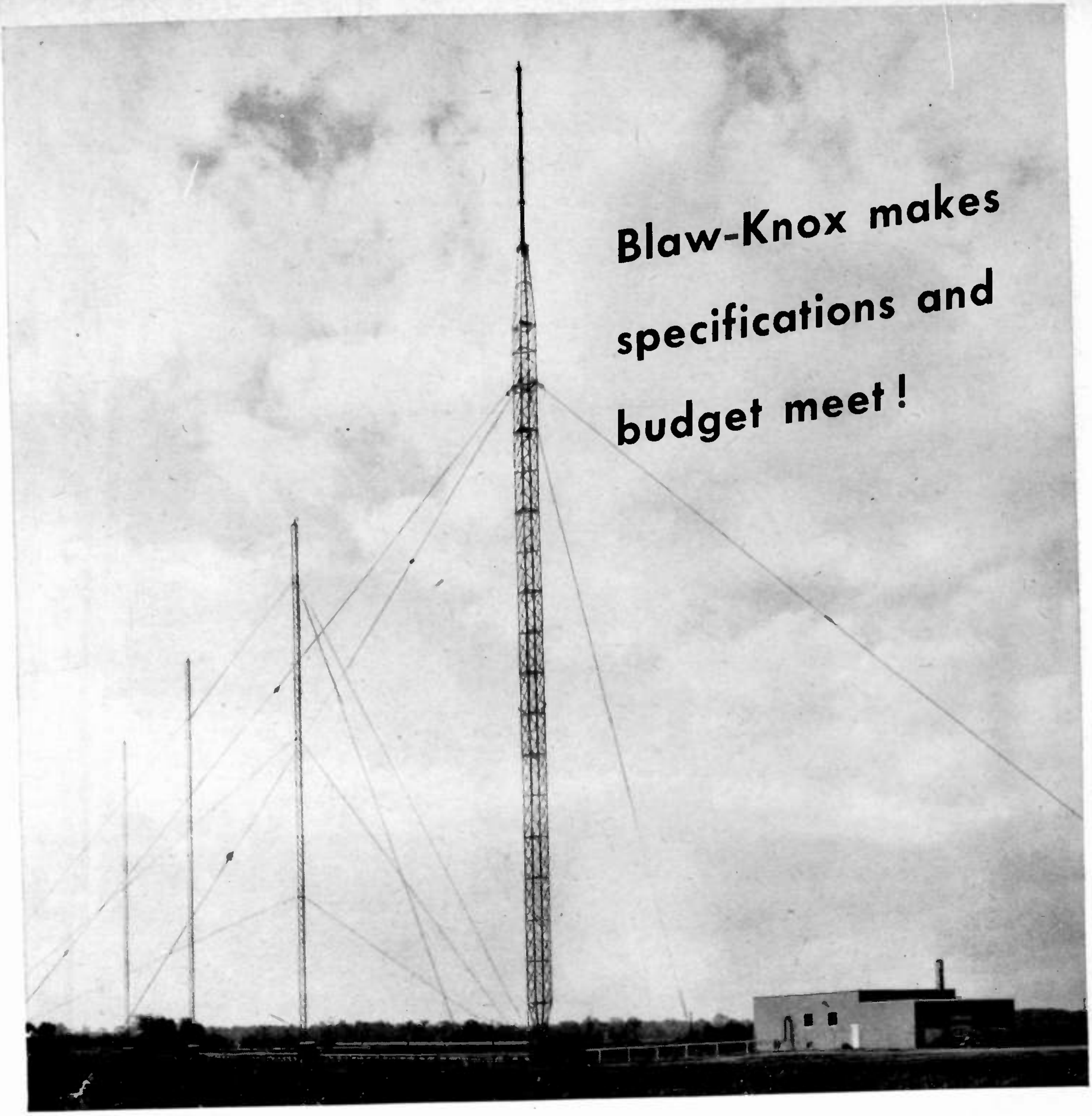

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BLAW-KNOX ANTENNA TOWERS

PROCEEDINGS OF THE I.R.E.

(Including the WAVES AND ELECTRONS Section)

Published Monthly by

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VOLUME 37 — *May, 1949* — NUMBER 5

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

WAVES AND ELECTRONS SECTION

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

Harold A. Zahl

DIRECTOR, 1949

Harold A. Zahl, director of research of the U. S. Army Signal Corps Engineering Laboratories at Fort Monmouth, N. J., was born in Chattsworth, Ill., on August 24, 1904. In 1927 he received the B.A. degree in physics and mathematics from North Central College, followed by the M.S. and Ph.D. degrees from the State University of Iowa in 1929 and 1931.

Joining the Signal Corps as a physicist in 1931, Dr. Zahl participated for the next eleven years in many of the Corps' research and development projects, including work on sound, infrared, electron tubes, radar, and so on. He was intimately connected with radar from its inception, and he was also responsible for research leading to a number of electron-tube types used in military equipment during World War II. In 1939 he became an Associate Member of the IRE.

In 1942 Dr. Zahl entered active military service as a major, and was promoted to the rank of lieutenant colonel three years later. As an army officer, he continued to serve with the Signal Corps Laboratories, dividing his time between technical and administrative matters pertaining to the development of electronic equipment and components for use by the armed forces. He was later awarded the Legion of Merit for his contributions to the fields of electron tubes and radar.

Upon leaving the Army in the spring of 1946, Dr. Zahl re-entered the Signal Corps Engineering Laboratories as a civilian, and in that year was also transferred to Senior Membership in the IRE. His first assignment at the Laboratories was connected with the atom bomb tests at Bikini during the summer of 1946.

A member of the American Physical Society, the New York Academy of Sciences, Gamma Alpha, and Sigma Xi, Dr. Zahl is a member of the board of directors of the Armed Forces Communications Association. He is the author of many scientific articles dealing with various research investigations conducted in the field of physics and electronics, and holds a number of patents pertaining to electron tubes, radar, communications, and the like.

Proposed Standard Frequency-Band Designations

The following proposed Standard Frequency-Band Designations are presented by the Standards Committee of the IRE in order that comments on them may be received from all interested parties prior to final standardization. Comments should be mailed to the Chairman of the Standards Committee, Professor J. G. Brainerd, at Institute Headquarters, 1 East 79 St., New York 21, N. Y. Final standardization will be considered in August, 1949.

At various times during the past several years, there have been proposals to establish standard frequency-band designations. In 1945, the Standards Committee of the Institute adopted a decade system which was described in the August, 1945, issue of the PROCEEDINGS OF THE I.R.E. In this system the band number indicated the exponent of the number ten corresponding to the midfrequency, e.g., band 6 centered on 10^6 cycles with limits of 300 kc and 3 Mc.

In the May, 1947, issue of *Electrical Engineering* an alternative system was described, proposed by H. M. Turner, in which the band number was the exponent of ten corresponding to the lower frequency limit, e.g., band 1 covered a range of 10–100 cps (one exception was that band 0 covered 0–10 cps, rather than 1–10 cps).

During the above period several other methods of designating portions of the frequency spectrum have been evolved, using letters or names, such as uhf band, etc., with the result that at the present time there exists no method of band designation accepted by the majority of engineers. In order to resolve this confusion, a re-study of the problem was authorized by the Standards Committee at its meeting on September 9, 1948.

The objective of a system of band designations is:

To provide a form of shorthand terminology which gives a general idea of what part of the frequency spectrum is referred to without stating exact frequencies, for use in papers, articles, oral presentations, text books, instruction books, etc.

The desirable characteristics of such a system are:

a. That it instantly convey to the listener or reader an idea of the frequencies pertinent to the subject at hand with minimum effort on the listener's part (no memory feat should be involved).

b. That it be readily extensible to bands either up or down in the frequency spectrum.

It is not anticipated that such a system shall rule out use of other designations, such as audible region, microwave region, infrared region, etc. There will also undoubtedly continue to be need for more specific band designations within one or more of these standard bands. It is felt, however, that the following system should contribute materially to uniformity in our written and spoken references.

The proposed Standard Band Numbers and corresponding spectrum limits are given in the following table.

Band No.	*Frequency Range*	*Approximate Equivalent Wavelength*
0	10^0–10^1 C (1–10 cycles)	3×10^8–3×10^7 m (300–30 megameters)
1	10^1–10^2 C (10–100 cycles)	3×10^7–3×10^6 m (30–3 megameters)
·		
·		
5	10^5–10^6 C (100–1000 KC)	3000–300 m (3000–300 meters)
6	10^6–10^7 C (1–10 MC)	300–30 m (300–30 meters)
·		
·		
15	10^{15}–10^{16} C (1–10 KMMC)	3×10^{-7}–3×10^{-8} m (300–30 millimicrons)

Notes:

1. The lower frequency is inclusive, upper exclusive. Thus 10 cycles belongs in Band 1.
2. The system may be extended below 1 cycle by the use of negative band numbers. Thus Band No. −1 covers 0.1–1 cycles.
3. Used as an adjective, the word "Band" shall precede the number; thus, "Band 3 · · · ."

The following table gives proposed standard units for specific frequency designations:

Unit	*Abbreviations*
Cycle	C
Kilocycle	KC
Megacycle	MC
Kilomegacycle	KMC
Megamegacycle	MMC

(Note—The "per second" is implied in the above frequencies.)

Megameter	Mm
Kilometer	Km
Meter	m
Centimeter	cm
Millimeter	mm
Micron	μ
Millimicron	mμ
Micromicron	$\mu\mu$

Theoretical Limitations on the Rate of Transmission of Information*

WILLIAM G. TULLER†, SENIOR MEMBER, IRE

Summary—A review of early work on the theory of the transmission of information is followed by a critical survey of this work and a refutation of the point that, in the absence of noise, there is a finite limit to the rate at which information may be transmitted over a finite frequency band. A simple theory is then developed which includes, in a first-order way, the effects of noise. This theory shows that information may be transmitted over a given circuit according to the relation

$$H \leqq 2BT \log (1 + C/N),$$

where H is the quantity of information, B the transmission link bandwidth, T the time of transmission, and C/N the carrier-to-noise ratio. Certain special cases are considered, and it is shown that there are two distinctly different types of modulation systems, one trading bandwidth linearly for signal-to-noise ratio, the other trading bandwidth logarithmically for signal-to-noise ratio.

The theory developed is applied to show some of the inefficiencies of present communication systems. The advantages to be gained by the removal of internal message correlations and analysis of the actual information content of a message are pointed out. The discussion is applied to such communication systems as radar relays, telemeters, voice communication systems, servomechanisms, and computers.

I. Introduction

THE HISTORY of this investigation goes back at least to 1922, when Carson,[1] analyzing narrow-deviation frequency modulation as a bandwidth-reduction scheme, wrote "all such schemes are believed to involve a fundamental fallacy." In 1924, Nyquist[2] and Küpfmüller,[3] working independently, showed that the number of telegraph signals that may be transmitted over a line is directly proportional to its bandwidth. Hartley,[4] writing in 1928, generalized this theory to apply to speech and general information, concluding that "the total amount of information which may be transmitted . . . is proportional to the product of the frequency range which is transmitted and the time which is available for the transmission." It is Hartley's work that is the most direct ancestor of the present paper. In his paper he introduced the concept of the information function, the measure of quantity of information, and the general technique used in this paper. He neglected, however, the possibility of the use of the knowledge of the transient-response characteristics of the circuits involved. He further neglected noise.

In 1946, Gabor[5] presented an analysis which broke through some of the limitations of the Hartley theory and introduced quantitative analysis into Hartley's purely qualitative reasoning. However, Gabor also failed to include noise in his reasoning.

The workers whose papers have so far been discussed failed to give much thought to the fact that the problem of transmitting information is in many ways identical to the problem of analysis of stationary time series. This point was made in a classical paper by Wiener,[6] who did a searching analysis of that problem which is a large part of the general one, the problem of the irreducible noise present in a mixture of signal and noise. Unfortunately, this paper received only a limited circulation, and this, coupled with the fact that the mathematics employed were beyond the off-hand capabilities of the hard-pressed communication engineers engaged in high-speed wartime developments, has prevented as wide an application of the theory as its importance deserves. Associates of Wiener have written simplified versions of portions of his treatment,[7,8] but these also have as yet been little accepted into the working tools of the communication engineer. Wiener has himself done work parallel to that presented in this paper, but this work is as yet unpublished, and its existence was learned of only after the completion of substantially all the research reported on here. A group at the Bell Telephone Laboratories, including C. E. Shannon, has also done similar work.[9,10,11]

II. Definitions of Terms Frequently Used

Certain terms are used in the discussion to follow which are either so new to the art that accepted definitions for them have not yet been established, or have

* Decimal classification: 621.38. Original manuscript received by the Institute, September 7, 1948; revised manuscript received, February 3, 1949. This paper is based on a thesis submitted in partial fulfillment of the requirements of the degree of Doctor of Science at the Massachusetts Institute of Technology. It was supported, in part, by the Signal Corps, the Air Matériel Command, and the Office of Naval Research.

† Melpar, Inc., Alexandria, Va.

[1] J. R. Carson, "Notes on the theory of modulation," Proc. I.R.E., vol. 10, p. 57; February, 1922.

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

[3] K. Küpfmüller, "Transient phenomena in wave filters," *Elek. Nach. Tech.*, vol. 1, p. 141; 1924.

[4] R. V. L. Hartley, "Transmission of information," *Bell Sys. Tech. Jour.*, vol. 7, p. 535–564; July, 1928.

[5] D. Gabor, "Theory of communication," *Jour. I.E.E.* (London), vol. 93, part III, p. 439; November, 1946.

[6] N. Wiener, "The extrapolation, interpolation and smoothing of stationary time series," National Defense Research Council, Section D_2 Report, February, 1942.

[7] N. Levinson, "The Wiener (RMS) error criterion in filter design and prediction," *Jour. Math. Phys.*, vol. 25, no. 4, p. 261; 1947.

[8] H. M. James, "Ideal frequency response of receiver for square pulses," Report No. 125 (v-12s), Radiation Laboratory, MIT, November 1, 1941.

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

[10] C. E. Shannon, "Communication in the presence of noise," Proc. I.R.E., vol. 37, pp. 10–22; January, 1949.

[11] The existence of this work was learned by the author in the spring of 1946, when the basic work underlying this paper had just been completed. Details were not known by the author until the summer of 1948, at which time the work reported here had been complete for about eight months.

been coined for use in connection with the research here reported. The definitions used in this paper for those terms are given below for the convenience of the reader. No justification of the choice of terms or of the definitions will be given at this point, since it is hoped that this justification will be provided by the bulk of the paper. Terms used in the body of the paper which are not defined below and are peculiar to the jargon of radio engineers will be found in the various "Standards" reports published by The Institute of Radio Engineers.[12]

Information Function—The information function is the function (generally instantaneous amplitude of a current or voltage as a function of time) to be transmitted by electrical means over the communication systems to be analyzed.

Intersymbol Interference—Intersymbol interference is the disturbance present in a signal caused by the energy remaining in the transient following the preceding signal.

Coding—Coding is the representation of the information function by a symbol or group of symbols bearing a definite mathematical relation to the original function, and containing all the information contained in the original function.

Binary Coding—Binary coding is coding in which the instantaneous amplitude of the information function is represented by a sequence of pulses. The presence or absence of these pulses at certain specified instants of time represents a digit (either one or zero) in the binary system of numbers.

Clearing Circuit—The clearing circuit is a circuit which will clear intersymbol interference from the output of a filter.

Quantized—A variable is said to be quantized when it varies only in discrete increments.

III. Communication System Transmission Characteristics

In general, the information which one wishes to transmit over a communication system is supplied to the system (neglecting any transducers which may be present to transfer the energy from its source to the electrical system) in the form of a time-varying voltage (or current), lasting for a period T.

For the purposes of the paper, it will be assumed that a width of pass band f_c has been selected, and the lower and upper limits of this band fixed. It will be assumed that all frequency components of the information function lying within this pass band are to be transmitted without distortion of any type, and that all frequencies outside the limits of the pass band are completely unimportant and may be suppressed an arbitrary large amount, enough to make them negligible. These assumptions are, it is realized, somewhat arbitrary, since a system satisfying them would cause considerable transient distortion of certain types of information function. However, these assumptions will serve as a first approximation.

[12] In particular, "Standards on Antennas, Modulation Systems, and Transmitters: Definitions of Terms, 1948."

IV. Definition of Quantity of Information

It must, of course, be recognized that any physical transmission system will have an upper bound to the amplitude of information function it will transmit. The accuracy of specification of the information function at any given time may be specified in terms of this maximum value. Thus, within the range of possible values of the information function, there will be a number s of these values that are significant. Similarly, if the information is examined over a period of time of length T, there will be a number n of times at which samples of the function may be taken and yet the information will be unchanged, since the function may be recreated from a knowledge of its values at these intervals. It is known from the work of Bennett,[13] and others, that n must be greater than $2f_cT$, using the nomenclature of the previous sections, in order to recreate exactly any arbitrary function. Considering the continuous information function shown in Fig. 1, the second statement given above permits us to consider its values only at specified, and in this particular case equispaced, intervals of time, as is shown in the solid stairstep curve. The statement of

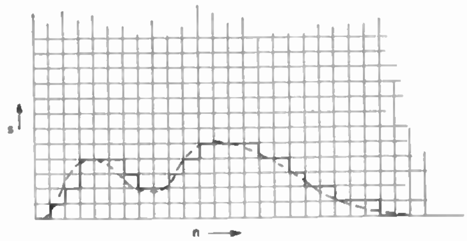

Fig. 1—The information function in n,s space.

the finite number of significant values of the function allows us to consider only certain discrete amplitudes, separated from each other by twice the error of specification. The information function may thus be redrawn so as to follow only certain lines in a rectangular co-ordinate system. Such a function is called quantized, since it takes on values chosen from a discrete set. A plot of such a function quantized, and drawn in n,s space, is also given in Fig. 1.

The question now before us is, "What is the information content of a function in the n,s plane?" The answer of Hartley is the "quantity of information" given by

$$H = kn \log s \qquad (1)$$

where k is a proportionality constant. The reasons for Hartley's choice may be expressed in a straightforward manner on the basis of two fundamental requirements of a definition of "quantity of information." These are:

(a) Information must increase linearly with time. In other words, a two-minute message will, in general, con-

[13] W. R. Bennett, "Time-division multiplex systems," *Bell Sys. Tech. Jour.*, vol. 20, pp. 199–222; April, 1941.

tain twice as much information as a one-minute message.

(b) Information is independent of s and n if s^n is held constant. This states that the information contained in a message in a given n,s plane is independent of the course of the information function in that plane, allowing only single-valued functions. With this restriction, the number of different messages that may occupy a given n,s plane is s^n. Transmission of one of these messages corresponds to making one of s^n choices. Stating that information is independent of s and n if s^n is constant means that we gauge quantity of information by the number of possible alternatives to a given message, not by its length or number of possible values at any given instant of time.

On the basis of these two requirements, it can be shown that (1) is the only possible definition of quantity of information.

V. Transmission of Information in a Noise-Free Universe

The preceding discussion has been set up on the basis of a system with noise, and indeed this is the most practical arrangement. However, most of the previously published works on the transmission of information have neglected noise, and come out with the conclusion that even in the absence of noise there is a limit, in any system containing elements capable of storing energy, to the rate at which information may be transmitted. This theory has been widely—in fact, almost universally—accepted by communication engineers. It is, therefore, believed worth while to show by example that this theory is incorrect, even though the correct theory to be derived later in this paper shows implicitly the error in the previous theories. The basic fact that has been neglected in earlier analyses and which resulted in their errors is that the output wave form of a network is completely determined for all time by the input wave form and the characteristics of the network. A method of utilizing this effect in practice is outlined in the follow-paragraphs.

Suppose that we choose to transmit the information by a series of modulated pulses according to any of the well-known methods of pulse modulation. If these pulses are passed through a filter that one would suspect as having too narrow a pass band to faithfully reproduce the pulses, there will result intersymbol interference, as shown in Fig. 2. That is, energy stored in the filter from the first pulse will appear at the output of the filter during the time at which the second, third, and all succeeding pulse outputs are present. However, if we know the shape of the pulses and the transient response of the filter, we may transmit our intelligence in the following manner, theoretically. Let us first transmit the pulse to be used as a standard of comparison. The output from the filter resulting from this pulse will be measured for a period of time sufficient to determine exactly the amplitude of the initial pulse. This may be done since, as was mentioned above, the output amplitude is uniquely determined by the input amplitude. Knowing the wave form of the output from our knowledge of system characteristics and the amplitude of the output from measurements, we may compute the exact output voltage to be obtained from the system at any time in the future.

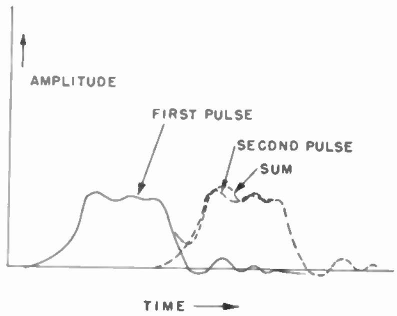

Fig. 2—Response of communication system to rectangular pulse.

This voltage wave form may be generated locally and subtracted electrically from the output of the filter. Alternatively, the output wave form may be recorded graphically, the component due to the first pulse computed, and this component subtracted by graphical methods, to give the system output free of intersymbol interference caused by this pulse. Either method may be applied repeatedly to remove intersymbol interference from the following pulses. Other methods may, no doubt, be used, but these two serve to indicate the problem involved.

To formalize the argument, suppose we are given an arbitrary signal $f(t)$ and an arbitrary filter pass band f_c. We wish to transmit an arbitrarily long message at the rate of one per second. Let us assign to the mth message we wish to transmit an integer M_m characterizing the message in a one to one manner. This number M_m may, for example, be derived from the number in the binary system of units corresponding to the message as transmitted in ordinary continental Morse Code, using a one to correspond to signal of unit length. In this fashion,

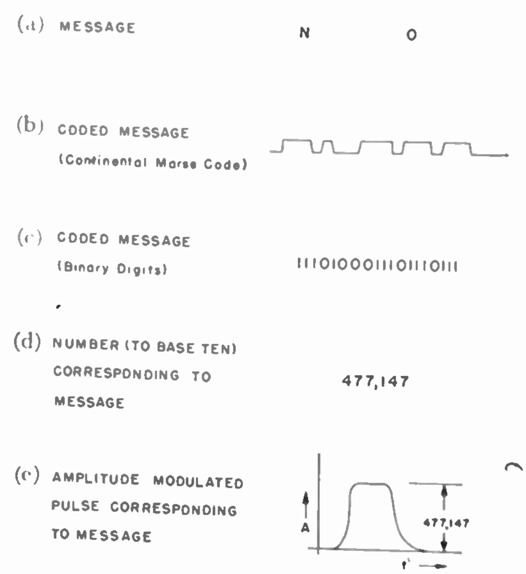

Fig. 3—Representation of message by an integer.

a one-to-one correspondence may be realized between some number (in the binary system of units in this case) and our message. Referring to Fig. 3, let us consider transmission of the message "NO." Fig. 3(a) shows the message in English, 3(b) in Continental Morse Code, 3(c) in binary digits, and 3(d) gives the number to the base ten that corresponds to our message. Fig. 3(e) shows that all the information contained in the message may be contained in one pulse, of arbitrary duration and 477,147 units in amplitude. The pulse of Fig. 3(e) may be used, then, as our message.

There have been three types of pulses mentioned in the preceding paragraphs, perhaps causing a certain amount of confusion. To distinguish among them, let us reconsider for a moment. The first pulse mentioned was a typical pulse in a pulse-modulation system, used as an example to show how intersymbol interference might be eliminated. The second pulses mentioned were those forming our message in Morse Code, used in a typical process for obtaining a one-to-one correspondence between a message and an integer M_m. The third pulse mentioned was one of arbitrary duration and M_m units in amplitude. It may, therefore, be used as our message. It may, further, be used as a channel pulse in a pulse-amplitude-modulation communication system, since the information it carries is solely contained in its amplitude.

To return to our original argument, we are given a wave form $f(t)$. Let $F(t)$ be the response of our channel, of bandwidth f_c, to $f(t)$. At time $t=0$ we transmit $M_0 f(t)$ where M_0 is a known calibrating amplitude. To calibrate our system we measure the voltage in the receiver channel at some time t_0. We may call this voltage $n_0\ F(t_0)$. Since we know $F(t)$ for all values of t, we know it for t_0. From this and our measurement we obtain n_0, the demodulated voltage at the output of the system corresponding to a modulating voltage M_0.

We now introduce the voltage $-n_0F(t)$, for t greater than t_0, into the output of our system. This clears the channel completely. We may then take a voltmeter reading at t_0+1, from which we get M_1. This second message may then be cleared from our system by the same procedure as used previously and the same process repeated. This may now go on until $t=t_m$. At this time we measure the voltage $n_mF(t_m)$. All energy that would ordinarily have been present at this time, because of intersymbol interference, has been eliminated by the clearing process. We know $F(t)$ for all values of t, so we know it for t_m. From this and the previously measured relation between n_0 and M_0, we may obtain successively n_m and M_m. Thus our message M_m is obtained regardless of the intersymbol interference present, and therefore regardless of f_c, providing only that we may measure to negligible error and that our system characteristics are accurately known.

A semipractical arrangement for accomplishing this is shown in block form in Fig. 4. Fig. 5 shows some of the wave forms involved.

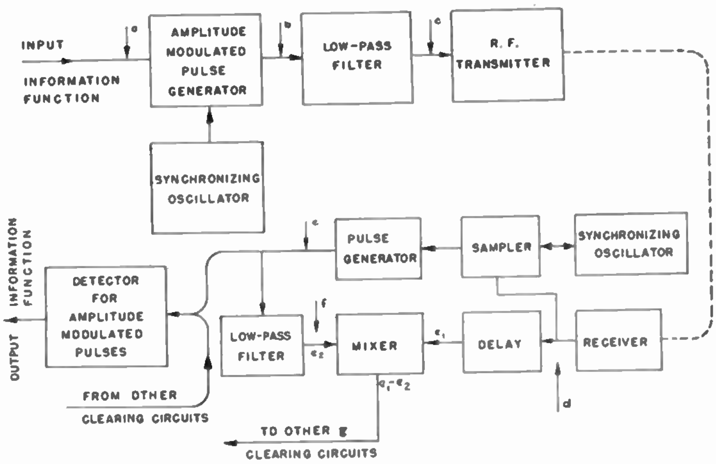

Fig. 4—Block diagram of a narrow-band communication system.

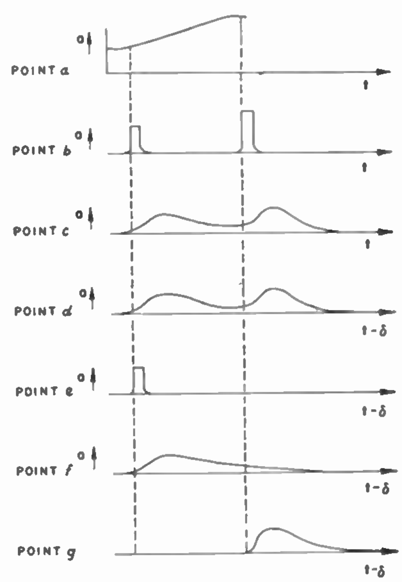

Fig. 5—Wave forms in the narrow-band communication system.

It has been observed by critics of the system proposed above that this scheme is theoretically unworkable, since in the limit each pulse will have an infinite precursor, which will not be eliminated by the clearing circuits. However, in this limiting case, the system has an infinite delay also. Therefore, the precursor begins at the time of transmission of our first pulse, and, though small, may be measured at this time. The components due to other pulses cannot exist before these pulses are impressed on the system, and hence cannot affect the measurement of the precursor to the first pulse. Once the precursor to the first pulse of the system has been measured, the amplitude of this first pulse is known, and its effects may be cleared from the system. The system proposed here does not, it should be emphasized, depend on the early results of Nyquist, who pointed out that systems can be constructed which are self-clearing at certain times, and hence can transmit a very large amount of information at these times. The system proposed here can transmit information at a rate only limited by the delay one wishes to tolerate. As has been pointed out by several workers, long delay is an unavoidable concomitant of any system conserving bandwidth to the utmost. This does not, however, affect the rate of transmission of information, the quantity under consideration here.

As a result of the considerations given above, we are led to the conclusion that the only limits to the rate of transmission of information on a noise-free circuit are economic and practical, not theoretical.

VI. Transmission of Information in the Presence of Noise

In some ways the discussion of the section immediately preceding this one represents a digression in the main argument to be continued below. It may be well, therefore, to review the main argument at this point, and to indicate the direction it is to take. So far, Hartley's definition of information has been investigated and shown adequate for this analysis. The early theories of transmission of information have been refuted. In the portion of the work that follows, a modified version of the Hartley law applicable to a system in which noise is present is derived. This is done for the general case and for two special types of wide-band modulation systems, uncoded and coded systems. As a result of these analyses the fundamental relation between rate of transmission of information and transmission facilities is derived.

Since we have shown that intersymbol interference is unimportant in limiting the rate of transmission of information, let us assume it absent. Let S be the rms amplitude of the maximum signal that may be delivered by the communication system. Let us assume, a fact very close to the truth, that a signal amplitude change less than noise amplitude cannot be recognized, but a signal amplitude change equal to noise is instantly recognizable.[14] Then, if N is the rms amplitude of the noise mixed with the signal, there are $1+S/N$ significant values of signal that may be determined. This sets s in the derivation of (1). Since it is known[13] that the specification of an arbitrary wave of duration T and maximum component f_c requires $2f_cT$ measurements, we have from (1) the quantity of information available at the output of the system:

$$H = kn \log s = k2f_cT \log (1 + S/N). \tag{2}$$

This is an important expression, to be sure, but gives us no information in itself as to the limits that may be placed on H. In particular, f_c is the bandwidth of the over-all communication system, not the bandwidth of the transmission link connecting transmitter and receiver. Also, S/N may not at this stage of the analysis have any relation to C/N, the ratio of the maximum signal amplitude to the noise amplitude as measured before such nonlinear processes as demodulation that may occur in the receiver. It is C/N that is determined by power, attenuation, and noise limitations, not S/N. Similarly, it is bandwidth in the transmission link that is scarce and expensive. It is, therefore, necessary to bring both these quantities into the analysis and go beyond (2).

The transmission system assumed for the remainder of this analysis is shown in block diagram in Fig. 6. The elements of this system may be considered separately.

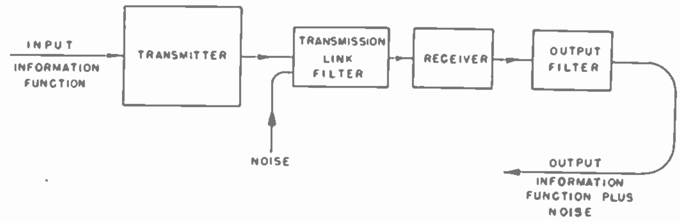

Fig. 6—Block diagram of the simplified communication system used in the analysis.

The transmitter, for example, is simply a device that operates on the information function in a one-to-one and reversible manner. The information contained in the information function is preserved in this transformation.

The receiver is the mathematical inverse of the transmitter; that is, in the absence of noise or other disturbance, the receiver will operate on the output of the transmitter to produce a signal identical with the original information function. The receiver, like the transmitter, need not be linear.

It is assumed throughout the remainder of this analysis, however, that the difference between two carriers of barely discernible amplitude difference is N, regardless of carrier amplitude. This corresponds to an assumption of over-all receiver linearity, but does not rule out the presence of nonlinear elements within the receiver. This assumption is convenient but not essential. If it does not hold, the usual method of assuming linearity over a small range of operation and cascading these small ranges to form the whole range may be used in an entirely analogous analysis with essentially no change in method and only a slight change in definition of C/N and S/N, here assumed to be amplitude-insensitive.

The filter at the output of the receiver is assumed to set the response characteristic of the transmission sys-system. (It should be noted that, when "transmission system" is referred to, all the elements shown in Fig. 6 are included. "Transmission link" refers only to those elements between the output of the transmitter and the input to the receiver.) The transmission characteristics of this filter are, therefore, those previously given for the over-all transmission system. Coming now to the elements of the transmission link, consider first the filter which sets the link's transmission characteristics. The phase shift of this filter is assumed to be linear with respect to frequency for all frequencies from minus to plus infinity. The over-all attenuation is assumed to be zero decibels at all frequencies less than B, and is assumed to be so large for all frequencies above B that energy passing through the system at these frequencies is small

[14] This assumption ignores the random nature of noise to a certain extent, resulting in a theoretical limit about 3 to 8 db above that actually obtainable. The assumption is believed worth while in view of the enormous simplification of theory obtained. For a more precise formulation of the theory, see footnote references 9 and 10.

in comparison with the unwanted disturbance, or noise, present in the output of the system. It should be obvious that this characteristic may be made band-pass or high-pass by the well-known transformations.

The noise is assumed to have a power spectrum whose amplitude is uniform over the range of frequencies passed by the filter in the transmission link. The noise spectrum is, therefore, identical with that of the pass band of the receiver.

From the above discussion, it is apparent that the transmission system sketched in Fig. 6 is a close approximation to most communications systems in which only one source of noise is important. The transmitter can be anything from a pair of wires connecting input and output up to and beyond a pulse-code-modulation generator modulating a high-frequency carrier. The lumping of noise into one generator, lumping the transmission characteristics of the link into one filter, and lumping the transmission characteristics of the over-all system into one other filter, as well as the assumption of linearity, are admitted to be unreal assumptions which, however, come reasonably close to the true facts, close enough for engineering purposes in many instances. The special shapes of the transmission characteristics assumed are, as has been mentioned, chosen for convenience and not necessity.

In addition to the general case, it is interesting to consider two special cases:

1. *Uncoded*—To every specified and unique point in the information function, there corresponds one specified and unique point in time of the information function as transformed by the transmitter. There are the same number of points in the transformed as in the original function. Information is conserved. The over-all time taken for transmission may, but need not, be equal at the input and output of the transmitter.

2. *Coded*—While information is conserved in this case also, one point in the transformed information function may, in this case, be specified so accurately as to contain all the information contained in a whole series of rather inaccurately specified points in the original information function, or vice versa. The transformation again must be reversible.

The uncoded transmission corresponds to direct transmission of the information function, transmission of an amplitude-modulated carrier, transmission of a frequency-modulated carrier, and the like. Coded transmission corresponds to pulse-code modulation or other similar systems.

The points to be shown about the three types of transmission are as follows:

1. In general, for large signal-to-noise ratios, the signal-to-noise ratio may be equal to or less than the carrier-to-noise ratio *raised* to the *power* B/f_c. (See (7).)

2. Coded transmission is capable of realizing the fullest capabilities of the general system; i.e., signal-to-noise ratio may equal carrier-to-noise ratio *raised* to the *power* B/f_c, for large signal-to-noise ratios. (See (18).)

3. In uncoded transmission the signal-to-noise ratio may be equal to or less than the carrier-to-noise ratio *multiplied* by the *factor* B/f_c, for large signal-to-noise ratios. (See (20).)

These points will be shown separately in the order given.

Let us first consider the general system. In this case, making the assumption that a change in carrier voltage equal to rms noise is just detectable, and applying the reasoning that led to (2) to the receiver input, we have, for the "quantity of information" at the receiver input,

$$H_{\text{in}} = k \cdot 2BT \log (1 + C/N). \tag{3}$$

We know from (2) that the "quantity of information" at the output of the receiver is

$$H_{\text{out}} = k \cdot 2f_cT \log (1 + S/N). \tag{4}$$

The receiver cannot be a source of information. By this we imply that to every value of C/N there corresponds one and only one value of S/N. This must be true if the system is to operate in the absence of noise, since otherwise there might correspond more than one value of S for a given C, an unworkable situation. We may, however, lose information in the receiver; i.e., it may not be perfect. Allowing for this,

$$H_{\text{out}} \leqq H_{\text{in}}. \tag{5}$$

Substituting (3) and (4) in (5) and clearing like quantities and logarithms from both sides of the inequality gives

$$(1 + S/N) \leqq (1 + C/N)^{B/f_c}. \tag{6}$$

Or, if $C/N \gg 1$ and $S/N \gg 1$,

$$S/N \leqq (C/N)^{B/f_c}. \tag{7}$$

Let us now consider coded transmission. In this case, as will be shown by an example, the equals sign of (6) may be achieved. Suppose, for example, we wish to transmit the message of Fig. 3(e). Clearly, this requires a carrier-to-noise ratio of at least 477,146 if it is to be transmitted as an amplitude-modulated pulse. Suppose, however, a carrier-to-noise ratio of but unity is available, so that the best we can do is distinguish between carrier off and carrier on. In this case we may still transmit the message in the form of Fig. 3(b), essentially coding it in binary digits. In this case, if we wish the message to be transmitted in the same time, we must transmit it in nineteen times as much bandwidth, since we must transmit nineteen time units during the duration of our message, instead of the original single pulse or time unit. At the receiver the pulses of Fig. 3(b) may be deciphered to form the single pulse of Fig. 3(e). One must be careful how he uses these data to avoid error. The various quantities of (6) might erroneously be considered to be, for this example,

$$1 + S/N = 477,147 \tag{8}$$

$$1 + C/N = 2 \tag{9}$$

$$B/f_c = 19. \tag{10}$$

We find, however, that

$$2^{19} = 522,288, \tag{11}$$

and therefore, in this case,

$$(1 + S/N < (1 + C/N)^{B/f_c}. \tag{12}$$

This does not correspond to the earlier statement that the equals sign of (6) can be realized. However, one message that could be sent over our system would be a long dash of 19 time units duration. The number corresponding to this dash would be 522,287. This fact gives a clue to the application of (6). In using this formula, $(1+S/N)$ must be the number of possible allowed states of the receiver output at any one time, and $(1+C/N)$ the number of possible allowed states of the receiver input for any one instant of time. In other words, $(1+S/N)$ is actually s, measured at the output of the receiver, and $(1+C/N)$ is s, measured at the input to the receiver. Considering things in this correct manner, we have

$$(1 + S/N) = 522,288 \tag{13}$$

$$(1 + (C/N) = 2 \tag{14}$$

$$B/f_c = 19 \tag{15}$$

$$2^{19} = 522,288 \tag{16}$$

$$(1 + C/N)^{B/f_c} = (1 + S/N). \tag{17}$$

If $C/N \gg 1$ and $S/N \gg 1$,

$$(C/N) = (S/N)^{B/f_c}. \tag{18}$$

Thus, we see that in this manner, at least, the equals sign of (6) may be realized.

Now let us consider uncoded transmission. In this case, one and only one functional value is specified for each original time value. That is to say, the total number of samples taken of the wave is maintained constant. However, to each unit, or quantum, of time in the original information function there correspond B/f_c resolvable units, or quanta, of time on the transmitted information function. This was, of course, also true in the case of the coded transmission. Now, however, we specify that during all but one of these B/f_c units the function be zero. During each of these periods we may have the carrier-to-noise ratio C/N and hence a possible range of values $(1+C/N)$.

Corresponding to our original possibility of one point on the original information function with any of $(1+C/N)$ possible significant amplitudes, we now have a possibility of one point[15] with any one of B/f_c times $(1+C/N)$ possible significant values. This comes about because the point may have $(1+C/N)$ possible significant amplitudes and, in consequence of the improvement in system resolution capability by the factor B/f_c, B/f_c possible independent time values. We have, therefore, increased the number of degrees of freedom of each point by the factor B/f_c. This argument is illustrated in Fig. 7, showing the original and transformed signals in amplitude-time and frequency-time space. We can, therefore, make one of B/f_c times C/N possible choices for this point. We have, therefore, as the number of possible independent states for the receiver over any period of $1/2f_c$ seconds,

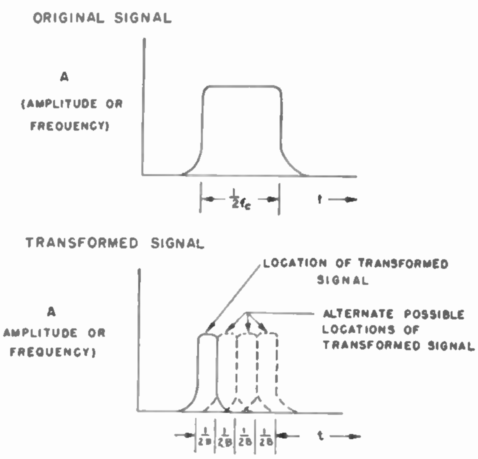

Fig. 7—Uncoded transformation.

$$(1 + S/N) \leqq B/f_c \cdot (1 + C/N). \tag{19}$$

Or, again, if $S/N \gg 1$ and $C/N \gg 1$,

$$S/N \leqq B/f_c \cdot (C/N). \tag{20}$$

Again the equality can be realized, as in the well-known frequency-modulation system.[16]

It is interesting to apply the results of (18) and (19) to the noise-free system. In this case, C/N is infinite. Therefore, if H is a very large but finite quantity, either T or B may equal zero. Thus we are led directly to the refutation of the early theories of transmission of information.

VII. The Use of Coded Information Functions

Binary coding is the only type of coding that has received attention in published papers. This type of coding represents the maximum sacrifice in bandwidth requirements and maximum decrease in required carrier-to-noise ratio, since the latter is reduced to unity. Time of transmission is unchanged in this coding system, as is true in most systems designed for two-way transmission.

The publicity given to binary coding, and its value in systems employing regenerative repeaters, have overshadowed the potentialities of other coding schemes. For example, consider the national radio broadcasting system. The standard system employing amplitude modulation has been recently supplemented by a frequency-modulation service offering increased freedom

[15] In contrast to the coded case, we are here forbidden to employ more than one point by the very definition of uncoded transmission.

[16] The absence of the $\sqrt{3}$ from (20) in the frequency-modulation case can be shown to be due to differences in definition of bandwidth between that assumed in this paper (for nonrectangular spectra) and that generally used in the conventional FM analyses.

from noise at a cost, not considering the additional transmitted audio-frequency range provided, of five times the spectrum width per station. Since frequency modulation is an uncoded transmission system, one expects that the signal-to-noise ratio would be improved about five times by this band widening. Suppose we wish to obtain this improvement by coding. We might choose a double-band code system, in which each point on the original information function was replaced by two points on the transformed information function. Taking a signal-to-noise ratio of 1,000 (60 db) as a reasonable figure, this could be accomplished with a carrier-to-noise ratio of but 32 (30 db). A frequency-modulation system similar to the present standard would require a carrier-to-noise ratio of about 120 (42 db) to accomplish this, or about four times the transmitter power, and, moreover, would use 2.5 times the bandwidth.[17] If higher signal-to-noise ratios were required, the difference between the two systems would be even more spectacular.

Coding may also be used to reduce bandwidth or time of transmission at the expense of carrier-to-noise ratio. The bandwidth required for transmission, for example, may be halved by combining the information contained in two points in the information function into one point on the transformed information function. This requires that carrier-to-noise ratio be the square of that required without coding, but does accomplish the required results. For example, if each point were expressed originally to an accuracy of 1 part in 10, the first might be nine units in amplitude and the second three. A system capable of transmission accuracy of 1 part in 100 could transmit one point 93 units high, which is easily recognized as containing all the information present in the previous two. The transmission of this one point might use the time previously required to transmit the previous two, and thus would require but half the bandwidth. There is no theoretical advantage to be gained in using bandwidth for transmission in excess of that required for the transmission of the coded information function; i.e., wide-band modulation systems using uncoded transformations are inherently inefficient in their utilization of spectrum. If a carrier is unnecessary and the signal-to-noise ratio attained by simply transmitting the information function directly is just adequate, then this is the most efficient utilization of spectrum. If a carrier must be used, single-sideband amplitude modulation, narrow-deviation frequency modulation, or some other "narrow-band" modulation system should be used. Coding may be used as desired to gain in one parameter at a sacrifice in some other or others without any loss in efficiency.

It should be pointed out that economic factors not considered in any of the analysis above modify these theoretical considerations. In particular, in the present state of the art, coded transmission is more suitable for point-to-point communication in which the number of transmitters and receivers are equal, rather than for broadcasting, where one transmitter services many receivers. In the latter case, an uncoded "brute force" scheme may be desirable, putting the burden on a big transmitter in order to permit the simplest possible receivers. The existence of such a situation should, however, only point the way to needed improvements in the art.

VIII. The Function with Maximized Information

Until now we have considered a rather general type of information function, limited only by a finite width of spectrum. It is of some importance to consider the amount of actual irreducible information contained in such a function. The transmission of information involves the transmission of one of a set of possible alternative choices. If certain analytic properties of the information function make the selection of a particular choice mandatory at some time, no actual choice is made since no alternatives can exist. Continuation of this line of reasoning, as is shown below, leads to the possibility of reducing the bandwidth required to transmit many types of information function. Further, this analysis shows that one particular type of information function, here called the function with maximized information, conveys the maximum intelligence from one point to another for a given set of transmission facilities. This function has the general characteristics of filtered random noise, except for its distribution function.

To derive the characteristics of this function, let us first consider the definition of "quantity of information." This definition was arrived at by considering a series of n selections, each made from a set of s possible choices. It should be obvious that if, in some selections, we choose from only s-j possible choices, we transmit less information than if s choices were available each time a selection were made. Nothing has been said previously about this point, but it should be recognized that s need not be a constant during a message, but may vary with time.

If s is not a constant, we must provide system facilities adequate to transmit the maximum value of s ever realized in the message. Considering the transmission link of Fig. 1, we have, therefore,

$$C/N \geqq s_{max} - 1. \tag{21}$$

The actual quantity of information contained in a system with variable s is

$$H = kn \log s_{ave} \tag{22}$$

where s_{ave} is the average value of s, obtained in the conventional manner.

Thus, in this case, since

$$s_{max} \geqq s_{ave}, \tag{23}$$

[17] If additional advantage were taken of single-sideband operation, the coded signal could be accommodated in a standard broadcast channel and give the same signal-to-noise ratio as frequency modulation with one-fourth the power and one-fifth the required spectrum.

the formula

$$H = k2BT \log (1 + C/N) \tag{24}$$

no longer holds, and, in fact, we have

$$H \leqq k2BT \log (1 + C/N). \tag{25}$$

To realize the equals sign in (25), we must achieve the equals sign in (23). This can only be done when s is a constant, since only in this case does $s_{max} = s_{ave}$.

We have, therefore, the fact that if s is not constant during a message the transmission of that message will require more time, bandwidth, or power than would be necessary to transmit the same quantity of information in a form in which s were constant. We now ask the implications of this statement. These are that unless, at any instant of sampling, the sample is equally likely to take on any of its allowed significant values, we are wasting time, bandwidth, or power; and further, that every message should be examined in detail for possible long-time-interval coherences before transmission. This implies delay and storage in the transmission of a message, since we can only make sure that s is constant by examining every portion of the complete message.

To carry the argument a step further, suppose the future amplitude of the function may not be exactly determined from a knowledge of the function in the past, but that it may be determined to a certain probability. Then only the range of values having high probability need be transmitted, since, by omitting that range having low probability, power is made available to transmit the high-probability range with greater accuracy. To give a numerical example, suppose that from some knowledge of the information function it is determined that the amplitude of the function will be within 10 per cent of the possible amplitude range at a given instant of sampling, to a probability of 0.9. Suppose the system has an s of 100. In this case a range of ten possible significant amplitudes will have a probability of occurrence of 0.9 and the remaining range of ninety has a probability of 0.1. We may then let the more probable range occupy 50 per cent of the scale, by a prearranged scheme, and express this range in fifty significant steps, rather than just ten. The accuracy of reproduction, so far as this most probable region is concerned, is thereby increased by a factor of five, at the expense of a similar reduction of accuracy of the remainder of the scale. Therefore, 90 per cent of the time the effect is to transmit with s five times as great as was formerly the case; 10 per cent of the time the effect is to reduce s by a factor of five. The average effect is roughly to increase s by 4.5, and the quantity of information that may be transmitted over the system by the logarithm of this quantity.

Another way of stating this requirement of maximized information is to state that there must be no possibility of analytic continuation of the information function to an accuracy of better than one part in s for the duration of the interval between samples. This may readily be arrived at by consideration of the arguments above.

To summarize, any information function which is not one of maximized information will require more time, bandwidth, or power to transmit a given quantity of information than will the maximized information function. It is, therefore, extremely important that, in any transmission requiring maximum efficiency in utilization of these three parameters, one make sure his input wave is one of maximized information. The spectrum of such a wave if the interval between samples is constant is that of white noise passed through an ideal low-pass filter. This is a convenient, although not sufficient, method of assuring that such a function has been obtained.

IX. Application to Other Fields

The point of view developed in the work described above has already been very useful in the analysis of systems not generally considered as belonging to the communications family, but which, as several people have recently come to believe, should be. Typical general fields in which information-transmission problems occur and, in fact, may completely govern system design are radar, radar relay, telemetering, servomechanisms, and computing mechanisms of the digital type. Application of the viewpoint here developed can show possible simplification in system design, unrealized information-handling capability, or the use of a system inefficient in that it supplies more information than is required.

Let us consider the radar problem. At the moment we shall only be concerned with radar search in two dimensions, azimuth and range, although expansion of the theory to three-dimensional search systems offers only slight additional complications. The problem to be solved by the radar is the determination of the existence or nonexistence of a reflecting body at any point within the range of the equipment. A refinement that might be useful, although not always essential, consists in knowing the "electrical size" of the target, i.e., the strength of the reflected signal. Service codes for reporting echo signal strength only recognize five possible strengths, realizing the difficulty of estimating by eye the amplitude of a constantly fluctuating signal on the face of a cathode-ray tube. Therefore, five digits and a zero are enough to tell all the significant facts about signal strength. If the maximum range of the equipment is R, and the desired range accuracy $\pm r$, then there will be a total number $R/2r$ ranges at which a target may be said to be located. Similarly, if search is to be carried out over 360°, to an azimuth accuracy of $\pm B$ degrees, there are $360/2B$ possible azimuth positions in which a target may be located. The total number of integers that must be transmitted to the operator for each complete scan are, therefore, $(R/2+360/2B)$. Each integer has six possible values, ranging from zero to five. A five-to-one carrier-to-noise ratio is, therefore, all that is required. This assumes separate search in range and azimuth. An alternative way of searching is to examine each elemental area bounded by the concentric range-accuracy circles and the radial angular-accuracy lines separately. This involves the transmission of $(R/2r \times 360/2B)$ integers, each having six possible values. Since the quantities involved are such that the second system of scanning always results in the transmission of more information than the first, we may say at once that the first scheme of scanning looks more efficient than the second, and ask why. The answer is not long in coming. The first system of scanning is adequate so long as there are never two targets in the same range-accuracy strip, or the same azimuth-accuracy wedge. If at any time there are two targets in such an area, the system will become confused and will report but one. Taking typical numbers to see the cost of this degree of data separation, R might be 100 miles, r, $\frac{1}{4}$ mile, and B, 1 degree. Then the first system of scanning calls for the transmission of $200+180=380$ integers, while the second calls for the transmission of $200 \times 180 = 36,000$ integers. Therefore, the second system of scanning should require almost one hundred times the bandwidth or transmission time of the first, holding the signal-to-noise ratio constant at five. This is the cost of the freedom from confusion. It would seem that the first type of scanning could advantageously be used for early-warning systems, sited so that target confusion is unlikely. The resultant decrease in information-handling capacity required of the system could be taken advantage of in system design to make possible the use of lower power or narrow bands or decreased search time. On the other hand, these parameters could be held constant and the effective range of the system increased until its information-handling capacity was fully utilized.

If we assume that target confusion is highly probable, then it may be worth while to examine the second system of scanning in some detail, to see if modern radar systems are as efficient as they might be. As we have observed, some 36,000 integers must be transmitted during each complete scan. Each of these integers must be transmitted during each complete scan. Each of these integers has one of six possible values. Suppose, as is reasonable, that we wish to scan the complete area under surveillance once per minute. The information must then be transmitted at a rate of 600 integers per second. A bandwidth of 300 cps is all that is required to transmit this information at a five-to-one signal-to-noise ratio. Other signal-to-noise ratios may be accommodated, or taken advantage of, by coding, with a resultant change in required bandwidth. The high speed of scan in range, forced by the velocity of propagation of radio waves, is immutable in radar systems of the pulsed type, and therefore one must accept a wide band of frequencies. It should not be necessary, however, to force the indicator circuitry to respond at this speed. If a delay and storage circuit is provided, it could accept signal information in short bursts and disgorge this information at the uniform rate of 600 integers per second, for use by the radar indicator and the operator.

Considering now a remote indicator or relay for the radar system outlined above, one could be provided with a bandwidth of but 300 cps and a minimum signal-to-noise ratio acceptable for good results of but five, using techniques already at hand. The storage could perhaps be provided by the conventional long-persistence-screen cathode-ray tube used as a radar-system indicator, with pick-off of data provided by television or facsimile methods, at very low scan rates. The cases considered here do not, it should be mentioned, contemplate relaying more information than that present on the local indicator, in contrast to some other proposed systems.

The telemetering problem is similar in many respects to the radar-relay problem, but simpler in that only one-dimensional information need be transmitted, usually a meter reading. If 1 per cent accuracy is required, the problem is that of transmitting one of fifty integers, at whatever rate is desired. The bandwidth required is, therefore, half the desired rate, and the signal-to-noise ratio a minimum of fifty. If this signal-to-noise ratio is not attainable under the most unfavorable conditions, best spectrum utilization is obtained, not by resorting to uncoded modulation schemes, which have been shown to be inefficient, but in coding the information. This can be carried out down to the point where a signal-to-noise ratio of unity is adequate for the accuracy of data transmission required, but in this case the bandwidth or time of transmission required must be increased by a factor of almost six, for most cases. To illustrate the gain in efficiency of this method over the various conventional but uncoded wide-band schemes, it need only be noted that any of them would require a bandwidth increase of fifty to accomplish the same results. Narrowest band of transmission and least time of transmission are, of course, always obtained by operating the system at the lowest usable signal-to-noise ratio.

A servomechanism may be regarded as a communication system. Its function is to communicate the position of some object, such as a rotatable shaft, to a distant point, and there to cause another object to move in accordance with the motions of the first, The motion of the first object may be, but seldom is, known with absolute accuracy; the motion of the second must always be specified to within certain definite limits. Uncertainties arise in the link between transmitter and receiver; these may be due to backlash, electrical or mechanical noise, instrument imperfections, and the like. The sum of these uncertainties in the transmission link corresponds to the noise discussed in the theory outlined above. The position of the second object, the output member, corresponds to the information required to be transmitted over the system. This information may be considered as a group of integers, each corresponding to a possible output member position. If the static position of the output member is to be specified to 1 per cent, then one of fifty possible integers must be transmitted every time the input member moves through one one-fiftieth of its possible range. It may be that certain elements in the transmission link limit the accuracy of the system (its effective noise-to-signal ratio) to 5 per cent. The conventional thing to do in this case is to transmit the data at a higher rate with multiple-speed data-transmission systems. This is actually a coding scheme, since effectively the single integer required to specify the position of the output member is transmitted as a two-digit integer, where the first digit transmits the rough data and the second the more precise. In this case, therefore, accepted practice coincides with most efficient.

A feature of the scheme of analysis presented in this paper is the breakdown of a continuous smooth curve of data into a series of equispaced points, the value of each point being restricted to one of a number of integral values. Since this technique of analysis is exactly that used in the solution of problems by digital computing machines, one might expect to find correlation between the problems found in this field and those discussed above. To some degree this is true at first glance, and a more thorough study would perhaps prove fruitful. For example, most new computing machinery uses coding to express numbers in the binary system of units; we have seen here that coding in the binary system of units obtains the maximum signal-to-noise ratio for a given carrier-to-noise ratio, consistent with the amount of frequency spectrum utilized. Therefore, we may say that the accuracy of the machine is least affected by noise and perturbations introduced in any of its various transmission links if the data is transmitted by the binary system. Therefore, this is logically sound.

Some Relations Between Speed of Indication, Bandwidth, and Signal-To-Random-Noise Ratio in Radio Navigation and Direction Finding*

H. BUSIGNIES†, FELLOW, IRE, AND M. DISHAL†, SENIOR MEMBER, IRE

Summary—Rate of phenomenon change and required speeds of indication are quite slow in many navigational and direction-finding systems, particularly those for long ranges. Therefore, the actual total required electrical bandwidths are also quite narrow, probably never greater than 100 cps or so, and in most cases much less than this. Even with complex wave forms, such small total bandwidths are possible if there can be designed a discontinuous-type bandpass filter having a multiplicity of very narrow pass bands occurring at the steady-state Fourier components of the complex signal; i.e., a "comb" filter. One practical method of producing such a discontinuous pass band is described briefly.

In view of the interest in new modulation schemes which give an output signal-to-noise ratio which is better than the input carrier-to-noise ratio, it is pointed out that all such systems have improvement thresholds, and many navigational systems provide satisfactory information at output signal-to-noise ratios lower than these threshold values. When this is the case, single-sideband and double-sideband amplitude modulation produce the most sensitive systems.

When postdetection bandwidth is very much narrower than predetection bandwidth, many navigational systems will perform satisfactorily even though the carrier-to-noise ratio at the input to the final detector is appreciably less than unity. When this is so, the phenomenon of "apparent demodulation" is encountered. Because it is of practical importance, an analysis, which is useful for most engineering purposes, is performed to find the relation between the open-circuit antenna carrier voltage or available power, the output signal-to-noise ratio $(S/N)_v$ required for satisfactory indicator operation, the percentage modulation m of the carrier, the predetection bandwidth Δf_{IF}, and the postdetection bandwidth Δf_v when a linear final detector is used in a double-sideband amplitude-modulation system.

It is shown that the above relation depends markedly on whether the quantity $4(\Delta f_{IF}/\Delta f_v)(N/S)_v^2m^2$ is greater or less than unity, and certain practical cases encountered in system design are investigated to show the dependence of the "required carrier for system operation" on the above quantity.

1. Symbols

C = rms carrier voltage in series with the equivalent-generator resistance R_g, which is seen looking back from the receiver input terminals.

P_a = available power from equivalent generator of output resistance R_g

N_{IF} = rms value of thermal-noise voltage appearing in the predetection bandwidth Δf_{IF} and across the input to the linear detector

N_D = rms value of thermal-noise voltage appearing across the load of the linear detector

N_v = rms value of thermal-noise voltage appearing in the postdetection bandwidth Δf_v

C_{IF} = rms value of carrier voltage appearing in the predetection bandwidth Δf_{IF} and across the input to the linear detector

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

S_D = rms value of signal; i.e., the envelope of the modulated carrier, appearing across the load of the linear detector

Δf_{IF} = predetection band-pass bandwidth

Δf_v = postdetection low-pass bandwidth (Δf_v may also be a band-pass bandwidth if $\Delta f_v \ll \Delta f_{IF}$ and the midfrequency of Δf_v is very much lower than Δf_{IF}

G_{IF} = value of predetection voltage gain

G_D = value of detector voltage gain

G_v = value of postdetection voltage gain

NF = noise figure of receiver

R_g = equivalent-generator output resistance, which is seen looking back from receiver terminals

KT = Boltzman's constant times absolute temperature

m = percentage modulation of the carrier.

Unless otherwise noted, all carrier, noise, and signal currents, voltages, and ratios throughout this paper are rms values.

2. Introduction

That a fundamental relation exists between "required minimum pass bandwidth" and "rate of transmitted information" is well known in the communication field. Also well known is the basic relation between thermal-agitation noise power and the pass bandwidth, and, in a more qualitative way, the relation between the pass bandwidth and effect of impulse-type noise on a received signal.

The implications of these relations will be considered with reference to the specific field of radio direction finding and navigation. These considerations are of particular importance in the design of long-range direction-finding and navigational systems. "Long range" is a relative term; for low-frequency navigational systems, distances of 1,500 miles may be considered, whereas for very- and ultra-high-frequency direction finding, line-of-sight distances in the neighborhood of 150 miles are "long" range.

Briefly, we can say that before final detection the rms thermal-noise voltage is proportional to the square root of bandwidth, and the peak voltage resulting from impulse-type noise is directly proportional to bandwidth. Therefore, the narrowest possible bandwidth should be used to minimize noise, and it is important to know what bandwidth actually is required to pass the transmitted information.

The problem of placing as many channels as possible in a given frequency band is an additional practical reason for using narrow-bandwidth systems.

This paper is written not to recommend specific systems but rather to present a point of view concerning the bandwidth required for the transmission of information concerning direction and distance.

3. Required Speed of Indication

With respect to navigational systems, this fundamental point must be realized: when there is no change in the direction of arrival of a signal or in the distance between receiver and transmitter, information involving direction or distance is being conveyed at zero rate and essentially an infinitely small bandwidth is required to indicate this. This is equivalent to saying that whatever wave form is used in a distance- or direction-indicating system, that wave form is in a steady-state condition when there is no change in relative position in the system.

Only when the direction of arrival of a carrier, or the distance between receiver and transmitter is changing, is directional or distance information being received, and the rate at which the direction or distance is changing determines the total bandwidth required at the receiver. This assumes that no communication or other information is to be received at the same time.

In practice, there are actually two "types" of speeds of indication to be considered; one type concerns the rate at which the phenomenon being measured is changing after the phenomenon has already reached its quasi-steady-state condition. This speed of indication is, of course, fixed entirely by the rate of change in the phenomenon; i.e., no matter how large a bandwidth is used, the speed of indication would not increase over its actual occurrence rate.

The other type of speed of indication is involved when the phenomenon is, in effect, "turned on" or "turned off"; e.g., when a direction-finder receiver is suddenly changed in frequency to take bearings on a "new" transmitter. For this case, of course, increasing the bandwidth will increase the speed at which the indicator gives the bearing.

Considering both types of speed of indication, we list below the approximate length of time in which it is desirable that information be obtained in some of the various types of radio direction finders and aids to navigation:

(A) Aircraft radio compass—Two or three seconds; must follow the changes of course of the aircraft.

(B) Omnidirectional radio range—Two or three seconds; could be much slower as it does not have to change rapidly according to plane course or position, except near the station.

(C) Distance-measuring equipment—Two or three seconds would be satisfactory. The present system requires much longer (say, 30 seconds) to provide protection against noise or interference (strobe technique).

(D) Long-range navigational systems—Two minutes is considered satisfactory for a fix. The classical loran is slow because of the procedure of operation.

(E) Radar—Radar of the rotating-antenna type is definitely limited to the time required to effect a complete turn, 1 to 30 seconds.

4. Required Bandwidth

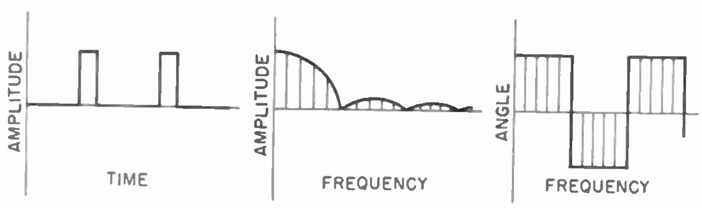
Before considering the specific bandwidths required by the speed of indication noted in Section 3, we would like to review the concept of a discontinuous pass band.

4.1 *Concept of Discontinuous Pass Band*

With respect to required bandwidth, this fundamental point must be realized: no matter how complex the modulation is on a carrier, the steady-state reception of a complex steady-state signal conveys information at zero rate and requires zero bandwidth; i.e., although a multiplicity of infinitely small bandwidth pass bands may be necessary, essentially no resultant bandwidth is required to receive this steady-state complex signal. Only when the envelope of the modulation on the carrier is changing is information conveyed at any rate, and the resultant required bandwidth is fixed by the rate of change of the envelope of the modulation.

As a specific example, consider a steady-state train of modulating pulses; i.e., the time picture before a carrier is modulated.

For the amplitude versus time wave form shown in Fig. 1(a), it is well known that the envelope of the distribution of amplitude with respect to frequency is that shown in Fig. 1(b), and the phase relation between the amplitudes is as shown in Fig. 1(c). Mathematically, the envelope of the frequency distribution shown in Figs. 1(b) and 1(c) is the Fourier transform of the time distribution shown in Fig. 1(a).



(a) (b) (c)

Fig. 1—Well-known frequency distribution for a time distribution consisting of pulses.

When essentially steady-state conditions are reached in the pulse train, the frequency distribution is not continuous, but the amplitudes of Fig. 1(b) occur at discrete points on the frequency scale.[1] (The first zero in the envelope of Fig. 1(b) occurs at a frequency equal to the reciprocal of the pulse width (the other zeros occur at harmonic multiples of this frequency) and the number of discrete frequency lines appearing in each cycle of the envelope is equal to one less than the ratio of the repetition width to the pulse width.)

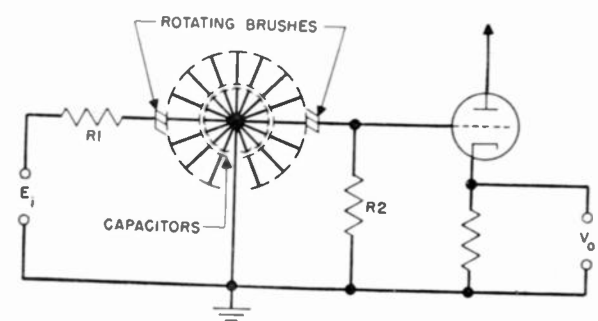
It is, therefore, possible, and in some cases even practical, to reproduce this steady-state pulse train satisfactorily by combining the outputs of a multiplicity of pass bands that are essentially zero cps wide and positioned in frequency so as to pass these separate frequency components. The total resultant bandwidth used can thus be extremely small.

Now, if the wave train shown in Fig. 1(a) should gradually change to a different steady-state amplitude, a certain bandwidth would be required in each of the separate pass bands to permit these multiple pass bands to follow the change in wave form. The bandwidth required depends on the rate of change of the envelope of the modulating wave form in passing from one quasi-steady-state to the next.

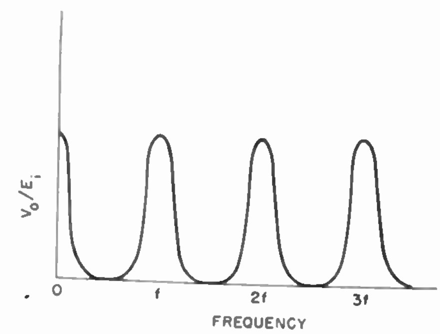
A fairly exact indication of the bandwidth required in each pass band to accommodate a given rate of change may be obtained by solving for the response of a band-pass circuit when a sinusoidal driving voltage is suddenly changed in amplitude; i.e., by finding the "rise time" or the "decay time" of the band-pass circuit. Making use of the LaPlacian transform or any of the other well-known methods of solution, it can be shown[2,3] that the decay (or rise) time in seconds Δt of a band-pass circuit of Δf is approximately $\Delta t \doteq 1/\Delta f$. It is then approximately true that a band-pass circuit will be able to transmit an envelope if the rate of change of that envelope is much slower than the decay or rise time of the circuit.

Each pass band provided for the Fourier series components must have a width equal to that indicated above and, therefore, the total bandwidth needed is directly proportional to the number of Fourier components required to reproduce the wave form satisfactorily.

It is worth repeating that no bandwidth is required to reproduce a steady-state wave form *no matter how com-*



(a)



(b)

Fig. 2—(a) Electromechanical circuit that produces a "comb filter" type of frequency response. (b) The type of frequency response produced. The mechanical rotational frequency f fixes the location of the pass bands, and the time constant $R_1C(R_2 \gg R_1)$ fixes the width of each pass band.

[1] T. E. Shea, "Transmission Networks and Wave Filters," D. Van Nostrand Co., New York, N. Y., 1929; pp. 417–426.

[2] E. A. Guillemin, "Communication Networks," John Wiley and Sons, New York, N. Y.; 1935, vol. II, pp. 474–506.

[3] M. J. DiToro, "Phase and amplitude distortion in linear networks," Proc. I.R.E., vol. 36, pp. 24–36; January, 1948.

plex: bandwidth is required only to follow the rate of change of the envelope of a wave form.

Considering the quite slow rates of change in the phenomenon to be measured and the speeds of indication noted in Section 3, most of the navigational aids require total pass bands that are between 1 and 100 cps wide (for a single-function system measuring only distance or direction) and in many cases are even narrower than this.

4.2 *"Comb Filter" Circuit*

Fig. 2 shows a circuit[4] that produces the discontinuous type of pass band described in Section 4.1. The rotational frequency of the brushes determines the frequency location of the pass bands and the time constant R_1C establishes the width of the pass bands ($R_2 \gg R_1$).

When it is impracticable to utilize such a comb filter and continuous pass bands must be used for navigational systems, the slow rate of phenomenon change should indicate the use of simple wave forms having quite long periods.

5. Some Relations Between Bandwidth and Thermal Signal-to-Noise Ratios

5.1 *Concerning the "Best" Type of Modulation*

Discussions have appeared recently of the concept that the output signal-to-noise ratio of an information-transmitting system can be increased by increasing the receiver bandwidth and using the proper kind of modulation.[5] Consequently, some readers may feel that the concepts of this paper apply only to one type of modulation; i.e., amplitude modulation. This limitation is not correct, as the following brief discussion should indicate.

The fact must be realized that in a great many, perhaps even in a majority of, cases satisfactory navigational information can be obtained at quite low output signal-to-noise ratios $(S/N)_v$; i.e., approximately 2:1, say.

Now in all modulation schemes where the output signal-to-noise ratio from the final detector is better than the input signal-to-noise ratio there is always a predetection carrier-to-noise ratio $(C/N)_{IF}$ *improvement threshold*, which must be passed before the improvement is obtained. Of course, the wider one makes the predetection bandwidth to obtain a larger improvement factor, the greater will be the number of microvolts required to reach this improvement threshold.

The input required to reach the improvement threshold in wide-band systems is always larger than that needed to produce a 2:1 output signal-to-noise ratio in a single-sideband amplitude-modulation system, which uses the narrowest possible band.

Thus, when satisfactory results can be obtained with output signal-to-noise ratios approximately equal to the predetection carrier-to-noise ratio improvement threshold for the various wide-band systems, the most sensitive system (i.e., the one that will operate satisfactorily with the smallest transmitted power) will employ single-sideband amplitude modulation; double-sideband amplitude modulation will be next best.

5.2 *Relations Between Received Carrier, Output Signal-to-Noise Ratio, and Pre- and Postdetection Bandwidths for Double-Sideband Amplitude Modulation with a Linear Final Detector*

When a linear detector is used in a given double-sideband amplitude-modulation system (and for a number of practical reasons this is usually the case), the output signal-to-noise ratio will be found to be directly proportional to the strength of the applied modulated carrier only down to a certain carrier level; below this value, the output signal-to-noise ratio approaches proportionality to the square of the applied modulated-carrier voltage.

This unfortunate effect is due to the phenomenon of "apparent demodulation," which occurs when the carrier-to-noise ratio at the input to a linear detector is in the neighborhood of unity or is less than unity. Ragazzini[6] has considered this phenomenon both analytically and experimentally for the case where the carrier-to-noise ratio at the input to the linear detector does not drop much below unity.

In the previous section, it was noted that satisfactory navigational information can be obtained at output signal-to-noise ratios as low as 2:1. Therefore, when the predetection bandwidth is much wider than the postdetection bandwidth (as is often the case), an output signal-to-noise ratio of 2:1 means that the predetection carrier-to-noise ratio can be appreciably less than 1:1. It thus seemed worth while to extend Ragazzini's work by investigating the phenomenon of apparent demodulation for the case of predetection carrier-to-noise ratios materially less than 1:1. Since it is of practical importance, we will consider the relations between carrier level C, percentage modulation m, predetection bandwidth Δf_{IF}, postdetection bandwidth Δf_v, and output signal-to-noise ratio $(S/N)_v$ in the postdetection bandwidth. This relation will be derived with satisfactory accuracy for most engineering applications.

Since the complete, exact analysis, for *any* predetection carrier-to-noise ratio, of the case where noise plus a modulated carrier is applied to a linear detector leads to quite formidable mathematical manipulations, we will obtain our desired engineering relations by a combination of analysis and experiment.

We will first follow the random noise through the system of Fig. 3 to obtain the resulting noise N_v in the final postdetection bandwidth Δf_v (for the sake of brevity, we will refer to this as the video-frequency bandwidth); the effect of the signal on the noise will, of

[4] Developed by G. R. Clark, of Federal Telecommunication Laboratories, Nutley, N. J.

[5] D. G. F., "Bandwidth vs. noise in communication systems," *Electronics*, vol. 21, pp. 72–75; January, 1948.

[6] J. R. Ragazzini, "Effect of fluctuation voltages on the linear detector," Proc. I.R.E., vol. 30, pp. 277–288; June, 1942.

course, be considered. Next, we will follow the signal through the system to find the resulting signal S_v in the final video-frequency bandwidth Δf_v; the effect of the noise on the signal will, of course, be considered. And, finally, our desired output signal-to-noise ratio $(S/N)_v$ will be given by the ratio of the above signal to the above noise.

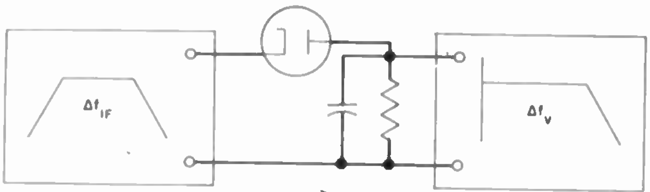


Fig. 3—The three parts of the receiver in which the relation between signal and noise are to be considered.

It should be noted that, to simplify reasoning, the response of the diode load is made at least equal to or greater than the full if bandwidth (i.e., not half that bandwidth). When video-frequency narrowing is obtained after the diode load, it is isolated from that load so as not to impair the frequency response of the diode load. This is normal good design practice insofar as negative-peak clipping, sideband cutting, and similar factors are concerned.

5.2.1 *Resulting Noise in Postdetection Bandwidth:* First we must consider the relation between the noise at the input to the detector and the resulting noise across the diode load.

Landon[7] has shown that, with no carrier present and neglecting diode efficiency or gain, the ac noise voltage output N_D from a perfect linear detector is equal to 0.655 of the noise oscillations N_{IF} entering the diode.

Bennett[8] has shown that, as carrier is added to the noise, the ac noise voltage across the diode load increases by a maximum of approximately 1.6 times as the carrier-to-noise ratio becomes very large. Actually, at a predetection carrier-to-noise ratio of about 3:1, 90 per cent of this increase has already occurred. Stated in another way, when a strong carrier plus noise is fed to the input of a linear detector, the ac noise voltage across the diode load N_D equals the input noise voltage N_{IF}, i.e., 1.6×0.655. Modulation on the received carrier produces an additional small percentage increase in the ac noise across the diode load.

In view of the above quite small (for most purposes) variation in noise across the diode load (i.e., from 0.655 N_{IF} to 1.0 N_{IF}) we will make the engineering approximation that

$$N_D \doteq 0.7 N_{IF} G_D. \tag{1}$$

It is well known that the noise in the if pass band at the input to the diode detector can be expressed in terms of the if bandwidth and noise figure NF of the receiver in the following manner.

$$N_{IF} = G_{IF}(4R_g KT\Delta f_{IF} NF)^{1/2} \tag{2}$$

where R_g is the equivalent-generator output resistance seen looking back from the receiver input terminals, and G_{IF} is the voltage gain of the receiver with reference to the open-circuit voltage of the equivalent generator; i.e., G_{IF} is the ratio of the rms voltage across the output terminals of the last if transformer to the rms open-circuit voltage of the equivalent generator.

Equation (2) can now be written as

$$N_D = 0.7(4R_g KT\Delta f_{IF} NF)^{1/2} G_{IF} G_D. \tag{3}$$

We must next consider the relation between this noise across the diode load and the resulting noise in the final video-frequency bandwidth.

This relation is complicated by the fact that, at input carrier-to-noise ratios below unity, the frequency spectrum of the noise across the diode load is approximately triangular, falling linearly to zero at a video frequency equal to the full if bandwidth; whereas, at input carrier-to-noise ratios above unity, the frequency spectrum of the noise across the diode load is approximately rectangular, cutting off at a video frequency equal to one-half the if bandwidth.[9] By the following approximations, we will obtain an expression for the desired relation that is satisfactory for the great majority of engineering applications.

First: For the case of carrier-to-noise ratios below unity, we will assume that the frequency spectrum of the noise across the diode load is essentially triangular. Figs. 4(a) and 4(b) then give the relation to be considered. The noise power across the diode load is given

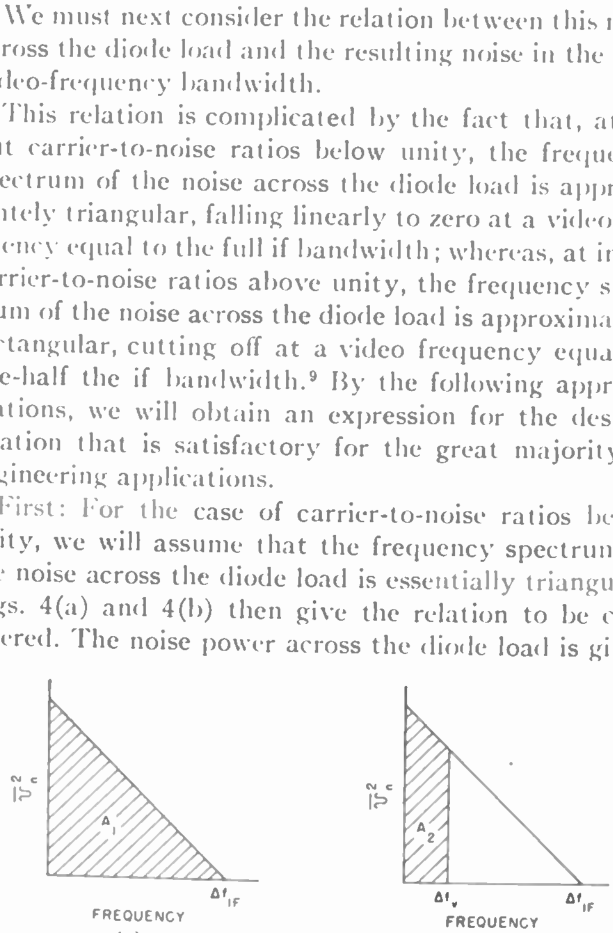


Fig. 4—The relation between the noise power across the detector load and the noise power in the postdetection bandwidth at very low predetection carrier-to-noise ratios.

by A_1 and the noise power in the video-frequency bandwidth is given by A_2. The square root of the ratio of these two values is our desired relation, and from simple geometric considerations can be written

$$\frac{N_v}{N_D} = \left(\frac{\Delta f_v}{\Delta f_{IF}}\right)^{1/2}\left(2 - \frac{\Delta f_v}{\Delta f_{IF}}\right)^{1/2} G_v. \tag{4}$$

It should be realized, of course, that the maximum meaningful value of $(\Delta f_v/\Delta f_{IF})$ in (4) is unity, because

[7] V. D. Landon, "Distribution of amplitude with time in fluctuation noise," PROC. I.R.E., vol. 29, pp. 50–55; February, 1941. Discussion, vol. 30, pp. 425–429; September, 1942.

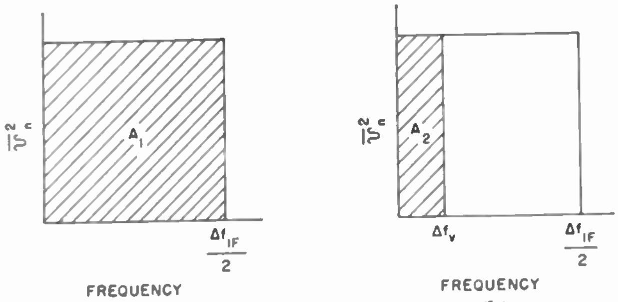
[8] W. R. Bennett, "Response of a linear rectifier to signal and noise," *Jour. Acous. Soc. Amer.*, vol. 15, pp. 164–170; January, 1944.

[9] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. Jour.*, vol. 24, p. 148; January, 1945.

after the video-frequency bandwidth becomes wider than the if bandwidth there is no change in the video-frequency noise output. It is evident that the second factor of (4) varies from a maximum of 1.4 to a minimum of 1.0. For most engineering purposes, this is not a large variation and, because in many practical cases the video-frequency bandwidth is much smaller than the if bandwidth, we will use (4a) as an approximation for (4).

$$\frac{N_v}{N_D} \doteq 1.4\left(\frac{\Delta f_v}{\Delta f_{IF}}\right)^{1/2} G_v. \tag{4a}$$

Second: Now let us consider the case of predetection carrier-to-noise ratios above unity. Here the frequency spectrum of the noise across the diode load is essentially rectangular and Fig. 5 gives the relation to be considered.



(a) (b)

Fig. 5—The relation between noise power across the detector load and noise power in the postdetection bandwidth at high predetection carrier-to-noise ratios.

As in Fig. 4, the desired ratio of N_v/N_D is given by the square root of the ratio of the areas A_1 and A_2 of Figs. 5(a) and 5(b). By simple geometry, the ratio is

$$\frac{N_v}{N_D} = 1.4\left(\frac{\Delta f_v}{\Delta f_{IF}}\right)^{1/2} G_v. \tag{5}$$

We see that (5) for the case of predetection carrier-to-noise ratios above unity is the same as our approximation (4a) for ratios below unity. Thus, for most engineering purposes we can use (4a) or (5) to give the desired relation between the noise in the video-frequency pass band and the noise across the diode load.

Combining (3) and (5), we have our final desired noise equation:

$$N_v \doteq (4R_gKTNF)^{1/2}\Delta f_v^{1/2} G_{IF}G_DG_v. \tag{6}$$

5.2.2 *Resulting Sine-Wave Signal in the Postdetection Bandwidth:* As previously mentioned, Ragazzini[6] has used an analytical approach to this problem that is valid (because of the use of a convergent series) down to predetection carrier-to-noise ratios of approximately unity. Ragazzini also considers the additional relatively small effects of modulation compression and modulation distortion, which we will neglect.

Fig. 6 is an experimentally obtained graph giving the desired relation between the resulting sine-wave signal across the diode load and the input modulated-carrier level as the carrier is varied above and below the value that gives unity carrier-to-noise ratio at the input to the second detector. (The apparatus used to obtain Fig. 6 is described in Section 8.) It will be noted that the abscissa and ordinates are expressed in terms of carrier level $C_{1:1}$, which makes the carrier equal to the noise.

The circles on the graph are the experimentally obtained points, and the solid-line curve is a plot of (7):

$$S_v = G_{IF}G_D \frac{mC}{\left[1 + \left(\frac{N_{IF}}{C_{IF}}\right)^2\right]^{1/2}}. \tag{7}$$

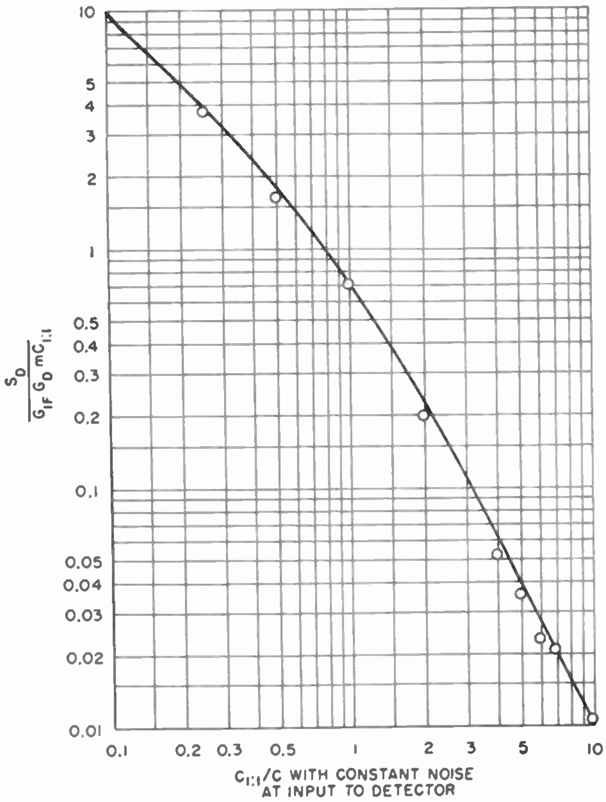


Fig. 6—The relation, for a linear detector, between the resulting signal that appears across the detector load, and the input carrier-to-noise ratio. $G_{IF}G_DmC_{1:1}$ is the signal actually fed to the input to the detector at a 1:1 carrier-to-noise ratio. At unity on abscissa scale, the rms values of carrier and noise are equal.

(Of course, both sides of (7) were divided by $G_{IF}G_DmC_{1:1}$ for plotting Fig. 6.) For most purposes, (7) is an excellent representation of the experimental data, and we will, therefore, consider that (7) adequately describes the phenomenon of apparent demodulation. (Equation (7) was obtained by a guess based on the shape of the experimental curve of Fig. 6 and on (18) of footnote reference 6).

Finally, since we assume that the video-frequency pass bandwidth is wide enough to accommodate the desired signal, we have the fact that

$$S_v = G_vS_D \tag{8}$$

and combining (7) and (8) and making use of (2), we obtain

$$S_v = \frac{mC}{\left[1 + \frac{4R_gKT\Delta f_{IF}NF}{C^2}\right]^{1/2}} G_{IF}G_DG_v. \tag{9}$$

5.2.3 *Value of Modulated Carrier Required to Produce a Given Signal-to-Noise Ratio in the Postdetection Bandwidth:* Dividing (9) by (6), we obtain

$$\left(\frac{S}{N}\right)_v \doteq \frac{mC}{(4R_g KT\Delta f_v NF)^{1/2}} \cdot \left[\frac{1}{1+\dfrac{4R_g KT\Delta f_{IF} NF}{C^2}}\right]^{1/2}. \quad (10)$$

For any given value of modulated carrier and pre- and postdetection bandwidths, we can calculate the approximate signal-to-noise ratio in the final pass band from (10).

Solving (10) for the required carrier value C, we obtain

$$C \doteq (4R_g KT\Delta f_v NF)^{1/2}\left(\frac{1}{m}\right)\left(\frac{S}{N}\right)_v \cdot \left\{\frac{1}{2}+\frac{1}{2}\left[1+4\left(\frac{\Delta f_{IF}}{\Delta f_v}\right)\left(\frac{N}{S}\right)_v^2 m^2\right]^{1/2}\right\}^{1/2} \quad (11)$$

or, in terms of available power from the equivalent-signal generator, we have

$$P_a = \frac{C^2}{4R_g} = KT\Delta f_v NF\left(\frac{1}{m}\right)^2\left(\frac{S}{N}\right)_v^2 \cdot \left\{\frac{1}{2}+\frac{1}{2}\left[1+4\left(\frac{\Delta f_{IF}}{\Delta f_v}\right)\left(\frac{N}{S}\right)_v^2 m^2\right]^{1/2}\right\}. \quad (11a)$$

Equations (11) and (11a) are the important equations of this section of the paper; for any combination of variables in the equation, we can obtain the required carrier to make the system operate.

There are two conditions worth considering wherein (11) reduces to quite simple useful forms. These conditions are set by the quantity

$$4\left(\frac{\Delta f_{IF}}{\Delta f_v}\right)\left(\frac{N}{S}\right)_v^2 m^2,$$

being either much greater or much less than unity.

6. Practical Applications

6.1 *Required Carrier When $4(\Delta f_{IF}/\Delta f_v)(N/S)_v^2 m^2 \gg 1$*

We will first give one example of a practical system in which the condition $4(\Delta f_{IF}/\Delta f_v)(N/S)_v^2 m^2 \gg 1$ is satisfied. A long-range navigational system is to be designed. The indicator reading time can be as long as 10 seconds; therefore, the effective postdetection bandwidth should be of the order of 0.1 cps. Because of transmitter and local-oscillator frequency stabilities, the predetection bandwidth will have to be of the order of 20 cps. Therefore, $\Delta f_{IF}/\Delta f_v$ will always be very much larger than unity. The indicator to be used will give satisfactory readings at a video-frequency signal-to-noise ratio of approximately 2:1. The percentage modulation on the carrier in the final if circuit will be as close as possible to 100 per cent. In this system, $4(\Delta f_{IF}/\Delta f_v)(N/S)_v^2 m^2$ is much larger than unity. Probably the most important single condition is the fact that, because of practical limitations, the predetection bandwidth must be very much larger than the postdetection bandwidth. This particular condition is true of a number of navigational systems.

When $4(\Delta f_{IF}/\Delta f_v)(N/S)_v^2 m^2 > 1$, (11) reduces to

$$C \doteq (4R_g KTNF)^{1/2}(\Delta f_{IF}\Delta f_v)^{1/4}\left(\frac{S}{N}\right)_v^{1/2}\left(\frac{1}{m}\right)^{1/2} \quad (12)$$

or, in terms of available power,

$$P_a = \frac{C^2}{4R_g} = KTNF(\Delta f_{IF}\Delta f_v)^{1/2}\left(\frac{S}{N}\right)_v \frac{1}{m}. \quad (12a)$$

Because these conditions (for which (12) and (12a) are true) apply to many practical systems, these equations are quite useful insofar as design considerations are concerned. As an example, we note that the required carrier for system operation is proportional to the *fourth root* of the if bandwidth. In the light of this result, let us again consider the example given at the beginning of this section. Suppose the frequency-stability problem mentioned there would be materially helped by doubling the if bandwidth (from 20 to 40 cps); we see from (12) that the required carrier for system operation would be increased by only 1.19 times. In view of this small loss in system sensitivity, it might be well worth while increasing the if bandwidth.

6.2 *Required Carrier When $4(\Delta f_{IF}/\Delta f_v)(N/S)_v^2 m^2 \lessapprox 1$*

This condition is usually satisfied by those navigational systems where it is not necessary to make the predetection bandwidth much wider than approximately twice the required postdetection bandwidth (as should be done in double-sideband amplitude-modulation systems). In practice, this usually means that the system requires a rather large postdetection bandwidth. A fast-reading-time automatic direction finder using a cathode-ray-tube indicator is an example of this type of system.

For this case, (11) and (11a) reduce to (13) and (13a).

$$C = (4R_g KTNF)^{1/2}\Delta f_v^{1/2}\left(\frac{S}{N}\right)_v \frac{1}{m} \quad (13)$$

$$P_a = KTNF\Delta f_v\left(\frac{S}{N}\right)_v^2\left(\frac{1}{m}\right)^2, \quad (13a)$$

When the navigational system satisfies the conditions of this subsection, we see that the required carrier for system operation is proportional to the square root of the video bandwidth, and any possible reduction in the video bandwidth will increase the sensitivity of the system in accordance with (13).

6.3 *Adapting a Wide-Band Receiver Design to a Narrow-Band Navigational System*

The title of this section describes a situation which does arise in practice. In a specific case, for example, it was requested that a proposed uhf receiver having an if pass band of 100 kc and a video-frequency pass band of 50 kc be used in a direction-finding system of a type which required a postdetection bandwidth of only 1 kc. Four practical questions arise.

First: What will be the maximum range of the direction finder if this wide-band receiver is used; i.e., what will be the required value of the open-circuit carrier voltage C_1 supplied by the antenna system?

Second: By what factor would we improve the range of the direction-finding system if we redesigned the *post*detection circuits of the receiver so that instead of a 50-kc bandwidth the minimum allowable postdetection bandwidth of 1 kc was obtained; i.e., what will be the required value of open-circuit carrier voltage C_2?

Third: By what factor would we improve the range of the direction-finding system if we redesigned the *pre*detection circuits of the receiver so that instead of a 100-kc bandwidth the minimum allowable predetection bandwidth of 2 kc was obtained; i.e., what will be the required value of open-circuit carrier voltage C_3?

Fourth: How much more improvement will be obtained if the *pre*detection redesign of the third question is carried out as compared to the *post*detection redesign of the second question; i.e., what is the ratio of the required carrier C_2 and C_3?

The important characteristic of the above questions is that a *large* bandwidth reduction ratio will result in either case.

First: Equation (11) answers the first question, and since satisfactory bearing information can be obtained at a 2:1 signal-to-noise ratio in the video-frequency bandwidth, the approximation of Section 6.2, and (13) applies, so that we can write for the required carrier value C_1:

$$C_1 = 0.7\Delta f_{IF_1}{}^{1/2}\left(\frac{S}{N}\right)_v \frac{1}{m}(4R_oKTNF)^{1/2}. \qquad (14)$$

Second: For the required carrier value in the second question, (11) applies again, and since the approximation of Section 6.1 and (12) applies in this second question, we can write for the new required value of the carrier C_2:

$$C_2 = (\Delta f_{IF_1}\Delta f_{v2})^{1/4}\left(\frac{S}{N}\right)_v^{1/2}\left(\frac{1}{m}\right)^{1/2}(4R_oKTNF)^{1/2}. \qquad (15)$$

The ratio of the two carrier values C_2 and C_1 will answer question two.

$$\frac{C_1}{C_2} = \frac{0.7}{m^{1/2}}\left(\frac{\Delta f_{IF_1}}{\Delta f_{v2}}\right)^{1/4}\left(\frac{S}{N}\right)_v^{1/2}. \qquad (16)$$

Note the important fact that, for the conditions of this practical example, a redesign of the postdetection circuits makes the ratio of the two required carriers proportional to the *fourth root* of the ratio of the original intermediate- to new video-frequency bandwidths; i.e., even if a large amount of bandwidth reduction is accomplished by redesigning the postdetection circuits, we will obtain only a rather small improvement in range. However, it should be realized that, in many cases, even the small resulting improvement may be ample payment for the redesign effort required.

Third: with reference to the third question, (11) in its approximation form of (13) applies; i.e., the conditions of Section 6.2 will effectively be satisfied, and so (17) gives the carrier C_3 that will be required for system operation if we redesign the predetection circuits of the receiver.

$$C_3 = \frac{0.7}{m}\Delta f_{IF2}{}^{1/2}\left(\frac{S}{N}\right)_v(4R_oKTNF)^{1/2}. \qquad (17)$$

The ratio of the two carrier values C_3 and C_1 will answer question three.

$$\frac{C_1}{C_3} = \left(\frac{\Delta f_{IF1}}{\Delta f_{IF2}}\right)^{1/2}. \qquad (18)$$

We see that, if we redesign the predetection circuits of the wide-band receiver, the ratio of the two carriers C_1 and C_3 required for system operation will be equal to the square root of the ratio of the original to the if bandwidths.

Fourth: The answer to the fourth question is given by the ratio of C_2 and C_3.

$$\frac{C_2}{C_3} = \frac{\left(\frac{\Delta f_{IF1}}{\Delta f_{IF2}}\right)^{1/2} 1.4(m)^{1/2}}{\left(\frac{\Delta f_{IF1}}{\Delta f_{v2}}\right)^{1/4}\left(\frac{S}{N}\right)_v^{1/2}}. \qquad (19)$$

Examination of (19) shows that if, in the contemplated redesign, the new postdetection and predetection bandwidths will be about the same width (much narrower than the original bandwidths), then markedly more improvement in range will be obtained if we redesign the predetection circuits of the receiver rather than the postdetection circuits.

It must be realized that the redesign of the predetection circuits of a receiver is usually a problem of a much greater magnitude than that of redesigning the postdetection circuits. Thus, it is necessary for the designer to weigh this greater complexity against the increased range obtained by a predetection redesign.

To illustrate the above points, Fig. 7 shows three photographs of a typical pointer-type indication obtained on the cathode-ray-tube indicator of a rotating-loop (or equivalent) direction finder in the 300-Mc region.

Fig. 7(a) was obtained with a uhf receiver having a predetector bandwidth of 100 kc and a postdetector bandwidth of 50 kc. The carrier level C_1 was set so that the output signal-to-noise ratio in the video-frequency circuits is approximately 3:1. The carrier was modulated 100 per cent downward by the direction-finding system.

To obtain Fig. 7(b), the postdetection circuits of this receiver were redesigned so as to have a low-pass bandwidth of 1 kc and, as indicated, it was then possible to use a new lower carrier level C_2 approximately $\frac{1}{3}$ the original value C_1 and still obtain a video-frequency signal-to-noise ratio approximately the same as that in Fig. 7(a). In view of the fact that the 1-kc video-frequency pass band of Fig. 7(b) had a very gradual cutoff, (12) seems to describe satisfactorily these experimental results for most engineering purposes.

To obtain Fig. 7(c), the predetection circuits of the receiver were redesigned so as to have a bandpass of 2 kc. It was then possible to use a new lower carrier level C_3 approximately $\frac{1}{7}$ the original value C_1 and still obtain a video-frequency signal-to-noise ratio approximately the same as in Fig. 7(a). It will be noted that (13) is in satisfactory agreement with the experimental results.

Thus, in this redesign, bandwidth narrowing before final detection produced a system that was more than twice as sensitive as that obtained when the same amount of bandwidth narrowing was accomplished after final detection.

In a practical application of the above reasoning, the range of a uhf direction finder was increased from 30 miles to more than 100 miles by decreasing the if bandwidth until it was as small as could be used.

In many cases, it is possible to design indicators that make use of seemingly nonlinear elements (e.g., rectifiers), and in these cases, the linear concept of a pass bandwidth may not seem to apply rigorously. However, if the *reading time* of the indicator is determined under good postdetection signal-to-noise conditions, then for most engineering purposes the postdetection bandwidth, if of the band-pass type, is quite accurately equal to the inverse of the reading time or to the inverse of twice the bandwidth if of the low-pass type. *Reading time* may be defined in the following manner: A constant bearing with good output signal-to-noise conditions is suddenly applied to the system; the reading time is the time that elapses between the instant when the resulting indication has changed from 5 to 95 per cent of the total amount by which it will change.

In many automatic direction finders, a receiving antenna pattern is rotated. It is possible to use this known rotational frequency to produce a number of different types of *synchronous indicators*. These synchronous indicators have the common characteristic that a known reference frequency can be used so that the *average* produced indication is proportional to only that component of receiver output that is of exactly the same frequency as the reference frequency.[10] Note the importance of the word *average* in the above sentence. An important part of a synchronous indicator is a narrow-bandwidth circuit (either band-pass or low-pass, depending on the type of synchronous indicator

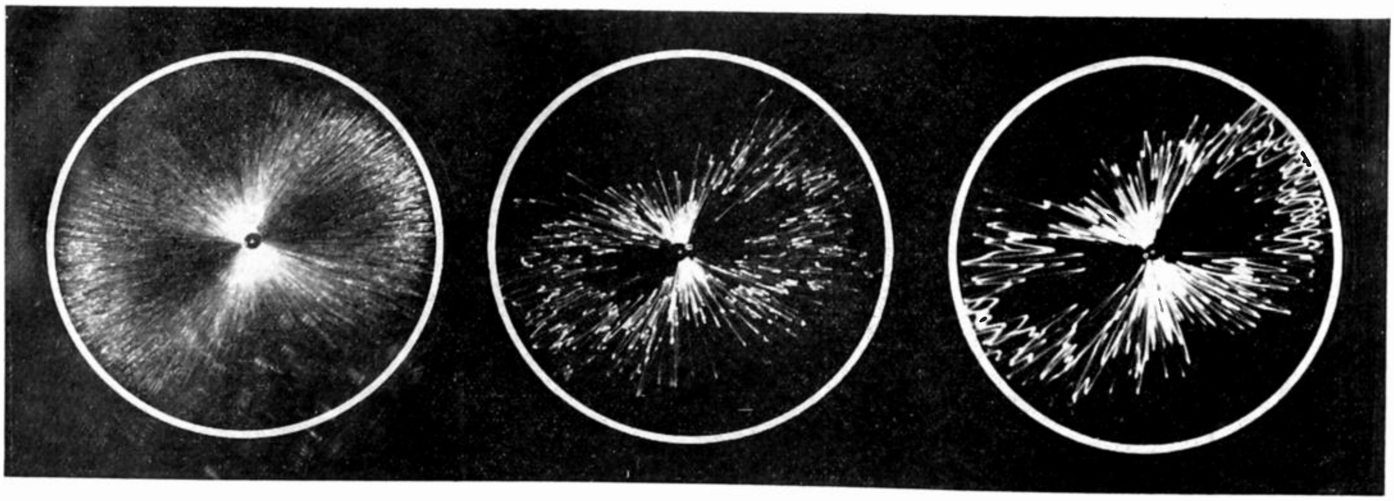


Fig. 7—Pointer-type pattern on a cathode-ray-tube indicator of a rotating-figure-eight direction finder at approximately 3:1 output signal-to-noise ratio. (a) Wide-band receiver, $\Delta f_{IF} = 100$ kc and $\Delta f_v = 50$ kc. (Required carrier was C_1.) (b) Postdetection circuits of receiver redesigned so that $\Delta f_v = 1$ kc. (Required carrier was $C_2 \doteq C_1/3$.) (c) Predetection circuits of receiver redesigned so that $\Delta f_{IF} = 2$ kc. (Required carrier was $C_3 \doteq C_1/7$.)

7. Postdetection Bandwidth

It is important to realize that the postdetection bandwidth Δf_v, continually referred to in this paper, means the bandwidth between the detector and the final indication device. (In some high-speed indicators, the bandwidth response of the human eye may set the postdetection bandwidth.)

[10] Insofar as operation at low predetection carrier-to-noise ratios is concerned, this exact knowledge of the law of effective rotation of the antenna pattern seems to be one of the fundamental advantages that direction finders of the "receiving type" have over direction finders of the "transmitting type."

In the transmitting case (e.g., an omnirange system), it is always necessary to send to the receiver "synchronizing information," and in many cases, the inability to detect this synchronizing information, rather than inability to detect the directional information, sets the limit on the operating range of the system.

used) whose function is to remove the instantaneous fluctuations from the bearing indication, leaving only the average indication. Since many of these synchronous indicators make use of nonlinear elements, the measurement or analysis of the *reading time* of the indicator as described previously is perhaps the best way to find the effective postdetection bandwidth.

The quantity S/N_v is an important term in the equations of this paper. It should be realized that it is possible to express this ratio in the postdetection bandwidth Δf_v in terms of the magnitude of fluctuation that will be seen about the true or average bearing. If the noise produces an error-function type of bearing fluctuation (which is true in many practical cases), then it is practical to express numerically the bearing fluctuation in terms of the statistical quantity, the standard deviation of the bearing from the average bearing.

8. Appendix: Experimental Determination of the Relation Between the Postdetection Signal-to-Noise Ratio and the Predetection Carrier-to-Noise Ratio When a Linear Detector is Used With Double-Sideband AM

A block diagram of the electrical apparatus is given in Fig. 8.

8.1 *Signal Generator and Modulation*

The signal generator and attenuator were high-quality standard laboratory instruments. Modulation at 80 per cent was produced by an alternator driven by a 15-cps synchronous motor, which was run from the power lines. The bandwidth of the postdetection band-pass filter requires that the 15-cps modulating frequency have a short-time frequency stability better than approximately 0.02 cps. By using the power lines to obtain the modulating frequency, the necessary short-time frequency-stability requirement was satisfied.

8.2 *Noise Generator*

The noise generator consisted of a high-gain double-superheterodyne receiver with a local oscillator and mixer added to its final if amplifier so that the resulting thermal noise was produced at the desired mid-frequency of 1.5 Mc. By using the multiple superheterodyne receiver, it was possible to obtain large thermal noise output with no observable regeneration caused by undesired feedback.

8.3 *Final IF Amplifier*

The final if amplifier from which the linear detector operated, and which set the predetection bandwidth of 50 kc, used five cascaded critical-shape-coupled double-tuned circuits and showed no regeneration effects whatsoever. Of course, the signal generator and noise generator, which operate into this amplifier, were carefully decoupled from each other so that any adjustments on one generator had no effect on the output of the other.

8.4 *Linear Detector*

The linear detector consisted of two sections in parallel of a 6AL5 double diode. The frequency response of the diode load used was approximately 150 kc, so that the total triangular noise spectrum was fully reproduced. The ac-to-dc impedance ratio of the total diode load for the modulation frequency of 15 cps was unity (for all practical purposes), so that the 80 per cent modulation used could be handled with no observable distortion. The power-output capabilities of the last if stage driving the linear detector was such that 50 volts peak-to-peak of detected signal could be obtained from the 80 per cent modulated carrier before appreciable distortion could be seen. In the experiment, the diode was operated at an essentially constant output level of 5 volts dc resulting from carrier plus noise. This output voltage was sufficiently below the overload point and sufficiently above the contact-potential voltage so that errors due to both effects were negligible. With reference to the linearity of the detector, no attempt was made to measure the departure from a constant value of the slope of the curve giving output dc versus input carrier; it is standard practice, however, to consider a diode detector to be satisfactorily linear above an output of approximately 2 direct volts, the upper limit being set by the stage driving the detector. Since this experiment dealt mainly with predetector signal-to-noise ratios below unity, it was possible to keep the operating level of the detector essentially constant; i.e., signal plus noise produced approximately 5 direct volts output from the detector.

8.5 *Postdetector Narrow-Band Filter*

Since an expected signal amplitude variation of at least 100:1 was to occur across the diode load, the attenuator shown in Fig. 8 preceding the narrow-band postdetection filter was used so that this filter and the following rectifier voltmeter could be operated at a

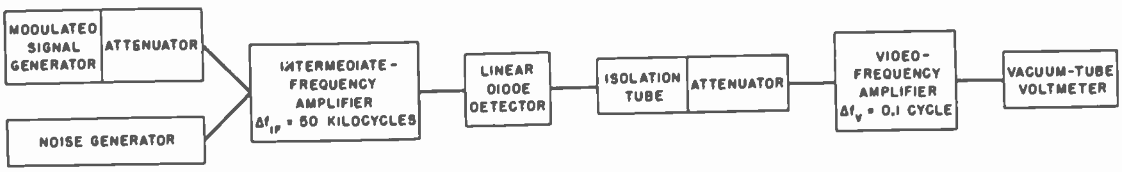


Fig. 8—Block diagram of the apparatus used for experimentally determining the law of "apparent demodulation" versus input carrier-to-noise ratio that occurs in a linear diode at input carrier-to-noise ratios down to 1:10.

constant level; thus the problem of the amplitude linearity of the vacuum tubes in the filter did not have to be considered. Since the circuits following this attenuator then operate at a low level, it is necessary to be sure that hum and af noise have negligible effect.

The required bandwidth for the postdetection filter was fixed by the predetection bandwidth of 50 kc, by the fact that it was desired to obtain readings at predetection carrier-to-noise ratios as low as 1:10, and by the fact that to read the output meter accurately it was necessary to have a reasonably good signal-to-noise ratio, say 5:1, in the postdetection bandwidth at this 1:10 input carrier-to-noise ratio. To calculate this required postdetection bandwidth, we combine (1) and (4a) and obtain

$$N_v \doteq \left(\frac{\Delta f_v}{\Delta f_{IF}}\right)^{1/2} N_{IF} G_D G_v. \tag{20}$$

Next, combine (7) and (8) to obtain

$$S_v = \frac{mC_{IF}}{\left[1 + \left(\frac{N_{IF}}{C_{IF}}\right)^2\right]^{1/2}} G_D G_v. \tag{21}$$

The ratio of (20) and (21) is

$$\left(\frac{S}{N}\right)_v = \left(\frac{\Delta f_{IF}}{\Delta f_v}\right)^{1/2} \frac{m\left(\frac{C}{N}\right)_{IF}}{\left[1 + \left(\frac{N}{C}\right)_{IF}^2\right]^{1/2}} \tag{22}$$

and, finally, for the case we are interested in where $(N_{IF}/C_{IF}) > 1$, we obtain

$$\frac{\Delta f_{IF}}{\Delta f_v} \doteq \frac{\left(\frac{S}{N}\right)_v^2}{m^2\left(\frac{C}{N}\right)_{IF}^4}. \tag{23}$$

For the set of conditions previously mentioned, we see that the ratio of predetection to postdetection bandwidth must be 2.5×10^5; thus we require a 0.2-cps wide postdetection bandwidth centered at the modulation frequency of 15 cps. In the actual experiment, a 3-db bandwidth of approximately 0.1 cps was used. A rough check of the signal-to-noise ratio in this postdetection bandwidth at a predetection carrier-to-noise ratio of 1:10 showed that (23) gives satisfactory engineering accuracy.

This bandwidth was obtained by cascading two band-pass amplifier stages. Each stage consisted of a feedback-amplifier chain using the well-known parallel-tee null network in the feedback path. Each feedback-amplifier chain had a gain of approximately 350 at 15 cps, thus giving a resonant-circuit Q of approximately 85. Each chain was dc coupled throughout, so that there was no low frequency at which the feedback could become regenerative; and by the simple expedient of dropping the high-frequency response of *one* of the stages in the chain, the magnitude of the $\mu\beta$ gain was made less than unity at the high frequency at which the phase shift about the feedback chain was 360°. Since there is 100 per cent negative feedback for direct current in each chain, dc drift troubles due to the dc coupling are not encountered.

It is important that the parallel-tee null network be so accurately aligned that its loss at the null frequency is greater than the gain of the feedback amplifier in which it is used, and it is necessary to devise a rigorous alignment procedure. It is also important to realize that the frequency-selective characteristics of this type of effective band-pass circuit are limited by the bandwidth of the feedback amplifier and by the signal-handling capabilities of the tubes in the chain. It will be remembered that, in the experiment, the postdetection circuits must reject noise-frequency components up to 50 kc. Rather than design the feedback amplifiers for this pass band, they were designed to handle a few thousand cycles and a few sections of *RC* low-pass filters preceded the feedback filter so that no frequency components above a few thousand cycles ever reached the feedback band-pass circuits.

8.6 *Setting for 1:1 Predetection Carrier-to-Noise Ratio*

Landon, in his discussion with Norton,[7] showed that, with only noise present, the average output A of a perfect linear detector (i.e., the dc value read on a long-time-constant dc voltmeter) is $A_N = 1.252\ N$, where N is the rms value of the if noise oscillations. We know that, when carrier alone is present, the average output A_c of a perfect linear detector is $A_c = 1.414\ C$, where C is the rms value of the if carrier oscillations. Therefore, if we have the ability temporarily to cut off the carrier input and thus read the dc output A_N due to noise only, and then temporarily cut off the noise input and read the dc output A_c due to carrier only, the ratio of the readings will be given by (24).

$$\frac{A_N}{A_c} = 0.886 \frac{N}{C}. \tag{24}$$

Thus, for our desired predetection carrier-to-noise ratio of 1:1, the ratio of the readings was made to be 0.886. This ratio was checked before each reading to make sure the noise generator was supplying a stable and constant output.

This setting was also checked in another way. Bennett[8] has shown that, when carrier is added to the noise at the input to a linear detector, the resulting dc output of the detector is approximately 1.4 times the dc output produced by noise alone. The frequency response of the diode load must be equal to or greater than the full if bandwidth, and there must be enough noise (e.g., 2 volts) to ensure that the diode is in its linear region. This ratio was found to check very satisfactorily with the ratio given by (24).

The graph of Fig. 6 was then obtained by varying the signal-generator attenuator from this 1:1 predetection carrier-to-noise condition and finding the corresponding required variation in the audio-frequency attenuator to maintain a constant output from the postdetection filter.

Calculation of Ground-Wave Field Strength Over a Composite Land and Sea Path*

H. L. KIRKE†, FELLOW, IRE

Summary—**A brief discussion of the problem is given, together with a description of the three proposed methods of solution. The results of a practical experiment are shown, and curves calculated by the three methods are compared with the observed results. It is shown that the BBC method gives field strengths much nearer to the observed values than the P. P. Eckersley method, and that the method proposed by Millington, while somewhat more laborious to use, also gives results which agree well with observed values. The difference between the three methods is small at low frequencies and when the effect of the discontinuity is not large.**

Introduction

The calculation of ground-wave field strengths for a homogeneous path has been given adequate mathematical treatment by a number of workers. As a result, the formulas which have been worked out are, in general, universally accepted.

This is particularly the case for distances up to 100 miles. For greater distances, where the earth's curvature is an important factor, the field strength depends to some extent on atmospheric refraction.

Norton has shown that measurements made in the United States agree well with values calculated using an effective value of 4/3 times the actual radius of the earth. This has some bearing on the calculation of field strength over mixed paths over long distances, as will be shown later.

The calculation of field strength for a nonhomogeneous path has been the subject of discussion for many years, but until the last few years no serious analytical treatment of the problem had been made. A number of empirical methods have been tried, however, with various degrees of success.

P. P. Eckersley proposed a method[1] which was used for some years. This method gave results which were in reasonable agreement with observed values for low frequencies (<600 kc) or where the effect of the discontinuity was not large.

In many cases, the P. P. Eckersley method gave results which were far from being in agreement with measured values. The problem later became of considerable importance, and several other empirical methods were suggested and tried by the BBC Research Department. Of these, only one, proposed by Somerville, who was at the time in charge of the Field Strength Section of the BBC Research Department, gave results which were in better agreement with measured values and was simple in its application.

An experiment was conducted over a land-sea path in the north of Scotland from Burghead to Melvich, in 1936, as a check on the method. This path was chosen as it gave extremes in conductivity: badly conductive land, and sea.

The results showed that the Somerville method was considerably better than the P. P. Eckersley method. The bleak and rugged nature of the country, and the changes in conductivity, made it impossible to obtain a sufficient number of measurements from which a reasonably accurate measure of the effective conductivities of the land path could be made. The results, though interesting, are not considered worthy of publication.

Later, as a result of a controversy on the subject, a further experiment was made on a path from Start Point (South Devon) to Happisburgh (Norfolk) in order to find out which of the two methods was the better. It was realized that one experiment does not necessarily give adequate proof, but it is certainly helpful.

The problem has now been treated analytically by T. L. Eckersley and Millington, and the latter has published a paper[2] in which, in addition to the theoretical analysis, he describes a further empirical method which, while somewhat more complicated to use, has a theoretical basis. This method gives results which are in reasonable agreement with practical observations.

Some aspects of the problem will now be discussed, together with a description of the three methods and some comparisons with observed results.

Discussion of the Problem and Description of the Empirical Methods

When a wave travels over a path of uniform conductivity, its rate of attenuation depends upon the wavelength, the conductivity and dielectric constant of the earth or water, and on the curvature of the earth; and the calculation of the attenuation is a relatively simple matter. When, however, a wave travels over a path which has different conductivities, the matter is more complicated.

A wave exists in the space above the ground as well as at ground level, and the field strength at different heights above ground will depend upon the ground constants and curvature of the earth (diffraction). At some height above the ground, the wave will be unattenuated and unaffected by the ground. The field at ground level and the rate of attenuation at ground level will depend upon the state of the wave above the ground, which will

* Decimal classification: R112.1. Original manuscript received by the Institute, December 22, 1948.

† Research Station, British Broadcasting Corp., Kingswood Warren, Tadworth, Surrey, England.

[1] P. P. Eckersley, "The calculation of the service area of broadcast stations," Proc. I.R.E., vol. 18, pp. 1160–1194; July, 1930.

[2] G. Millington, "Ground wave propagation over an inhomogeneous smooth earth," *Jour. IEE* (London), part III, vol. 96, p. 53; January, 1949.

depend upon what has happened to the wave during its passage from the transmitter.

For example, if the wave has passed over ground of poor conductivity, the ground wave will be considerably attenuated but the space wave will be considerably less attenuated, and if, subsequently, the wave travels over land of good conductivity, a considerable amount of energy is available to "fill in" and make up to some extent for the previous losses. Relatively, more energy will be available than would be the case had the previous transmission been over land of good conductivity, when for the same ground-wave field strength the transmitter power required would have been less. The rate of attenuation over the subsequent ground path of good conductivity may be expected to be less than would have been the case had that conductivity been the same throughout the path and the transmitter power less.

In the case of transmission over a path of good conductivity and then over bad conductivity, the ground-wave attenuation for the first part of the path would be low and the relative energy in the space wave less than had the attenuation been high and the transmitter power increased to produce the same ground-wave field strength.

Since, in this case, there is relatively less energy in the space wave, there is less energy available to make up for the losses in the subsequent land of lower conductivity, and the rate of attenuation will be greater than had the whole path been of the lower conductivity and the transmitter power increased.

It can be said that a wave is aware not only of its ground-wave field, but of its space field the power of the transmitter from which it originated, the distance from that transmitter, the ground conductivity, and the curvature of the earth. Any method used to calculate or estimate field strengths over the path of mixed conductivities must, in effect, take all these factors into account. Millington points out that for any method to be theoretically correct the reciprocity condition must be satisfied. That is to say, if the transmitter and receiver are interchanged, the field strength at the receiver remains the same. It is probably more correct, however, to treat the matter as a four-terminal network and not interchange the antennas.

P. P. Eckersley Method

The first method used by the BBC was proposed by P. P. Eckersley, and assumes that the rate of attenuation of the wave over any homogeneous part of a composite path is that which would occur if the whole path was of that conductivity but for a transmitter of different power. If the first part of the path were sea and the second land, the curve for land all the way is used and the field strength multiplied by a factor which is equal to the ratio of field strengths at the sea-land junction for sea and land for this part of the path. If this ratio is, say, 5, then the field strength over the land path would correspond to that for a land path all the way but for a transmitter of 25 times the power.

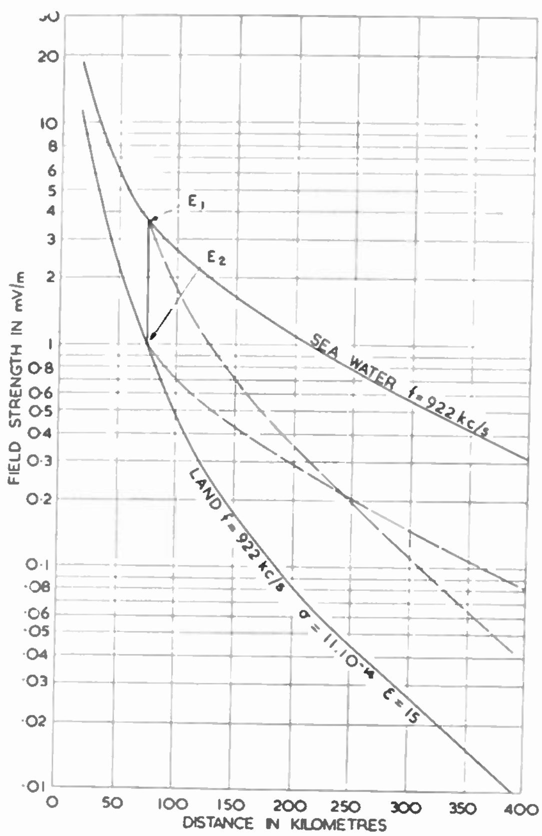


Fig. 1—P. P. Eckersley method, the equivalent-power method.

Fig. 1 illustrates the method. In the case of a sea path followed by a land path, the curve for sea is drawn up to the junction. In this case, 75 km has been chosen as the junction point. To obtain the curve for the subsequent land path, the field strengths for land all the way are multiplied by the ratio of land and sea fields at the junction. For a land path followed by a sea path, the curve for land is drawn up to the junction. The curve for sea is then drawn by dividing values for an all-sea path by the ratio of fields for sea and land at the junction.

A graphical method can be used if, as is usual, the field strength is plotted on a logarithmic scale, or else in decibels relative to a datum level on a linear scale. It is immaterial whether the distance scale is linear or logarithmic. For this method, the curves for land all the way and sea all the way are drawn. With a piece of tracing paper the curve for the appropriate conductivity up to the junction is then drawn, and the tracing paper moved up or down until the other curve coincides at the junction point, and the curve is then continued as for the conductivity of the second part of the path.

There is no justification, however, for the assumption that the rate of attenuation over any part of a mixed path will be the same as would have been the case had the conductivity been the same throughout the whole path unless, as Millington shows, the wave has traveled a sufficient distance from the discontinuity that it can be considered to have "settled down" after the disturbance caused by the discontinuity. The P. P. Eckersley method does not allow for the "filling in" which takes place after the wave has been severely attenuated by land and then travels over sea, or the reverse.

If the wave existed as a very thin layer, confined, say, between the earth and an imaginary boundary some small distance above the earth, the conditions might then be equivalent to transmission along a nonhomogeneous transmission line, and the rate of attenuation in each part of the line would then correspond with the constants of that part of the line, and would be unaffected by the rate of attenuation in the preceding part except as regards reflection.

Many observations have shown that the P. P. Eckersley method described above gives results which are much too high when the first part of the path is sea and the second land, and further (a point which is perhaps not so well realized), too low when the first part of the path is land and the second sea.

Somerville Method

Because of the failure of the P. P. Eckersley method in many cases, a second method, proposed by Somerville, was tried. This method does not directly take into account the factors mentioned above and which were appreciated at the time, but appears by the results it gives to do so indirectly.

In this method, the actual transmitter is replaced by an imaginary transmitter of equal power but at a different distance, the distance being such that the field at the junction of the two paths is the same as that from the actual transmitter, but as if the conductivity of the whole path was that of the second part of the path.

In the case of a sea path followed by a land path, the actual transmitter is replaced by an imaginary transmitter moved nearer to the receiver. In the case of a land path followed by a sea path, the imaginary transmitter is moved further away from the receiver.

This method can be conveniently called the equivalent-distance method, while the P. P. Eckersley method can be called the equivalent-power method.

The Somerville method is found, in practice, to take some account of the effect of "filling in" when the wave has been attenuated over a land path and then travels over sea, and the reverse effect. It gives results which are very near to actual values obtained in practice.

While reciprocal agreement is rarely, if ever, obtained with the P. P. Eckersley method, it is, in many cases, approached very closely in the Somerville method. This is a fortunate coincidence rather than a justification of the method.

The Somerville method can also be used graphically, but it is essential that the distance scale be linear.

Fig. 2 illustrates the method. The curves for land and sea are drawn as in the previous case, and the appropriate curve up to the junction is traced; the tracing paper is then moved to the left or right along the distance scale until the curve for the conductivity appropriate to

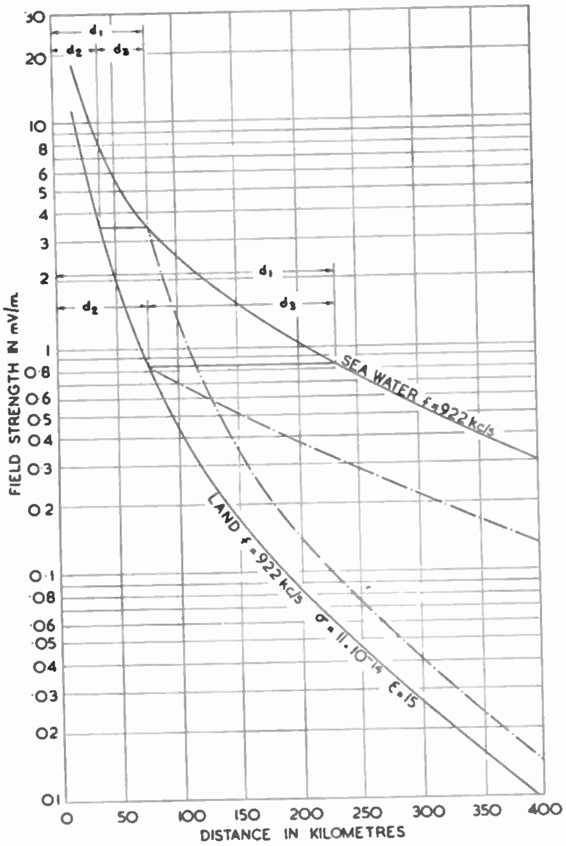


Fig. 2—Somerville method, the equivalent-distance method.

the second part of the path coincides at the junction with the one already traced up to the junction. The tracing is then continued along the new curve.

An equally convenient way to use either method graphically is by the use of a pair of dividers used vertically in the one case and horizontally in the other.

Millington Method

A full description of this method and an analysis of the problem is contained in Millington's paper. Briefly, his method is as follows.

A calculation is first made by the P. P. Eckersley method, and the field strength E_1 noted. The transmitter and receiver are interchanged and a further calculation made, and the field strength E_2 noted. The true field strength is then the geometric mean of E_1 and E_2.

A graphical method can be used for the Millington method. It is an advantage in this method if field strength is plotted in decibels above or below a datum point and on a linear scale. It is not considered necessary to illustrate the method graphically.

Millington points out that an interesting result of using his method is that, where the attenuation over the first part of the path is high, the field strength subsequent to the junction can be greater than that at the junction. This possibility was realized at the time when the Somerville method was evolved, and the Burghead-Melvich experiment showed a small increase.

Millington also suggests that it may be an advantage in some cases to move the transmitter further away from the receiver if, by doing so, the first part of the path is of better conductivity than it otherwise would be.

Description of Experiments

The experiments described below were carried out before Millington's method had been suggested. Comparisons were originally made, therefore, between measured values and those calculated by the P. P. Eckersley and the Somerville method, respectively. Calculations have now been made using Millington's method, and the results are also shown.

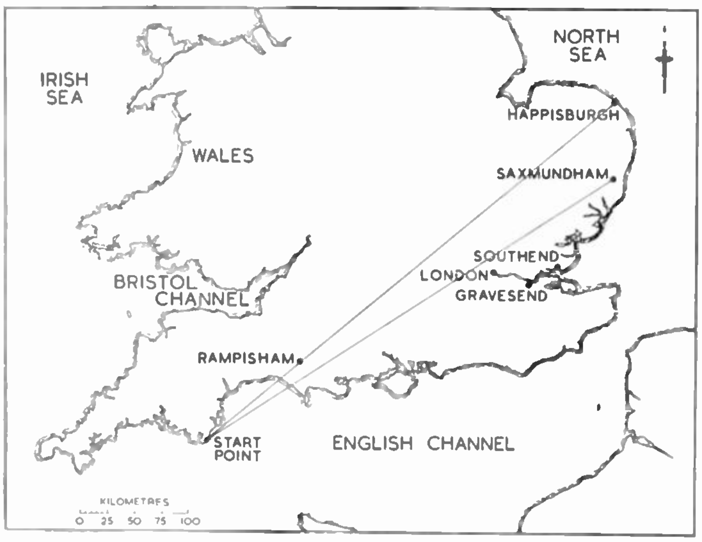


Fig. 3—With a transmitter erected at Start Point, field-strength measurements were made at many points on a line with Happisburgh and Saxmundham.

A low-power transmitter was erected at Start Point, South Devon, and field-strength measurements were made at many points on a line joining Start Point with Happisburgh in Norfolk, starting at the Dorset coast (see Fig. 3). The length of the sea path was 82 km and the total distance was 460 km. The measurements were made at three frequencies, viz., 537, 922, and 1,240 kc.

A second low-power transmitter was also erected at the BBC transmitting station at Rampisham, near Dorchester, Dorset (see Fig. 3), and field-strength measurements made on the same line to Happisburgh and at the same places as the measurements from Start Point.

The effective power radiated was determined in both cases by local field-strength measurements in the well-known and accepted manner.[3] These values of radiated power were used to find a multiplier with which to bring all the field-strength measurements up to an equivalent power of 1 kw for comparison with the standard field-strength curves which are plotted for that power.

The effective conductivity of the ground from Rampisham to Happisburgh was found by comparing the measured values of field strength from the Rampisham transmitter with the standard-field-strength curves.

[3] K. A. Norton, "The propagation of radio waves over the surface of the earth and in the upper atmosphere," Proc. I.R.E., vol. 24, pp. 1367–1388; October, 1936; p. 1371.

For a complete investigation, it is necessary to carry out measurements to determine separately the conductivities of the various portions of the path. It was not possible in the time available to carry out such tests.

It was found by inspection of the results that the best general fit with the measured values from Rampisham was obtained using the FCC (Norton) curves. These curves have, therefore, been used in computing the field strength for the composite path by the three methods.

In the above experiments, as will be shown later, there were some irregularities in the measured field-strength curves at distances over 300 km, and the power of the transmitter was too low for accurate field-strength measurements to be taken at distances much greater than 300 km, particularly at the higher frequencies. Therefore, a second experiment was made.

In this experiment, the high-power transmitter at Start Point was used on its normal frequency of 1,050 kc, using the normal mast radiator. Two field-strength runs were made, one in the direction of Happisburgh on the same line as in the previous experiment, and one in the direction of Saxmundham.

The field-strength measurements given in this paper were made by the BBC Research Department, but some measurements were also made by engineers of the Government Communications Center. Comparisons of the

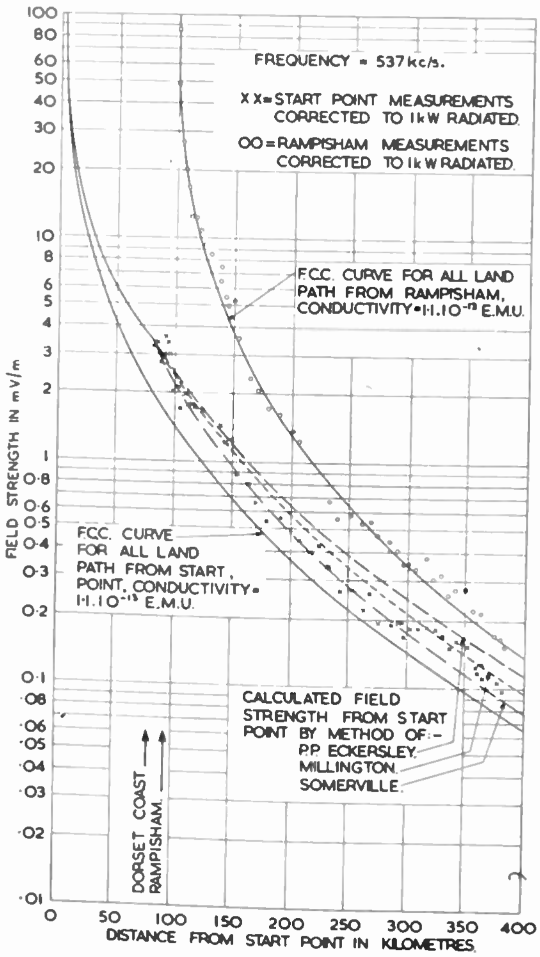


Fig. 4—Results at 537 kc, low-power transmitters at Start Point and Rampisham.

measurements showed good agreement except in cases where the field was very weak.

Results of Experiments and Comparison with the Empirical Methods of Calculation

Experiment 1: Low-Power Transmitters at Start Point and Rampisham

The measured values of field strength from both transmitters are shown in Figs. 4, 5, and 6.

Results at 537 kc: In Fig. 4, it will be seen that the measured values for the Rampisham transmitter agree well with the calculated curve for $\sigma = 11 \times 10^{-14}$, but that the measured values are somewhat higher at distances (from Start Point) between 250 and 380 km and lower for greater distances. This indicates that the conductivity is somewhat better for distances of 250 to 380 km, and less at greater distances. These differences between measured values and those calculated on the assumption of uniform conductivity must be taken into account in comparing the measured and calculated values from the Start Point transmitter.

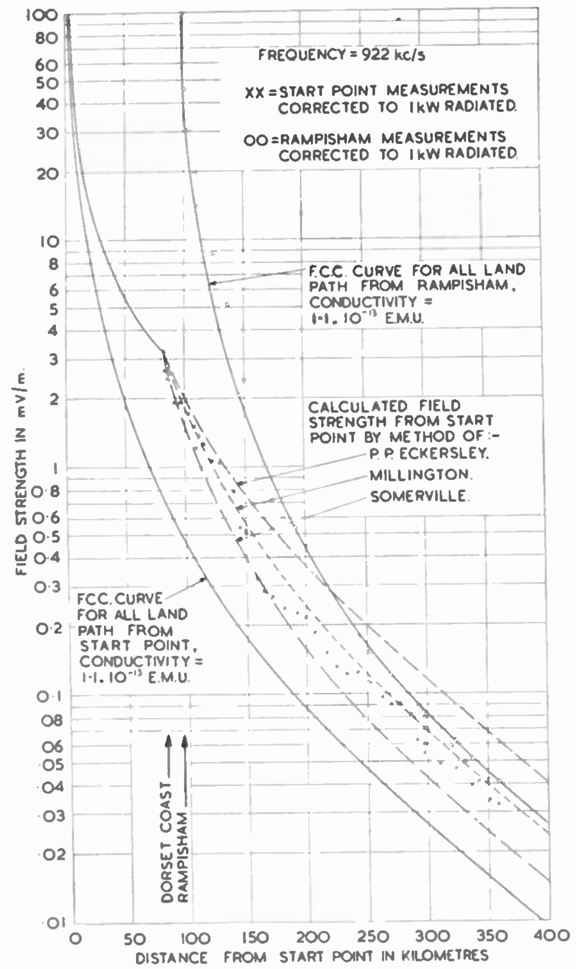


Fig. 5—Results at 922 kc, low-power transmitters at Start Point and Rampisham.

On 537 kc, the Somerville method shows closer agreement with measured values than the Millington method, although the results of the latter are sufficiently close for most practical purposes. The Eckersley method gives results which are always too high. There is not, in any case, a great difference between any of the methods at this frequency and with such values of conductivity.

Results on 922 kc: In Fig. 5, it will be seen that the measured values of field strength from the Rampisham transmitter over the all-land path are also in generally fair agreement with the calculated curve for $\sigma = 11 \times 10^{-14}$. The measured values for distances less than 50 km from the transmitter are, however, higher than the calculated values. The choice of a different value of conductivity in making the calculation would not make any appreciable difference in the calculated values, and it must be presumed that either the measurements are in error or that the corrections for 1 kw radiated are incorrect.

In the results from Start Point, the measured values give the best fit with the Millington curves, and are somewhat higher than the Somerville curve at the greater distances.

Results on 1,240 kc: In Fig. 6, the measured values for the Rampisham transmitter again show reasonable agreement with the calculated curve for $\sigma = 11 \times 10^{-14}$.

The results from Start Point on this frequency are in

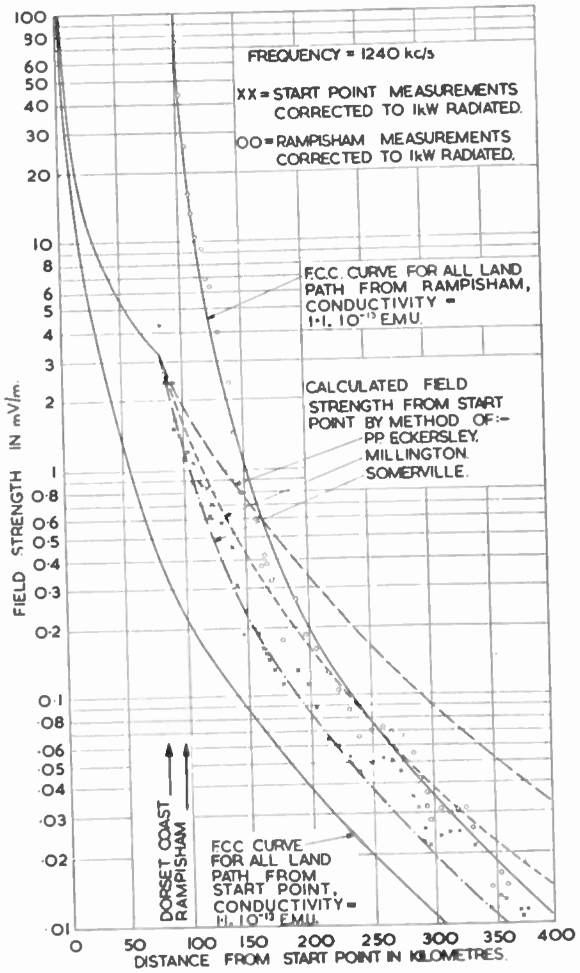


Fig. 6—Results at 1,240 kc, low-power transmitters at Start Point and Rampisham.

closer agreement with the Somerville than the Millington curve. They are somewhat high, having regard to the comparison of calculated and measured values from Rampisham, while the P. P. Eckersley curve gives results which are always higher than the measured values.

It is to be noted that in Figs. 4, 5, and 6 the disagreement between the P. P. Eckersley method and measured values is greater at higher frequencies.

Experiment 2: High-Power Tests from Start Point on 1,050 kc

As stated before, these tests were made at high power in order to obtain higher field strengths at the greater distances and greater accuracy in measurement. They were made on this frequency, which was the normal operating frequency of the station. They were made in two slightly different directions (see Fig. 3) in order to investigate the peculiarity at the greater distances. Figs. 7 and 8 show the results on the runs to Happisburgh and Saxmundham, respectively, and calculations by all three methods are shown for comparison.

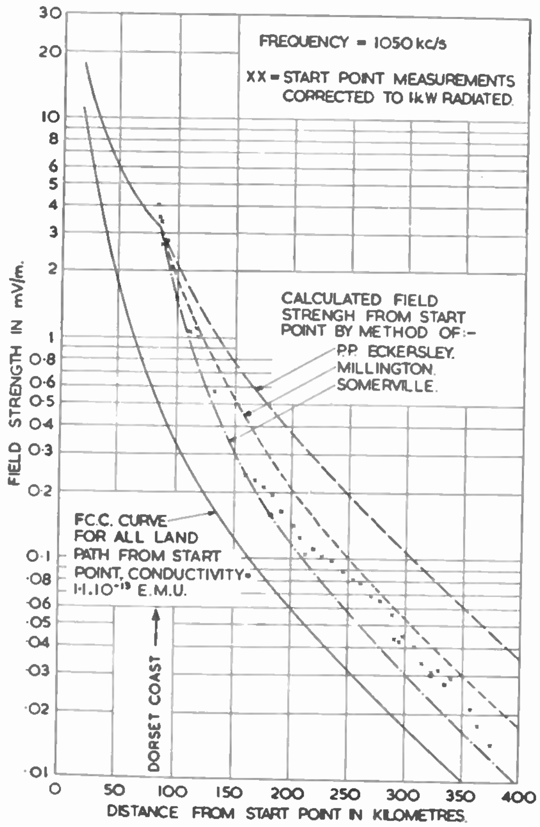


Fig. 7—Results at 1,050 kc, high-power tests from Start Point to Happisburgh.

In both cases, the Somerville method is in closer agreement with measured values than either of the others. As before, the P. P. Eckersley method gives results which are very much higher than observed values (over 3 to 1 at the greater distances).

Measurements in Denmark

Since this paper was commenced, CCIR Paper No. 148E of Study Group No. 2 (Propagation) has become

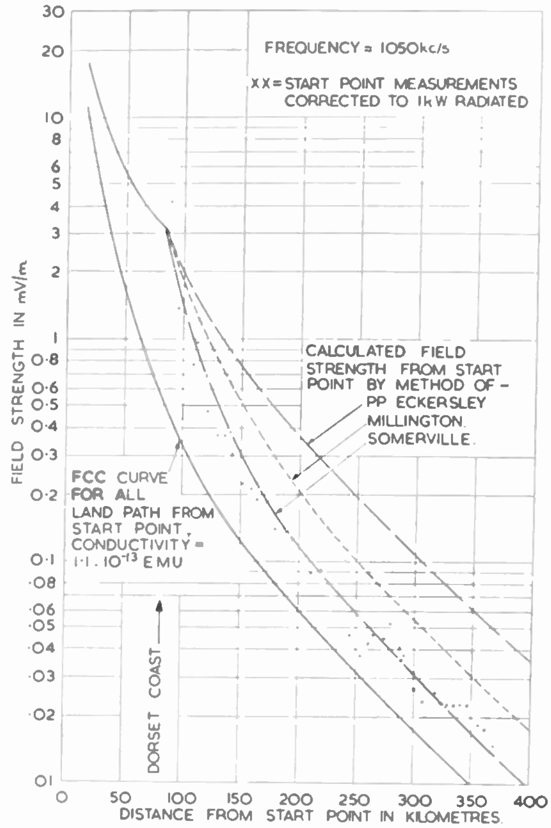


Fig. 8—Results at 1,050 kc, high-power tests from Start Point to Saxmundham.

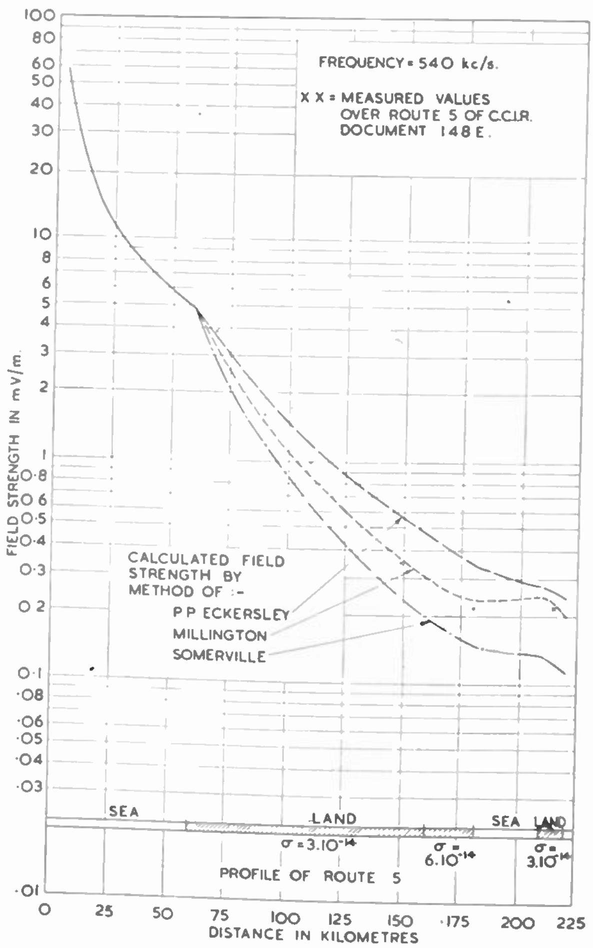


Fig. 9—Results at 540 kc over route 5 of CCIR Paper No. 148E.

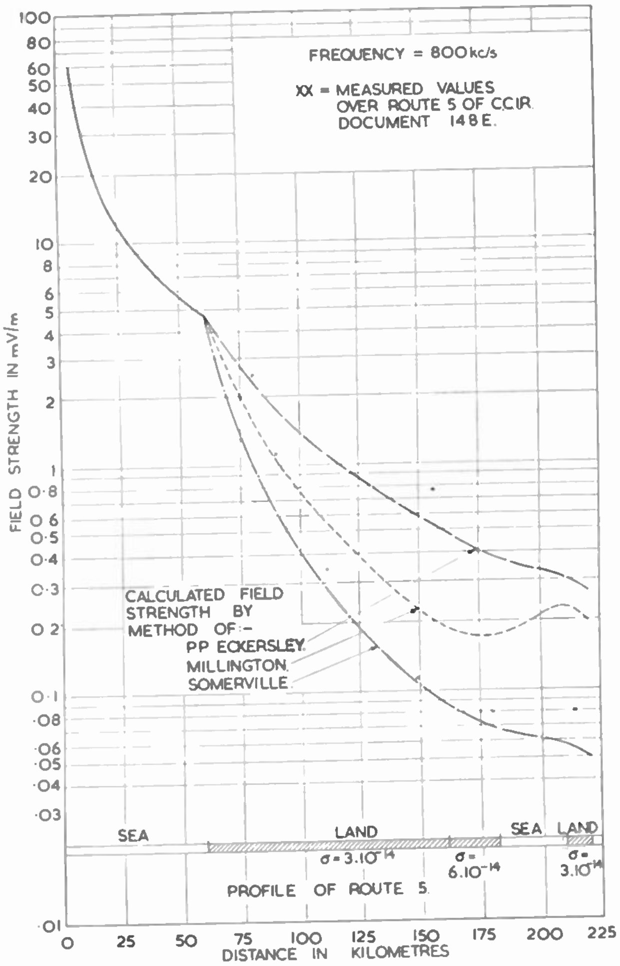


Fig. 10—Results at 800 kc over route 5 of CCIR Paper No. 148E.

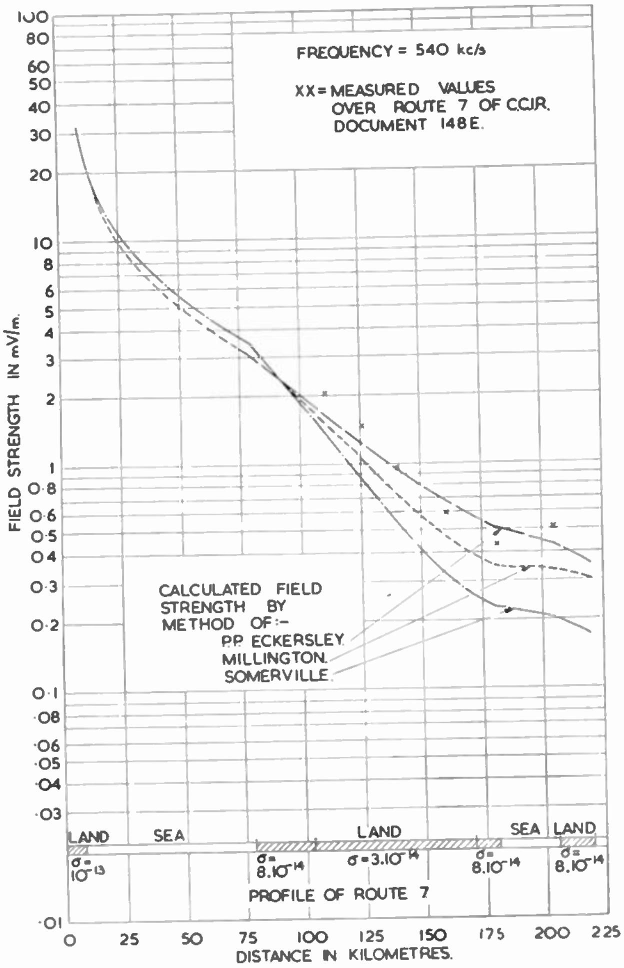


Fig. 11—Results at 540 kc over route 7 of CCIR Paper No. 148E.

available. That paper gives a number of measurements carried out over composite paths in Denmark by engineers of the Department of Posts and Telegraphs. Permission has been obtained from that authority to quote the results in this paper.

Figs. 9 through 13 show the measured values, together with curves calculated by the three methods using the values of conductivity quoted in the CCIR paper.

From Fig. 9 it will be seen that the measured values correspond well with the curve calculated by the Millington method.

In Fig. 10, which is over the same route as Fig. 9 but at a higher frequency, there is better correspondence with the Somerville method at the greater distances, but it is to be noted that this method does not show the rise in field strength at 200 km when the wave travels over a sea path after traveling over poor land. The Millington method does show this rise.

In Fig. 11, over another route, the P. P. Eckersley method shows the best correspondence with the measured values, although the Millington method is in fairly good agreement. It would appear that all the measured values are high. This could be accounted for by an incorrect multiplier to bring the results to 1 kw radiated.

In Fig. 12, on the same route as Fig. 11, the Millington method shows the best correspondence with the measured values, and again shows a rise in field strength at 200 km.

In Fig. 13, on a different route and a different frequency, the comparison of the three methods is interesting. The Millington method shows a rapid drop in field strength when the wave travels over poor land after a passage over sea, and a rise in field strength (fill up) when it again reaches the sea having passed over poor land. It is unfortunate that no measured values are available for the intermediate points between 80 and 230 km, but the agreement between the measured values and those calculated by the Millington method for distances over 230 km is remarkable.

Conclusions

The empirical method proposed by P. P. Eckersley is inadequate when the difference between the land and sea attenuation is large.

The Somerville method gives a closer approximation to the measured values than the Eckersley method. It is very simple to use, and is probably adequate for rough calculations where, in many cases, the conductivity data are of doubtful accuracy.

The Millington method does not give such a good cor-

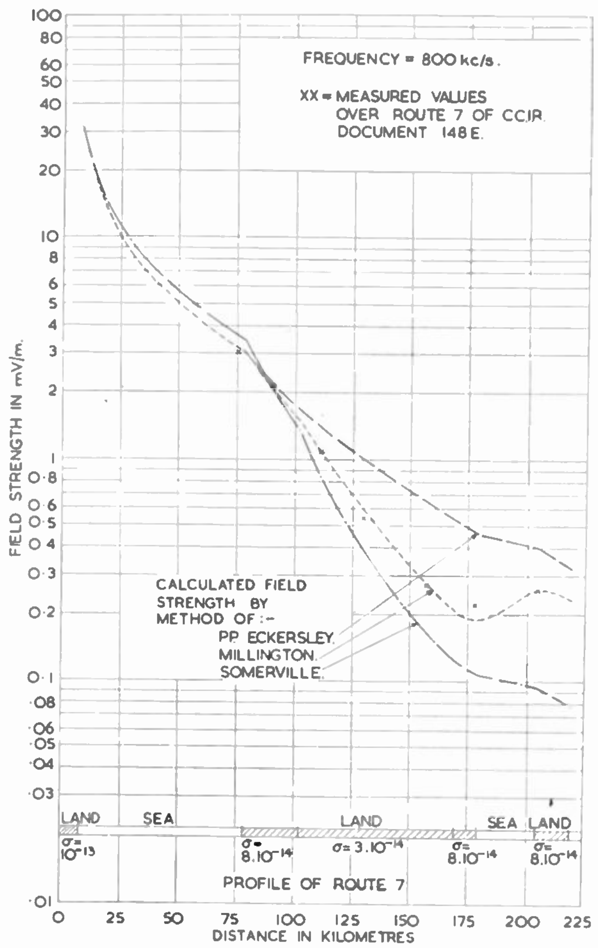


Fig. 12—Results at 800 kc over route 7 of CCIR Paper No. 184E.

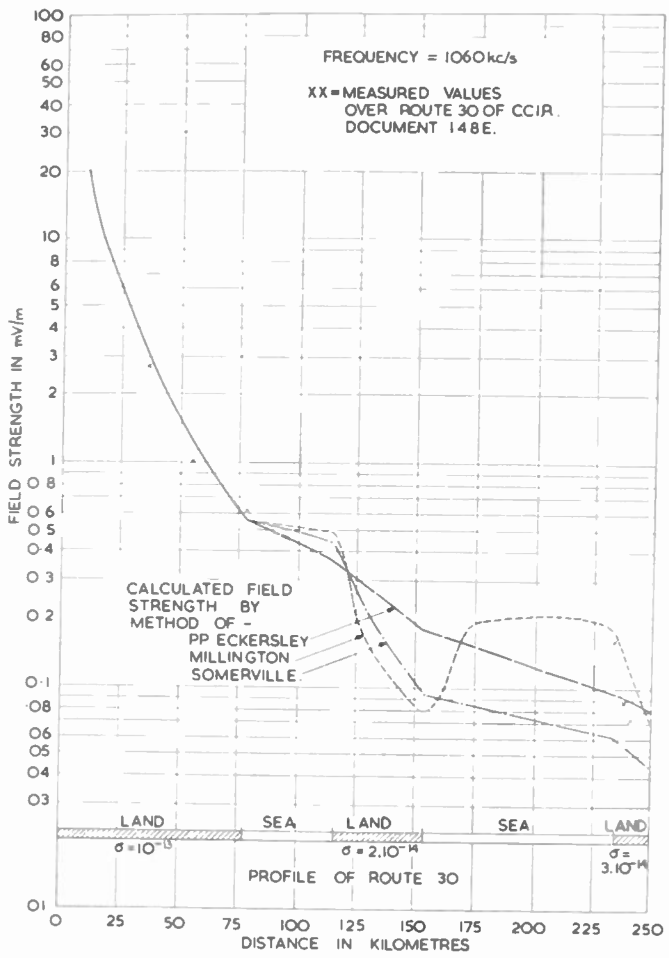


Fig. 13—Results at 1,060 kc over route 30 of CCIR Paper No. 148E.

respondence with measurements in England as the Somerville method. It appears to give better correspondence with some measurements made in Denmark.

There is insufficient evidence to justify the adoption of any empirical method for all conditions. In general, the Somerville method gives sufficiently good results for most practical purposes, and is easier to use than the Millington method. The Millington method has a better justification for its use on theoretical grounds, and in extreme cases shows "filling in" effects which are in agreement with theoretical considerations and practical observations.

It is hoped that this paper will stimulate others to carry out theoretical and experimental work and make further comparisons between measured and calculated values. Further work over the "disturbance region" is particularly necessary. It would be of special interest to carry out tests to investigate the reflection from a discontinuity. Any further work should be in the nature of true research rather than an attempt to justify or disprove any empirical method.

Acknowledgments

The author wishes to thank the British Broadcasting Corporation for permission to publish this paper, the Danish Posts and Telegraphs Department for permission to use the data in CCIR Paper No. 148E of Study Group No. 2 (Propagation), and T. L. Eckersley and G. Millington for their help in innumerable discussions on the subject.

He wishes to acknowledge, with grateful thanks, the past and present work of the Field Strength and Propagation Section of the BBC Research Department for their unfailing help and zeal: in particular, T. Somerville, who was Head of the Section when the work described was carried out, and R. A. Rowden, its present Head. He also wishes to thank S. F. Brownless for his valuable help in preparing the curves.

Automatic Frequency Phase Control of Television Sweep Circuits*

E. L. CLARK†, SENIOR MEMBER, IRE

Summary—**This paper describes three different types of automatic-frequency-control circuits: (1) sawtooth type, (2) sine-wave type, and (3) pulse-time type.**

The sawtooth system forms a sawtooth from the pulse present across the deflection yoke. This sawtooth is compared in phase with the synchronizing pulse to produce a control voltage for frequency phase control of the sweep circuit.

The sine-wave type comprises a stable sine-wave oscillator which is controlled in phase and frequency by the synchronizing pulse, and in turn controls the sweep circuit.

The pulse-time system measures the area of the synchronizing pulse as it rests on the edge of a shelf. Phase variations change this area, and provide information to control the sweep circuit.

A TELEVISION PICTURE can be considered as being composed of horizontal lines of varying brightness. A line is traced by a flying spot starting at the left of the picture and uniformly traversing the raster to the right side of the picture, and then rapidly returning to the start of the next line. In this manner, the lines composing the picture are built downward from the top to the bottom of the picture.

In order that the picture be perfect, the elements of successive scanning lines must coincide exactly. To accomplish this, a trigger signal or synchronizing pulse accompanies every line to initiate retrace. When the circuit used is such that every line must be started by the synchronizing pulse, as in older television receivers, it is termed triggered-type synchronizing. This type of synchronizing is satisfactory if there is sufficient signal and no interference present. However, in practice these requirements are not always met, in which case some of the lines are triggered by noise or interference. This random triggering by the noise results in an imperfect picture.

To improve the picture register and reduce the effects of interference, the automatic-frequency-control system of synchronizing was developed. In essence, this system consists of integrating a number of synchronizing pulses to provide a control, rather than controlling each scanning line individually.

There are three commonly used systems for obtaining automatic frequency control. They are often referred to as (1) the sawtooth type, (2) the sine-wave type, and (3) the pulse-time type.

I. Sawtooth System

A schematic diagram of the sawtooth-type automatic-frequency-control circuit as applied to horizontal deflection is shown in Fig. 1. This diagram includes a synchronizing amplifier tube $V1$, a phase detector or keying-circuit tube $V2$, a dc-amplifier tube $V3$, and a blocking-oscillator tube $V4$. $T1$ is a pulse-type transformer that passes the horizontal synchronizing pulses, but not the vertical synchronizing pulse. The polarity of the pulse in the secondary of $T1$ is such as to cause current to flow in the diodes $P1$, $K1$ and $P2$, $K2$ of tube $V2$. The diode current charges capacitors $C1$ and $C7$ with polarity as shown, which is such as to bias the diodes of tube $V2$ open except when the pulse is present. This diode circuit is in the form of a bridge circuit with a voltage from $P1$ to $K2$, but no voltage from $P2$, $K1$ to ground. In operation, it can be used as a keying circuit to key the voltage existing on the transformer $T1$ centertap (℄) to capacitor $C2$. This keying action is present only when the diodes of tube $V2$ are caused to conduct by the action of the synchronizing pulses applied to them. A sawtooth of voltage as shown at A is applied to the centertap (℄) of transformer $T1$. This sawtooth of voltage is produced by partially integrating a positive pulse (curve B) obtained from the horizontal deflection transformer (not shown). The amplitude of this sawtooth is adjusted so that it is less than the voltage developed across capacitors $C1$ and $C7$ by the synchronizing pulses, and hence will not cause diode conduction by itself. However, during the time the synchronizing pulse is present, the diodes conduct and can be considered as momentarily connecting the sawtooth to capacitor $C2$. With normal adjustment, the synchronizing pulse occurs at point 1 on curve A. Point 1 is on the ac axis of the sawtooth voltage; hence no charge will be acquired by capacitor $C2$. Assume that the phase of the sawtooth voltage changes so that the synchronizing pulse occurs at point 2 on curve A. The voltage at this point is negative with respect to the ac

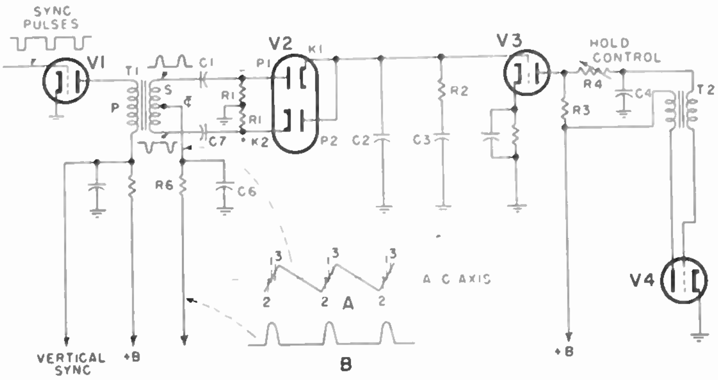


Fig. 1—The schematic diagram of the sawtooth-type automatic-frequency-control circuit, as applied to horizontal deflection.

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† RCA Victor Division, Camden, N. J.

axis and will be keyed to capacitor $C2$, which will gradually acquire a negative charge equivalent to the voltage present at point 2. The grid of tube $V3$, being connected to capacitor $C2$, is also negative, which reduces its plate current, resulting in an increase of voltage on its plate. This increased voltage is connected by way of resistor $R4$ and transformer $T2$ to the grid of the blocking-oscillator tube $V4$. Its effect is to change the phase of the blocking oscillator in such a way that the synchronizing pulse is returned toward its original position 1 on curve A.

If the phase of the sawtooth had changed so the synchronizing pulse occurred at point 3 of curve A, the charge acquired by capacitor $C2$ would be positive and the voltage on the plate of tube $V3$ would be reduced, which would cause the phase of the blocking-oscillator tube $V4$ to change, again returning the position of the synchronizing pulse toward its original position, point 1. Thus, the frequency and phase of the blocking oscillator is held in step by integrating the synchronizing pulses. It does not matter whether the oscillator changes or the rate of the synchronizing pulses change; the action is the same. The grid circuit of tube $V3$ has no dc return, and its potential can change only through the action of the keying-circuit tube $V2$. The time constant of the system is governed by the values of $C2$ and $R2$, $C3$.

This automatic-frequency-control system can also be applied to the vertical sweep circuits.

II. Sine-Wave System

The circuit of the sine-wave type of automatic frequency control is shown in Fig. 2. This circuit embodies a stable sine-wave oscillator of the Hartley type ($V2$), a comparitor ($V1$), and a reactance tube ($V3$).

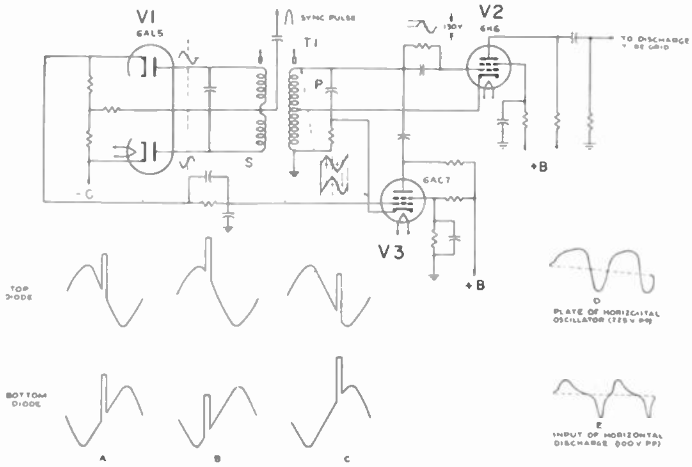

Fig. 2—The schematic diagram of the sine-wave type of automatic-frequency-control circuit, as applied to horizontal deflection.

Its features are automatic operation and good immunity from interference. Tube $V2$ is an extremely stable sine-wave oscillator operating at the horizontal rate of 15,750 cps. In operation, the phase of the sine wave and the synchronizing pulses are compared. A phase change will produce dc information which is applied to the grid of the reactance tube $V3$, which controls the frequency of the Hartley oscillator.

A double-diode tube $V1$ is used as a comparitor, and functions in much the same way as an FM detector. The plates of this tube are connected to the center-tapped coil S of $T1$, which is inductively coupled to P of $T1$, the tank coil of the Hartley oscillator.

Referred to the centertap of S, sine-wave voltages of equal amplitude and opposite polarity are applied to each of the diode plates of tube $V1$. The synchronizing pulses are applied to the centertap, and consequently this voltage appears in the same phase and with equal amplitude on each diode plate.

When the synchronizing pulses and sine wave are properly phased (as shown in Fig. 2, curve A), there will be zero voltage developed at the output of the comparitor.

If the phase of the synchronizing pulse changes with respect to the sine wave (as shown in curve B), then the top diode will produce more voltage output than the bottom diode, resulting in a positive voltage at the comparitor output. In curve C the reverse condition exists, and a negative voltage will appear. Obviously, then, the dc output of the comparitor will run from negative through zero to positive, depending on the phase relation of the pulse and the sine wave. In this way, the necessary control information is produced and applied to the grid of the control tube ($V3$) through a filter network which removes interference pulses and other misinformation, which would otherwise affect the frequency of the oscillator.

The oscillatory action takes place between the screen, grid, and cathode of tube $V2$. The peak-to-peak sine-wave voltage on the oscillator grid is approximately 130 volts. This grid swing produces the wave shape on the plate (as shown in curve D), which is differentiated by an RC network. The wave after differentiation is shown in curve E. The positive portion of this differentiated wave operates the discharge tube. In practice, it is necessary to phase the synchronizing pulse with the differentiated wave used to trigger the discharge tube. To do this, the winding S of transformer $T1$ is tuned off resonance from the oscillator tank circuit P of $T1$. This circuit is tuned on the low-frequency side of the oscillator tank circuit, which operates at 15,750 cps. When properly adjusted, the picture raster will show a small percentage of blanking on the right side of the picture, and will also be blanked to the same extent on the left side. With this adjustment, the picture is properly phased with respect to the synchronizing pulses. If the blanking bar occurs in the middle of the picture, the winding S of $T1$ is badly out of adjustment, or the diode plates of tube $V2$ are interchanged with respect to the winding S of $T1$.

III. Pulse-Time System

The schematic diagram of the pulse-time or width-control type of automatic-frequency-control circuit is shown in Fig. 3. Its operation is based on what may be described as "width modulation" of the synchronizing

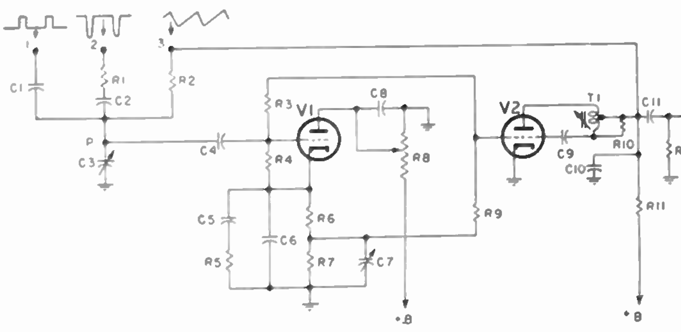

Fig. 3—The schematic diagram of the pulse-time type of automatic-frequency-control circuit, as applied to horizontal deflection.

pulse. A locally generated sawtooth is phased to allow a varying portion of the synchronizing pulse to fall atop the positive corner of the sawtooth, while the remaining portion slides down the steep side. Control voltage is a function of the pulse width atop the positive corner of the sawtooth, the peak amplitude of the combined waves being essentially constant.

Width modulation constitutes the basic difference between the pulse-time and other forms of automatic-frequency-control systems. Satisfactory operation of this circuit depends upon the proper wave shape being formed to apply to the grid of the control tube.

An exploded view of the wave-shaping network and wave shapes as applied to the grid of the control tube is shown in Fig. 4.

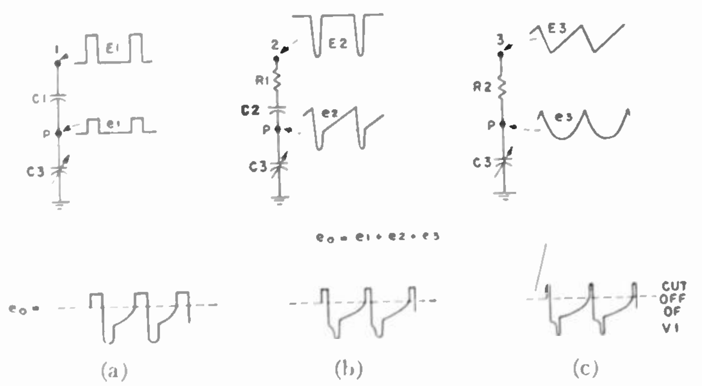

Fig. 4—An exploded view of the wave-shaping network and wave shapes, as applied to the grid of the control tube of the pulse-time type of automatic-frequency-control circuit.

The synchronizing pulse E_1 is fed into the circuit at point 1, and appears as e_1 at point P, being attenuated by capacitors $C1$ and $C3$.

The wave shown as E_2 is obtained from the horizontal deflection system and is in the form of a high-voltage negative pulse, which is fed into the system at point 2. The network $R1$, $C2$, $C3$ partially integrates and attenuates this pulse E_2 to form the wave e_2 at point P. The third wave shape E_3 is obtained from the discharge capacitor connected to the tap on the blocking-oscillator transformer, and is fed into the network at point 3. Resistor $R2$ and capacitor $C3$ attenuate and integrate E_3 to form the parabola e_3 at point P.

A parabola has the advantage over a sawtooth of having a steeper slope near the peak, and therefore provides increased gating. The voltage e_2 is necessary to produce a sharp downward slope immediately following the peak of the parabola e_3.

The combined wave e_0 at point P, which is coupled to the grid of the control tube, is shown for three different conditions of phase between the local oscillator and the synchronizing signal: (1) the curve at (a), when most of the synchronizing pulse is atop the parabola; (2) at (b), when one-half of the synchronizing pulse is atop the parabola; and (3) at (c), when most of the synchronizing pulse is down the slope.

Tube $V1$ (Fig. 3) is the control tube and is biased near cutoff by the dc component of the oscillator grid voltage applied through resistors $R3$ and $R4$. Its plate current consists essentially of pulses whose width is determined by the relative position of the synchronizing pulse atop the peak of the parabola. The voltage developed across resistor $R7$ by this average plate current is injected from the cathode circuit of the control tube $V1$ into the grid of the oscillator tube $V2$ by way of resistor $R9$, and thus maintains the phase of the oscillator with respect to the synchronizing signal within very close limits. The cathode circuit is an integrating network with the following properties: a fast response as $C6$ is relatively small, and a slow response as $C5$ and resistors $R6$ and $R7$ are relatively large. The former integrates the pulses of current and also acts to prevent hunting, while the latter maintains control over a longer period of time, and filters out disturbances of greater duration. The plate circuit contains a potentiometer $R8$, which acts as a vernier speed control.

The capacitor $C3$ is made adjustable, so that the control-tube's grid voltage can be varied to suit the characteristics of the individual tube, and thus maintain the control range at a uniform level.

The blocking-oscillator circuit used in this system is somewhat different from that of conventional blocking oscillators. The transformer $T1$ is an autotransformer arranged in an if-transformer can, and uses a powdered-iron core which permits a certain amount of frequency adjustment. This is limited by coupling requirements, and additional range is obtained by the use of a trimmer capacitor $C7$, which is connected across a portion of the blocking-oscillator's grid resistor $R7$. Tube $V2$ functions not only as the blocking oscillator, but also as the discharge tube. A sawtooth of voltage is developed across capacitor $C10$, and is used for horizontal deflection.

The synchronizing separator used with the pulse-time automatic-frequency-control system must be of a type that provides synchronizing pulses of constant amplitude. This system has been called a width-modulation system; actually, it can be considered a variable-area system. If the synchronizing pulse amplitude is not maintained constant, the area of the effective portion of the synchronizing pulse will not contain the proper information, and poor results will be obtained.

Conclusions

In comparing these three systems of automatic frequency control, it can be said that they all provide better synchronizing than can be obtained by the use of a triggered synchronizing system.

The sawtooth type of afc (Fig. 1) as shown does not snap into synchronism immediately, but takes an appreciable time. However, this undesirable feature of operation has subsequently been overcome.

The sine-wave type snaps into synchronism the instant the signal is applied. It is also the most noise-immune of the three systems. It requires more "B" current than the others and uses the most tubes.

The pulse-time system uses the fewest tubes of the three systems. The picture snaps into synchronism as soon as the signal is applied. The noise immunity is equally as good as the sawtooth type, and approaches the sine-wave type very closely.

Superregeneration—An Analysis of the Linear Mode*

HERBERT A. GLUCKSMAN†, ASSOCIATE, IRE

Summary—**Although superregeneration is usually discussed as a nonlinear problem, the linear mode is an example of a linear circuit problem with a time-varying parameter. The present analysis determines the effect on the behavior of a tuned circuit when its damping factor is subjected to sinusoidal variation. The amplitude and frequency of this variation are considered the fundamental parameters which distinguish the superregenerator from an ordinary resonant circuit. Sensitivity and selectivity are studied as functions of these parameters. It is shown that the solution of the differential equation predicts the phenomenon of multiple resonance and other well-known properties of a superregenerator in the linear mode.**

I. A Linear Circuit Problem

A superregenerator will be here defined as a tuned circuit, as in Fig. 1, in which the damping factor $u \equiv G/2C$ varies periodically with time. The superregenerator will be said to be

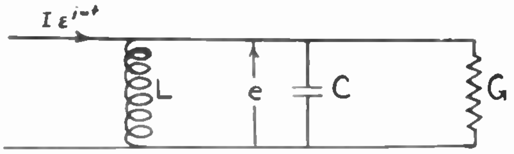

Fig. 1—Equivalent superregenerator circuit.

operated in the linear mode when u is a function of time alone, and not of the voltage e between the nodes of the circuit. The circuit parameters used in the analysis will be the damping factor or decrement u, the natural angular frequency $\omega_0 \equiv 1/\sqrt{LC}$, and the capacitance C. Except for u, these will be considered constant. The linear mode, thus defined, implies a linear circuit problem. The signal is represented by the driving current $I\epsilon^{j\omega t}$ entering one node of the circuit and leaving the other, while the response is the voltage e across the nodes.

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† Cambridge Field Station, Air Matériel Command (Formerly Watson Laboratories), Cambridge, Mass.

An example of an actual superregenerator circuit using grid quench, is shown schematically in Fig. 2. By

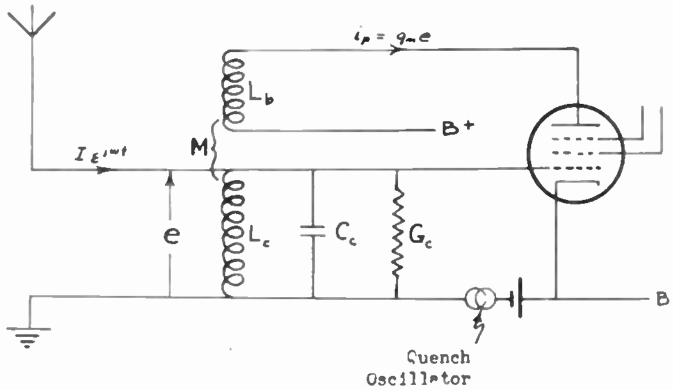

Fig. 2—Tuned-grid superregenerator.
Equations for reduction to Fig. 1:
$L = L_c$
$C = C_c$
$G = G_c - M/L\ g_m$.

the usual pentode approximation, the plate current through the tickler coil is $g_m e$ where g_m is the transconductance. The approximate differential equation of this circuit is, therefore,

$$C_c\dot{e} + G_c e + \frac{1}{L_c}\int\left[e - M\frac{d(g_m e)}{dt}\right]dt = I\epsilon^{j\omega t}$$

where ω is the angular driving frequency.

This reduces to

$$C_c\dot{e} + \left(G_c - \frac{M}{L_c}g_m\right)e + \frac{1}{L_c}\int e\,dt = I\epsilon^{j\omega t}.$$

This is also the equation of Fig. 1, if

$$L = L_c,$$

$$C = C_c \quad \text{and}$$

$$G = G_c - \frac{M}{L}g_m.$$

Other types of superregenerators may also be reduced to Fig. 1. In the above example, the effect of the quench oscillator is to vary g_m, and hence G, or the damping factor u which has already been defined as $G/2C$.

II. The Differential Equation and Its Solution

The equation for Fig. 1, obtained from Kirchhoff's law, is

$$C\dot{e} + Ge + \frac{1}{L}\int e\,dt = I\epsilon^{j\omega t}. \tag{1}$$

Written in terms of the parameters defined above, this becomes

$$\dot{e} + 2ue + \omega_0^2\int e\,dt = \frac{I}{C}\epsilon^{j\omega t}. \tag{2}$$

Differentiation yields

$$\ddot{e} + 2u\dot{e} + (2\dot{u} + \omega_0^2)e = j\omega\frac{I}{C}\epsilon^{j\omega t}. \tag{3}$$

The dissipation term, having the variable coefficient $2u$, may be conveniently eliminated by use of the transformation[1]

$$e = x\epsilon^{-\int_0^t u\,dt}. \tag{4}$$

This reduces (3) to

$$\ddot{x} + (\omega_0^2 - u^2 + \dot{u})x = j\omega\frac{I}{C}\epsilon^{j\omega t}\epsilon^{\int_0^t u\,dt}. \tag{5}$$

Here u^2 and $\dot{u}$ may be considered negligible compared with ω_0^2, since the damping factor u is the half-bandwidth of the circuit (expressed in angular frequency) and $\dot{u}$, for sinusoidal u, is at most the product of the amplitude of u by the angular frequency of its variation. This frequency, generally called the quench frequency, is in practical cases much smaller than ω_0.

Equation (5) may, therefore, be simplified, for practical cases, to

$$\ddot{x} + \omega_0^2 x = j\omega\frac{I}{C}\epsilon^{j\omega t}\epsilon^{\int_0^t u\,dt}. \tag{6}$$

Since u is a periodic time function, it can be expressed as a Fourier series of sines and cosines. When this is integrated term by term to obtain $\int_0^t u\,dt$, another series of sines and cosines results. Since a power of ϵ with a sinusoidal exponent can be reduced by means of Bessel functions to a complex Fourier series of simple exponentials, the righthand member of (6) can be reduced to a product of several such complex Fourier series. Once this is done, the solution of (6) is quite straightforward.

[1] This transformation was suggested by P. David in "Les superréactions," *L'Onde Elect.*, vol. 7, pp. 217–260; June, 1928.

The solution will be given here for the case where u has only one sinusoidal term:

$$u = a - b\sin pt.$$

At this point it is well to fix in mind the roles of the constants a, b, and p. A reference tuned circuit whose parameters are C, ω_0, and a may be conceived as a standard for comparison. The half-bandwidth or damping factor of this circuit is a. The superregenerator differs from the reference circuit only because of the variational damping term $b\sin pt$. The amplitude b and angular frequency p of the variational damping term are, therefore, the distinctive parameters of the superregenerator. These two parameters provide a variational damping plane over which it will be of interest to study the behavior of the superregenerator.

Use of the above expression for u reduces (6) to

$$\ddot{x} + \omega_0^2 x = j\omega\frac{I}{C}\epsilon^{(a+j\omega)t}\epsilon^{b/p\cos pt}\epsilon^{-b/p}. \tag{7}$$

By means of the Fourier expansion

$$\epsilon^{b/p\cos pt} = \sum_{n=-\infty}^{\infty} I_n\left(\frac{b}{p}\right)\epsilon^{jnpt}, \tag{8}$$

in which the I_n are the modified Bessel functions, (7) is reduced to the soluble form

$$\ddot{x} + \omega_0^2 x = j\omega\frac{I}{C}\epsilon^{-b/p}\sum_{n=-\infty}^{\infty} I_n\left(\frac{b}{p}\right)\epsilon^{[a+j(\omega+np)]t}. \tag{9}$$

The solution in x can be used to obtain the solution in e of (3) through transformation (4). If the transient term is omitted, this solution is

$$e = j\omega\frac{I}{C}\epsilon^{j\omega t}\epsilon^{-b/p\cos pt}$$

$$\sum_{n=-\infty}^{\infty} I_n\left(\frac{b}{p}\right)\frac{\epsilon^{jnpt}}{[a+j(\omega-\omega_0+np)][a+j(\omega+\omega_0+np)]}. \tag{10}$$

This can be simplified for practical purposes by noting that the second bracketed factor of the denominator of the general series term is approximately $2j\omega$ for cases not far off resonance, except for terms of higher order, which terms are generally negligible compared with the sum of the lower-order terms. With this approximation and the definitions

$$\left.\begin{aligned} \alpha &\equiv \frac{a}{p} \\ \beta &\equiv \frac{b}{p} \\ \gamma &\equiv \frac{\omega_0-\omega}{p} \\ e &\equiv E\epsilon^{j\omega t} \\ a &\equiv \frac{G_0}{2C} \end{aligned}\right\}, \tag{11}$$

the solution may be written

$$E = \frac{I}{G_0} \epsilon^{-\beta \cos pt} \sum_{n=-\infty}^{\infty} I_n(\beta) \frac{\epsilon^{jnpt}}{1 + j\dfrac{n-\gamma}{\alpha}}. \quad (12)$$

Note that α and β are the average and variational damping factors normalized to the angular quench frequency, that γ is the number of angular quench frequency units by which the signal departs from resonance, that E may be called the complex envelope of the voltage e, and that G_0 is the average value of the conductance G.

The complex envelope E is at all times real only for the case $\gamma=0$, i.e., at resonance. It may be written, in general,

$$E \equiv |E| \epsilon^{j\phi} \quad (13)$$

where both the magnitude $|E|$ and the phase angle ϕ are functions of time.

III. Properties of the Linear Mode Predicted by the Solution

The term "linear mode" was originally chosen to describe operation in which the oscillations do not involve nonlinear portions of the tube characteristics; that is, a condition of saturation is not approached.[2] Under such conditions the circuit damping is independent of the oscillations, in agreement with the definition used above for the linear mode. Experiments show that the principle of superposition holds in the linear mode, further justifying the use of a linear differential equation.

Solution (12) states that the response to a continuous wave at a frequency not too far from resonance is a wave whose envelope is a periodic function the fundamental frequency of which is that of the damping variation, i.e., quench frequency. A factor of the response is a series of terms of the form

$$I_n(\beta) \frac{\epsilon^{jnpt}}{1 + j\dfrac{n-\gamma}{\alpha}}.$$

Since this general term has a maximum magnitude at $\gamma=n$, the over-all response can be expected to have maxima at integral values of γ. This phenomenon, called multiple resonance, is a well-known property of superregeneration. As observed, the maxima decrease in height as γ departs from zero in either direction. This is predicted by the fact that $I_n(\beta)$ decreases as n increases. The observed symmetry about $\gamma=0$ is checked by the fact that $I_n(\beta)=I_{-n}(\beta)$. Multiple resonance is not observed at low quench frequencies. This is also predicted, for, if $1/\alpha \equiv p/a$ is small compared with unity, then the change in the imaginary part of the denominator when γ changes by one integral step is insignificant compared with the real part of the denominator. Figs. 3 and 4

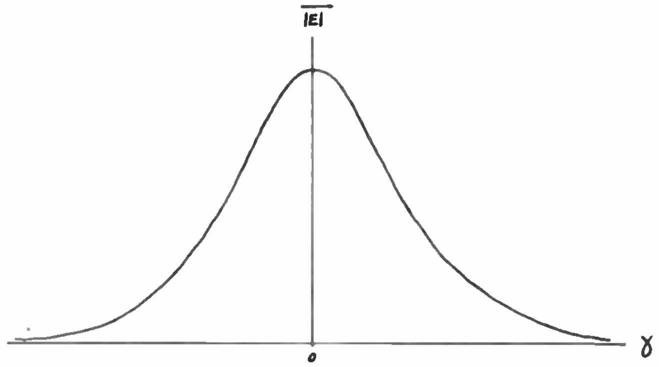

Fig. 3—Frequency response with no observable multiple resonance [$p/a \ll 1$].

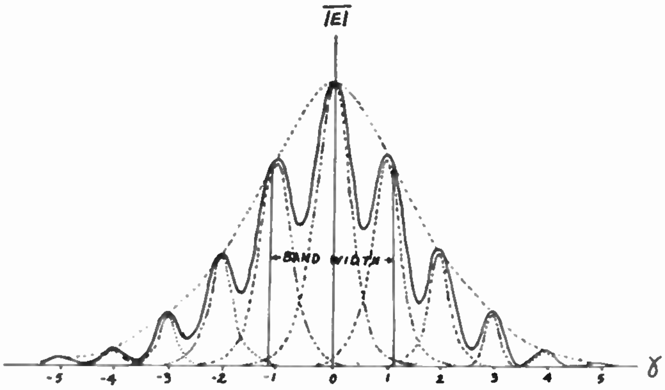

Fig. 4—Frequency response with pronounced multiple resonance [$P/a > 1$].

show how the response appears for $p/a \ll 1$ and for $p/a > 1$. In the latter case the over-all response is shown as the resultant of many responses, each with a resonance at an integral value of γ. Since the bandwidth of each component response is 2α, the condition $p/a > 1$ means that the spacing between resonances is more than half the bandwidth of each response. The condition $p/a \ll 1$, on the other hand, implies that each response is broad compared with the spacing between resonances, so that a smooth over-all response, as in Fig. 3, is the result.

The envelope of oscillations observed when an oscilloscope synchronized to the quench frequency is placed across the nodes is given by $|E|$, plotted as a function of time. This can be obtained from (12) by first separating E into its real and imaginary parts and computing these parts for given α, β, and γ as functions of the angle pt. Both $|E|$ and the phase angle ϕ can thus be obtained for special cases.

The work of computing $|E|$ and ϕ as functions of pt for $\alpha=2$, $\beta=6$, and several values of γ was given to the Center of Analysis of the Massachusetts Institute of Technology. The results are plotted in Fig. 5. At the top of Fig. 5, the function $u=a-b \sin pt$ is plotted for reference. It is noteworthy that the point at which each envelope has its maximum is the point at which the damping factor or decrement u has a zero with a posi-

[2] For a description of various modes of operation, see F. W. Frink, "The basic principles of superregenerative reception," Proc. I.R.E., vol. 26, pp. 76–107; January, 1938.

tive slope. Physical considerations predict this, since oscillations grow when the damping factor is negative and decay when it is positive. It has been assumed that u does become negative, since in an actual superregenerative circuit there is a growth and decay of oscillations. It has also been assumed that $a>0$, since for $a<0$ the transient term, omitted in the solution, would grow to infinity rather than vanish with time.

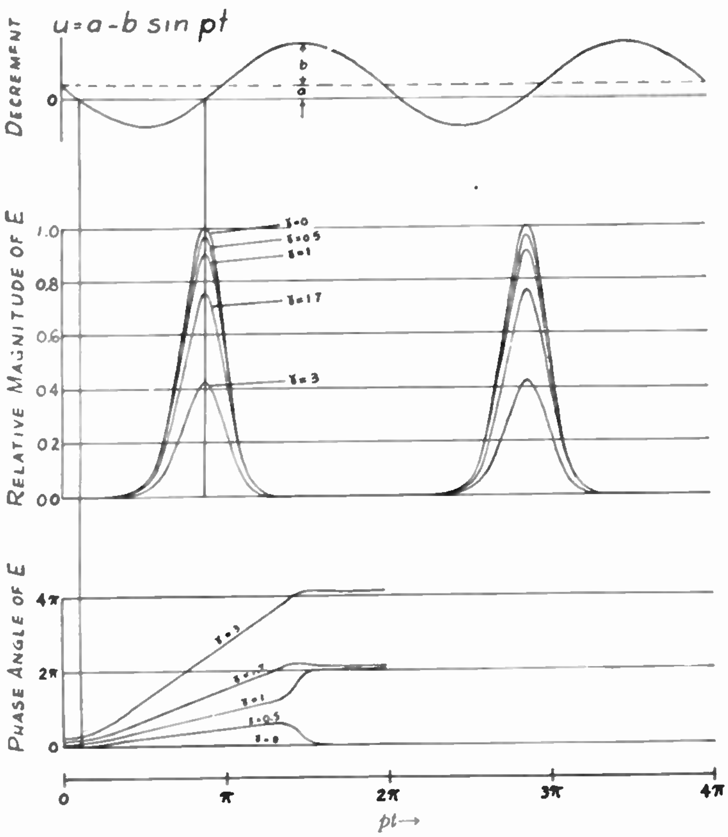

Fig. 5—Decrement u and magnitude and phase angle of the envelope E for $\alpha=2$, $\beta=6$, and several values of γ.

The phase angle ϕ is in every case very approximately a straight-line function of pt with slope equal to γ, except during the intervals in which $|E|$ is very small. Moreover, all these straight lines pass through the point where u has a zero with a negative slope. With these facts, the approximate equation relating ϕ to pt may be written as

$$\phi = \gamma\left(pt - \arcsin \frac{a}{b}\right), \tag{14}$$

except when $|E|$ is very small. When $|E|$ becomes small, ϕ quickly approximates some constant value and remains fairly constant until $|E|$ begins to grow large again. Relation (14) is identically true for $\phi=0$.

Since the instantaneous voltage is

$$e = |E| \epsilon^{j(\omega t+\phi)}, \tag{15}$$

relation (14) yields

$$e = |E| \epsilon^{j(\omega_0 t-\gamma \arcsin (a/b))}. \tag{16}$$

It may be concluded from (16) that, while $|E|$ is large, free oscillations exist (i.e., oscillations at frequency ω_0), and from (15) that, while $|E|$ is small and ϕ is constant, the oscillations are forced (i.e., at frequency ω).

IV. Sensitivity and Selectivity

The useful output of a superregenerator is that obtained by rectification and filtering of the oscillation trains. A measure of this output is the average value of the magnitude of the complex envelope. This is given by

$$\overline{|E|} = \frac{1}{2\pi}\int_{-\pi}^{\pi} |E| \, d(pt). \tag{17}$$

The value of this integral will be independent of time except for the slow variations of modulation frequency in the amplitude of the signal current.

An approximate expression for $\overline{|E|}$, which holds for integral values of γ, is

$$\overline{|E|} = \left| \frac{I}{G_0} \sec\left(\gamma \arcsin \frac{\alpha}{\beta}\right) \cdot \sum_{n=-\infty}^{\infty} \frac{(-1)^n I_n(\beta) I_{n-\gamma}(\beta)}{1+\left(\frac{n-\gamma}{\alpha}\right)^2} \right|. \tag{18}$$

Its derivation is based on the assumption that the phase angle obeys (14). This assumption appears to be safe enough, since it has been seen to be good except for intervals which do not contribute measurably to the integral in (17).

Equation (18) may be used to obtain the frequency response when α and β are given. Only the multiple-resonance peaks can be computed; i.e., the response for integral values of γ. These points are sufficient, since the bandwidth is best defined by means of a smooth simple curve, as the outer dotted curve of Fig. 4, joining them. If the upper and lower -3-db points on this curve are γ' and γ'', then $\gamma'-\gamma''/2\alpha$ is the ratio of the bandwidth of the superregenerator to that of the reference circuit. This ratio, here called the bandwidth factor, may be obtained for any α and β by constructing a response curve from (18) and measuring the bandwidth graphically at -3 db. The services of the MIT Center of Analysis were enlisted in computing the response data for many pairs of values of α and β. From these data the bandwidth factor was obtained and tabulated as a function of

$$\frac{p}{a} \equiv \frac{1}{\alpha} \quad \text{and} \quad \frac{b}{a} \equiv \frac{\beta}{\alpha}.$$

This resulted in the contours of constant-bandwidth factor, one of the two families of curves given in Fig. 6.

The plane on which these contours are plotted may be called the variational damping plane, since its coordinates are the parameters of the variational damping factor $b \sin pt$ normalized to the average damping factor a, which is also the half-bandwidth of the reference circuit. The curve of bandwidth factor 1.0 is the locus of points on the variational damping plane where the superregenerator has the same selectivity as the reference circuit. Below this curve the superregenerator is more selective, and above it less selective, than the reference circuit. The slope of the damping-

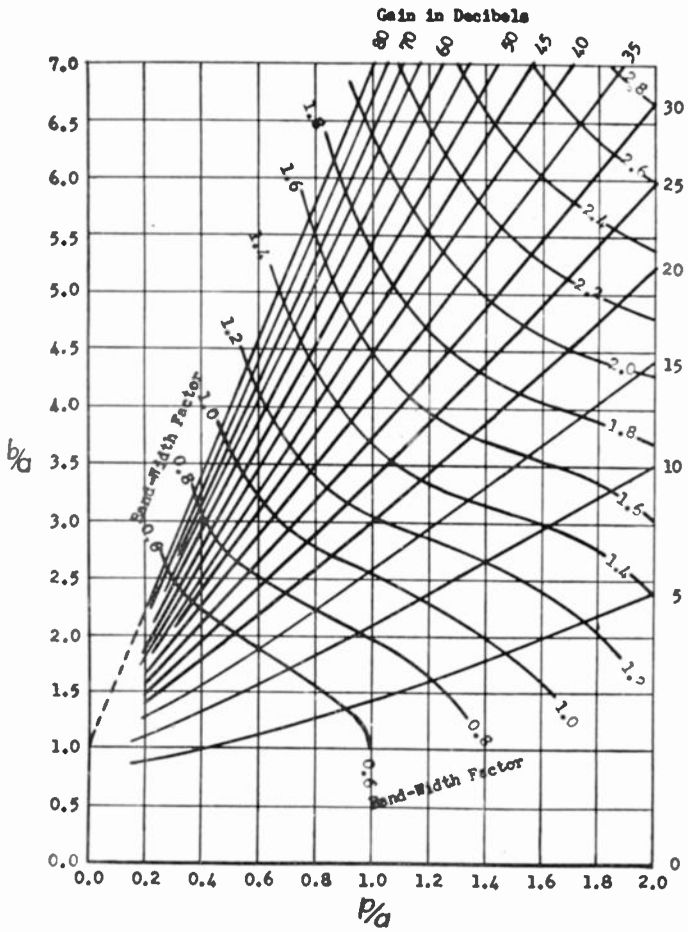

Fig. 6—Contours of constant variational damping gain and contours of constant-bandwidth factor on the variational damping plane.

factor curve, Fig. 5, at the point where $u=0$ with a negative slope is $-p\sqrt{b^2-a^2}$. If this is set equal to a constant, say $-ka^2$, the resulting equation is

$$\frac{p}{a}\sqrt{\left(\frac{b}{a}\right)^2 - 1} = k. \tag{19}$$

If this equation is plotted on the variational damping plane, the resulting curve will be seen to agree fairly well with one of the contours of constant bandwidth factor, depending on the value assigned to k. This agrees with the observation, often made by students of superregeneration, that the selectivity is determined for the most part by the slope of the damping function where it crosses the time axis into the negative region.[3] The bandwidth increases with this slope.

If (18) is specialized for the case of resonance ($\gamma=0$) and divided through by I/G_0, which is the amplitude of the voltage across the reference circuit at resonance, the resulting ratio is

$$A \equiv \sum_{n=-\infty}^{\infty} \frac{(-1)^n I_n^2(\beta)}{1+\left(\frac{n}{\alpha}\right)^2}, \tag{20}$$

which may be called the variational damping gain, as it expresses the advantage of the superregenerator over the reference circuit, from which it differs only in the variational damping term. This variational damping gain is a good measure of the relative sensitivity of the superregenerator. Values of A may be calculated and tabulated for pairs of values of p/a and b/a, as for the bandwidth factor, and constant-A contours may then be plotted on the variational damping plane. These are given in Fig. 6 with the values of the ratio A converted to decibels.

The area bounded by the time axis and the damping-factor curve, plotted in Fig. 5, in the interval where $u<0$, is

$$2\left(\frac{a}{p}\arccos\frac{a}{b} - \frac{1}{p}\sqrt{b^2-a^2}\right).$$

If this area is set equal to a constant, say $-2k$, the resulting equation is

$$\sqrt{\left(\frac{b}{a}\right)^2 - 1} - \operatorname{arcsec}\frac{b}{a} = k\frac{p}{a}. \tag{21}$$

If (21) is used to plot p/a as a function of b/a for various values of k, a family of curves similar to the constant-A contours is obtained. In particular, these curves all pass through the point $(p/a)=0$, $(b/a)=1$, as the contours appear to do. This supports the contention often made that the sensitivity of a superregenerator in the linear mode is primarily a function of the area bounded by the time axis and the damping function, where the latter is negative.

Experimental evidence for the constant-A contours is available in the form of contours of constant output voltage at modulation frequency plotted on a plane whose co-ordinates are quench frequency and quench amplitude. These experimental contours are the work of Sze-Hou Chang,[4] who completed them prior to the present analysis. They agree very well with the constant-A contours. All of them appear to pass through a point on the quench-amplitude axis, and all approximate straight lines. The similarity applies only to the region where Chang's circuit was operated in the linear mode. In the region on the other side of Chang's "transition line," which separates linear mode from logarithmic mode, the behavior is totally different. The transition line is approximately straight, and in the general direction of the constant-output contours. Obviously, Fig. 6 should be expected to apply to any given circuit only where that circuit is linear, since the analysis here given assumes linear operation.

[3] G. G. Mac Farlane and J. R. Whitehead, "The superregenerative receiver in the linear mode," *Jour. IEE* (London), vol. 95, part III, pp. 143–157; May, 1946.

[4] Sze-Hou Chang, "Theoretical and Experimental Studies in Superregeneration," Ph.D. Thesis, Cruft Laboratory, Harvard University, October, 1947.

A Modulator Producing Pulses of 10^{-7} Second Duration at a 1-Mc Recurrence Frequency*

MILLETT G. MORGAN†, SENIOR MEMBER, IRE

Summary—**The development of a pulse generator and a modulator which will produce pulses of 10^{-7} second duration and 3,000-volt amplitude at a 1-Mc recurrence frequency is described. The modulator is developed for the special case of a nonlinear load impedance consisting of 125 $\mu\mu$f capacitance for the leading half of the pulse and a spark discharge for the trailing half. The high recurrence frequency gives rise to problems not generally encountered in the design of pulse circuits. Other applications, such as the second modulation of a high-recurrence-frequency pulsed carrier with voice frequencies, suggest themselves.**

A new pulse-generating circuit, capable of producing positive pulses whose duration is one-tenth the period of the 1-Mc input sine wave, is developed. This pulse generator can be operated at moderate power levels sufficient to drive the modulator.

1. Introduction

As an approach to the radar countermeasures problem, it was desired to produce brief, rapidly recurring rf pulses of high intensity. With brevity, bandwidth is achieved, and rapid recurrence aids in producing contiguous images on the radar oscilloscope, thus blanketing the screen. The modulator described herein was developed for use with a damped-wave spark oscillator whose pulse duration was approximately 2×10^{-8} second.[1] Because deionization was required between pulses, it was desired that the modulating voltage be removed as soon after the termination of the rf pulse as possible, Not only does this increase the time available for deionization, but it reduces the spread of ionization which occurs if an arc is permitted to ensue after the rf pulse is ended; and it improves the efficiency of the modulator. From the point of view of modulator design, the use of this kind of load has disadvantages and advantages: the modulating voltage must be built up across a wholly capacitive load, and this requires low modulator resistance; the occurrence of the spark, however, reduces the load impedance to a low value and aids in producing a rapid decline of the modulation voltage.

2. Specifications for the Modulator

The choice of pulse duration, recurrence frequency, and amplitude were influenced by the known characteristics of the spark oscillator, the intended application, and estimated design limitations.

One can estimate the required recurrence frequency for jamming efficacy as follows: As an example, consider the Navy Type FD fire-control radar. It has a recurrence frequency of 1,640 cps and sweep length of 100,000 yards range. This represents a sweep duration of 610 microseconds. The time between transmitted pulses is 1/1,640 = 610 microseconds also. Hence, in this case, the sweep lasts for the full time between pulses. Consider pulses being received at a 1-Mc recurrence rate. In 610 microseconds, 610 pulses would be received. Taking an actual sweep length on the cathode-ray tube of 4 inches, each pulse would be separated by 4/610 or 0.00656 inches. It is apparent that the useful part of the screen would then be fully occupied and that no echo could be discerned. In order to find the shortest sweep duration which can be effectively blocked in this manner, it is necessary to establish the minimum spot diameter which can be used on the indicator tube. Taking 0.025 inch as the practical minimum, and dividing the sweep length (4 inches) by this dimension, we see that 160 is the minimum number of pulses required to blank the tube. For a 1-Mc recurrence frequency, this corresponds to a sweep duration of 160 microseconds, or 26,300 yards. For shorter sweeps, some jamming effectiveness would be lost without going to a recurrence frequency greater than 1 Mc.

Above about 50 kc, gas tubes are not usable,[2] and so we may consider that any recurrence frequency which will be effective for jamming in this way will require hard-tube pulse generation and modulation. In radar equipments using gaseous-discharge transmit-receive tubes of the best design, it is possible to observe echoes from targets as little as 200 yards away. This represents an elapsed time of only 1.2 microseconds for transit of the signal, and it means that the t-r tube must be sufficiently deionized in this time to switch on the receiver input. Also, because echoes from near-by objects are large, the deionization must be sufficiently complete that the strong received signal will not re-establish the discharge. Based on this consideration, and the computation preceeding, 1 Mc was selected as a recurrence frequency which would be at the limit of operation of the oscillator and would have worth-while jamming effectiveness.[3]

The desired modulator pulse width would be something less than the rf envelope width. Practically, this

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† Thayer School of Engineering, Hanover, N. H.

[1] This paper is confined to a description of the pulse generator and modulator.

[2] L. G. Kersta, "Gas-tube harmonic generator," *Bell Lab. Rec.*, vol. 22, p. 53; October, 1943.

[3] It may be reasoned similarly that gas tubes could be designed to operate as pulse generators and modulators up to 1 Mc. However, it was desired to confine this problem to the oscillator.

means that the shortest attainable modulator pulse would be the best. 10^{-7} second was taken as a minimum specification, and was the best obtained at the required output amplitude.

The choice of modulator pulse voltage was determined by minimum requirements.[4] Whereas the jamming effectiveness is an increasing function of the signal strength, it was not the intent in the initial design to attain more than moderate amounts of power. It was felt that with the available tube resistances, the oscillator capacitance (125 $\mu\mu$f) could be charged to 3,000 volts, using a 6,000-volt source, in something appreciably less than 10^{-7} second.

3. Approach to the Circuit Problem

High-level amplification of very short pulses is limited to one stage, with a positive pulse input. This results from the fact that amplification causes polarity reversal, and amplifiers having negative pulse inputs have intolerably low efficiency because they are conducting during the quiescent period.[5] The spark oscillator offered no polarity restriction on the modulator output. It was desired to produce also a 250-kc synchronous, linear sweep. Because of instability and synchronization problems, no form of relaxation oscillator was seriously considered. This follows the trend of conservative radar design.

Briefly, then, the approach was to use a 250-kc base oscillator, derive a 250-kc linear sweep from this for indication, and, from a harmonically generated 1-Mc sine wave, generate a positive pulse of sufficient amplitude to drive a normally cut-off amplifier to required output. Because of the short pulse duration and high duty ratio (0.1), a pulse transformer was not considered.

4. Circumstances Leading to the Discovery of a New Pulse-Generating Circuit

Overdriving a "resistance-coupled" amplifier reveals an interesting effect when grid-leak bias is used with no cathode or fixed bias. If the grid is initially at ground potential and the driven side of the coupling capacitor is at some value, above ground, corresponding to the axial value of the sinusoidal signal applied, the grid potential rises as the applied signal rises, the grid-cathode region becomes conducting, and a low-resistance circuit is established through the tube. If the coupling capacitor is not too large, it will charge nearly as fast as the applied signal rises, and will continue to do so until the crest of the driving wave is reached, at which point the driving voltage starts to fall. Immediately, the grid voltage is carried downward by the capacitor, and the grid-cathode region becomes nonconducting. The capacitor, which has taken on an additional charge during the rising portion of the applied signal, then begins to discharge into the grid-to-ground resistor. This discharge, together with the applied signal, controls the instantaneous potential of the grid. The amount of the discharge determines the point on the driving wave at which the grid will again become positive and conducting. Because the capacitor has not discharged completely to its initial voltage, this point will be above the axis of the driving wave. This operation may be thought of as, in effect, moving the axis of the driving wave downward so that only the crest portion is amplified by the tube. This is somewhat in the manner of a class-C amplifier, biased so that plate current flows for less than 180°. It was found, however, that steep negative pulses with flat tops could be produced by the automatic biasing arrangement, whereas with fixed bias the pulses had much smaller amplitude for the same width (angle of plate-current flow) and did not have flat tops.

By properly proportioning the circuit constants, very satisfactory negative pulses of 2.5×10^{-7} second duration were produced in this way across a plate-circuit resistor. Furthermore, it was found that, by use of a low-resistance triode, 125 $\mu\mu$f of output capacitance, plus the output capacitance of the tube, could be charged nearly to supply voltage in 0.5×10^{-7} second. Of course, a long trailing edge is produced unless the spark oscillator is connected. By means of an inductor shunting the plate load resistor, and a clipper tube, a negative pulse with a steep trailing edge can be had. It will be recalled that, for negative pulses, the clipper tube can be connected in such polarity that the interwinding capacitance of its filament transformer does not shunt the pulse output circuit.

Reasonably good results were also obtained by overdriving a cathode follower. With the input circuit of the previous amplifier unchanged, the plate load resistor was moved to the cathode, and positive pulses obtained. However, because the cathode cannot rise above the instantaneous grid potential, the output pulse cannot ex-

[4] The behavior of low-voltage sparks is erratic. This is due principally to the close gap spacing required. Depending upon the condition of operation, electrode material may be carried away, thus increasing the gap spacing, or the gap may be bridged by the formation of compounds at the electrodes.

[5] The bootstrap circuit (cathode follower with degeneration removed by returning the input to the cathode instead of to ground) will produce voltage amplification without polarity reversal. However, the entire driving circuit must be isolated from ground. This means that all of the capacitance to ground must be charged and discharged at a rate equal to the rate of rise and fall of the pulse. (See G. N. Glasoe and J. V. Lebacqz, "Pulse Generators," Vol. 5, Radiation Laboratory Series, McGraw-Hill Book Co., Inc., New York, N. Y., 1948; p. 123.) For pulses of the order of 10^{-7} second duration, this is not feasible. Another source of capacitance to ground is that of the cathode itself. In the case of small tubes, this is manifest in the capacitance of the cathode sleeve to the heater, and may be as low as 5 to 10 $\mu\mu$f. However, inasmuch as it is not possible to insulate the cathode sufficiently well from the heater to permit the generation of large signal voltages on the cathode, while the heater is grounded, filament-type tubes must be used in large amplifiers, and the cathode must be tied to the heater in medium-power cathode-type tubes. The principal source of capacitance to ground then becomes the interwinding capacitance of the filament transformer. This is commonly of the order of 100 to 300 $\mu\mu$f, but may be materially reduced by special design. This cathode-to-ground capacitance is also very troublesome whenever it is desired to connect a tube upside down, so to speak; i.e., with its cathode, rather than its anode, following the instantaneous pulse voltage. This generally occurs in using a clipper tube with a positive pulse, but not with a negative pulse. It also arises when attempting to linearize high-speed sweeps with a constant-current series pentode.

ceed in amplitude, nor rise as fast as, that portion of the driving wave being used when the grid is above cut-off voltage. Also, both the filament-to-ground capacitance of the cathode follower and the clipper tube now appear across the pulse output. This situation can be improved somewhat by using tubes requiring the same filament voltage, permitting the use of a common filament transformer.

The radiation-cooled tantalum-electrode tubes with high-melting-point, noncontaminative grids were found to be the only type which would operate indefinitely under these conditions. Good results were had with such tubes as the 100TL or 250TL. These tubes can readily withstand the required grid dissipation. Parallel operation was found to be desirable within limits, inasmuch as the advantage gained by paralleling the tube resistance more than offsets the disadvantage incurred by increasing the output capacitance when there is additional external capacitance connected. At the same time, this increases the available plate dissipation. The 304TL, containing four parallel-connected triode sections within one evacuated envelope, yields a lower output capacitance than can be had when four separate envelopes are used with external connections. This was found to be the most satisfactory tube tried, and dispelled any ideas of building a special tube.

The Western Electric type 715C, a tetrode developed for use as a pulse modulator for multicavity magnetrons, is not satisfactory as an overdriven amplifier. Because it is intended for use with signals having a much lower duty cycle, its plate dissipation is low; and because it is used with a low-capacitance load, its resistance is not especially low, although the peak current provided by the large oxide-coated cathode is large for large applied voltage.

A simple calculation will show how low the modulator-tube resistance must be to charge the oscillator capacitance in the required time. Using an oscillator capacitance of 125 $\mu\mu$f, a time of charge of 0.5×10^{-7} second, a voltage reached of 3,000, and a charging voltage of 5,000, we have

$$e = E(1 - \epsilon^{-t/RC})$$

$$R = \frac{t}{C \ln \dfrac{E}{E-e}} = \frac{\frac{1}{2} + 10^{-7}}{125 + 10^{-12} \ln \dfrac{5{,}000}{5{,}000 - 3{,}000}}$$

$$= 437 \text{ ohms.}$$

The mean transconductance given by the manufacturer for the type 304TL tube is 16,700 micromhos, and the average value of amplification factor is 12. This gives a plate resistance of $12/16{,}700\times10^{-6} = 720$ ohms. Two tubes in parallel then give a value of 360 ohms. Allowing for approximate values used, this compares well with the computed value required.

Although negative pulses rising in less than 0.5×10^{-7} second were produced with the overdriven amplifier, using the required output capacitance, no means was found to reduce the pulse duration beyond 2.5×10^{-7} second. One idea that came to mind was to produce a sharp negative pulse in the plate circuit of an overdriven amplifier and then to apply a negative pulse, similarly generated in another overdriven amplifier, to the grid of the first overdriven amplifier to "turn the tube off" before the originally applied sine wave had done so. This circuit would require that the phase of the shut-off pulse be properly controlled.[6]

In order to try out this circuit, two identical 1-Mc power stages were provided. The inputs to these were connected in parallel and supplied by the driver.[7] The necessary phase shift was obtained with delay lines.

Difficulties arose from attempting to apply the negative shut-off pulse to the grid of the modulator tube. Trouble was experienced in arranging the input impedance of the first overdriven tube, together with the impedance seen looking back into the sinusoidal amplifier, so that the negative shut-off pulse from the second overdriven tube could be developed across it. However, the troublesome feeding of sinusoidal voltage to the plate of the shut-off tube resulted in the discovery of a circuit which proved to be a solution to the problem.

5. A New Pulse-Generating Circuit

A large sinusoidal 1-Mc voltage wave is applied to the grid of a triode having low internal resistance. A grid coupling capacitor, and a grid leak are used to supply automatic bias. This is the arrangement described for the generation of a negative pulse in the plate circuit, 2.5×10^{-7} second long, with a steep leading edge. Then, instead of applying a dc voltage to the plate, a 1-Mc sinusoidal signal is applied and its phase so adjusted that, after it rises for some distance beyond zero, the input signal on the grid causes the tube to conduct and reduces the impedance of the tube to a low value. This can be accomplished by feeding voltage to the plate, from the grid, through the grid-to-plate interelectrode capacitance. A resistance is then connected from plate to ground. When the tube is not conducting, the sine voltage is developed across the plate-to-ground resistance. When the tube is conducting, the voltage is developed across the much lower tube resistance. The series circuit made up of the grid-to-plate capacitance and plate-to-ground resistance draws a leading current. Hence, the voltage on the plate leads that on the grid. By properly controlling this angle of lead, the voltage on the plate is allowed to rise for something less than 10^{-7} second and is then caused to collapse as the tube becomes conducting, producing a brief positive pulse on the plate. The phase angle of the sinusoidal voltage on the plate, and

[6] Because the generated pulse would be negative, no subsequent amplification would be possible.

[7] The driver contained the 250-kc oscillator and provided synchronous 1-Mc sinusoidal and 250-kc sawtooth outputs.

hence the length of the pulse, can be adjusted as desired and an over-all length of 10^{-7} second was used.[8] The circuit is shown in Fig. 1.

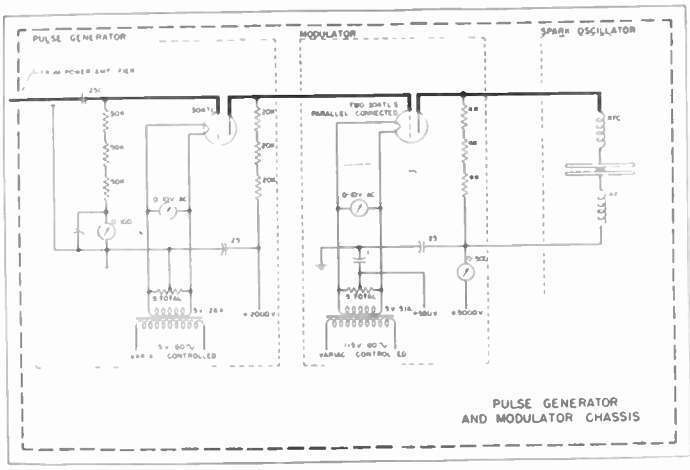

Fig. 1—Schematic diagram of the pulse generator and modulator chassis.

It is to be noted that an essential feature of this pulse generator is that it utilizes the steep, inflectional portion of the sine wave, rather than the slowly changing crest portion. An appreciation of this difference is had in comparing the Peterson coil[9] with the overdriven amplifier.

In order to use the steepest part of the wave, it is desirable to raise the axis above zero so that each side of the inflectional region may be utilized. The application of plate bias to accomplish this is somewhat troublesome due to the loss of voltage across the plate resistor. An attempt was made to shunt-feed this bias to the plate, but no satisfactory isolating impedance was found through which some current could be drawn. The problem is made difficult by the fact that the plate-to-ground resistor is large (60,000 ohms). Best results were finally secured by returning the resistor to a sufficiently high-voltage (2,000 volts), low-impedance source.

Some work was done on other arrangements for applying sinusoidal voltage of the proper phase to the plate of the pulse-generating tube. If an inductor were used from plate to ground, instead of a resistor, the application of plate bias would be greatly simplified. If the series combination of this inductance and the grid-to-plate capacitance were tuned to the inductive side of series resonance, a lagging current would be drawn. The voltage across the inductance would lead the current by 90°. Inasmuch as it is desired to have the voltage lead by only a small amount for the production of short pulses, it is apparent that the inductive reactance would have to be large compared with the capacitive reactance. However, it is not desirable to attempt to make the inductance too large, for the distributed capacitance then rises rapidly. Likewise, it is not desirable to increase the grid-to-plate capacitance too much in order to increase the disparity between the inductive and capacitive reactances, for then the time taken to charge this capacitance, when the tube conducts, becomes appreciable, and the rate of fall of the pulse is too low. In general, it was found to be more satisfactory to use a resistor in the plate circuit.

The parallel combination of the impedance from grid to ground, via the plate (series combination of grid-to-plate capacitance and plate resistor) and the direct grid-to-ground impedance (parallel combination of grid-to-cathode capacitance and grid resistor), determines the load seen by the exciting power amplifier. Since it is desirable to develop as large a signal as possible on the plate of the pulse generator, these impedances are kept as large as possible so that the voltage on the grid will be large.[10]

The high-voltage sine-wave input to the pulse generator is brought from the power amplifier with a coaxial cable (RG-11/U). In order to obtain proper operation of the biasing arrangement in the grid circuit of the pulse generator, it is necessary to place the dc blocking capacitor at the load end of the coaxial line rather than in the power-amplifier chassis. If this is not done, the cable impresses a large capacitance across the input of the pulse generator. This capacitance draws a large steady-state sinusoidal current and perturbs the "overdriven" operation of the tube. The tank impedance of the power amplifier is limited by the minimum capacitance which can be used, and thus principally by the length of coaxial output cable required.

6. The Modulator

The positive pulses thus generated are applied to the grid of the modulator by direct connection from the pulse-generator plate to the modulator grid. Because the input capacitance of the modulator interferes with the operation of the pulse generator,[8] this capacitance is kept as low as possible.

There is a considerable negative swing of the voltage at the plate of the pulse-generator tube due to the negative half of the applied sine wave (see Fig. 2). To eliminate this by means of a clipper would mean connecting the filament of the clipper tube to the plate of the pulse-generating tube. This would again introduce the problem of interwinding capacitance in the filament transformer. Fortunately, however, this negative swing does

[8] The capacitance of any load connected must be taken into consideration in adjusting the phase angle. Also, because such capacitance reduces the output impedance, it results in reduced pulse amplitude. Hence, the circuit is not suitable for driving high-capacitance loads directly. Fortunately, its output is positive, and can, therefore, be amplified once by a low-resistance modulator.

[9] The Peterson coil generates voltage pulses from a sinusoidal current by abrupt magnetic saturation. Although the symmetrical output is attractive for application to the spark oscillator, the efficiency of the device is much too low for high-level operation. E. Peterson, J. M. Manley, and L. R. Wrathall, "Magnetic generation of a group of harmonics," *Bell Sys. Tech. Jour.*, vol. 16, p. 437; October, 1937.

[10] As the grid resistor is increased, the coupling capacitance must be reduced in order to preserve the desired time constant for "overdriven" operation.

The plate resistor can be altered only by changing the grid-to-plate capacitance. Increasing the capacitance requires reducing the value of the plate resistor to restore the desired phase angle.

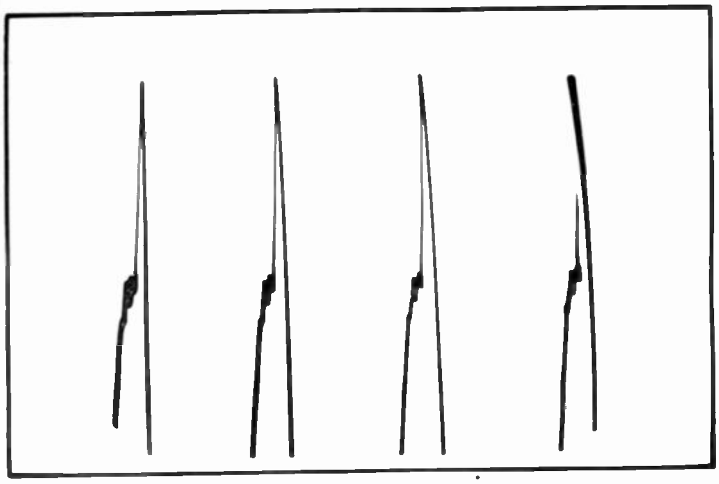

Fig. 2—The pulse-generator output. The pulse recurs 10^6 cps; hence it is clear that the duration (of the upper portion) is 10^{-7} second at the base. The lower portion is eliminated by proper biasing of the modulator. Time progresses from right to left in this figure.

no harm on the grid of the modulator tube provided the tube can withstand the stress and ion current. The tubes used (304TL) do so with ease. In the modulator tubes, just as in the pulse-generator tube, there is a feeding of the grid signal to the plate by means of the grid-to-plate capacitance. In spite of the large negative excursion of the grid, this has not been found objectionable. The effect is actually desirable, for it provides a reversed voltage on the oscillator to assist in ion sweeping.

Enough fixed cathode bias is used to hold the modulator tubes at cutoff when no signal is applied to the grid. The plate bias must not be applied to the pulse generator until a signal voltage is present. If this precaution is not taken, the pulse-generator plate bias will place a dc potential on the modulator tube grids which will, in turn, rise to a potential equal to the modulator cathode bias. This will cause the modulator tubes to become conducting, and excessive power will be dissipated at the plates.

Fig. 3 is an oscillogram of the voltage on the modulator plate; i.e., the modulating voltage developed across

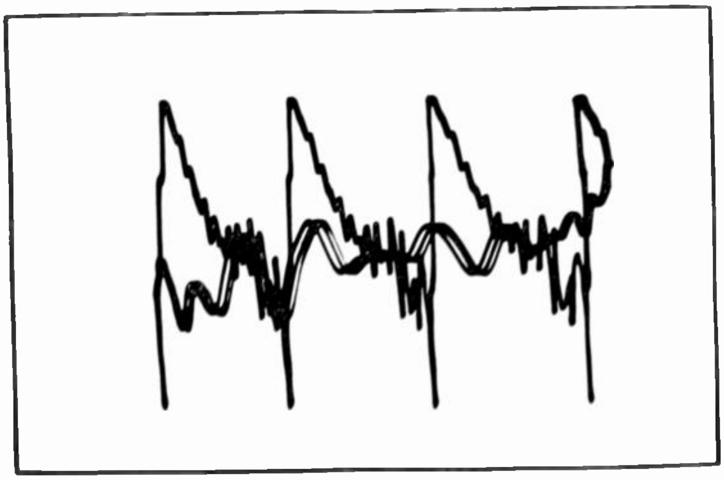

Fig. 3—The modulator output with the spark oscillator connected. The oscillations following each pulse are spurious and are generated by the indicator. Time progresses from right to left.

the spark oscillator. This oscillogram is somewhat unsatisfactory because of the oscillations shown following the pulse. In order to view the modulating pulse with the spark present in the oscillator for rapid discharge of the output capacitance, it is necessary to employ so much voltage that the pulse cannot be viewed without the use of a dc-blocking and amplitude-control capacitor. Enough has been observed on the defocused tube by direct connection and using excessive centering voltages to show conclusively that the spurious oscillations are initiated in the indicator circuit with each pulse and are not actually present across the oscillator with the indicator directly connected.[11] The amplitude of this pulse was successfully measured with the aid of a specially designed low-capacitance filament transformer having an interwinding capacitance of only 25 $\mu\mu$f, a diode which could withstand a large inverse voltage (HK-253), and a high-voltage power supply. With the diode's filament connected to the modulator plates, and the diode's plate connected to the power supply, and by-passed to ground, the diode plate voltage was raised until current just began to flow in the diode. This value of diode plate voltage then represented the minimum voltage of the modulator tubes' plate excursion. The maximum voltage reached by the modulator tube plates is somewhat less than the supply voltage if the output capacitance is not fully discharged between cycles. This can be measured by connecting the diode plate to the modulator plates and returning the diode filament to a source of high voltage. The voltage to which the diode filament is returned is then reduced from an amount equal to the modulator supply voltage until diode current commences to flow. This voltage then represents the maximum voltage of the modulator tubes' plate excursion. By subtraction, the total excursion is then obtained, and thereby the amplitude of the output pulse.

By this means, the amplitude of the pulse shown in Fig. 3 was found to be 1,100 volts. It was found that, with the spark in operation, an output pulse of greater amplitude could not be obtained, presumably due to incomplete deionization. However, with the spark oscillator disconnected and 125-$\mu\mu$f vacuum capacitors connected in its stead, a 3,000-volt output pulse was obtained with a modulator supply voltage of 4,500 volts.

7. Acknowledgments

The author gratefully acknowledges the helpful suggestions and criticisms offered by the following men during the conduct of this work: Hugo Benioff, associate professor of seismology, and F. E. Lehner, research assistant, California Institute of Technology. He is also indebted to the late Harold J. W. Fay, president, and I. C. Clement, vice-president in charge of engineering, Submarine Signal Co., for sponsorship.

[11] See chap. by O. T. Fundingsland, p. 673, of footnote reference 5.

Circuits for Traveling-Wave Tubes*

J. R. PIERCE†, FELLOW, IRE

Summary—The comparison between traveling-wave-tube circuits of different geometries can be facilitated by the use of phase velocity, group velocity, and stored energy as parameters. For a given stored energy per unit length, lowering the group velocity with respect to the phase velocity increases circuit impedance (and, hence, gain), increases attenuation, and narrows the band. A filter-type circuit consisting of pillbox resonators is shown to be much inferior electrically to a helix. Expressions are presented showing the effect of gap length in filter-type circuits consisting of pillbox resonators, and the attenuation is calculated for such circuits.

Introduction

VARIOUS CIRCUITS have been used or proposed for traveling-wave tubes: the helix, coiled waveguides, and a series of resonators coupled together to form a filter. It is often desirable to regard the beam voltage, or the phase velocity, as a design parameter, to be held constant in comparing circuits. In this case, the goodness of circuits, as far as gain per wavelength is concerned, is expressed by the parameter

$$(E^2/\beta^2 P)^{1/3}. \tag{1}$$

The quantity $(E^2/\beta^2 P)$ has the dimensons of impedance. E is the peak longitudinal field acting on the electrons, P is the power flow in the circuit which produces the field E, and β is the phase constant;

$$\beta = \omega/v_\phi \tag{2}$$

where ω is radian frequency and v_ϕ is phase velocity. For optimum gain, the electron velocity should be near to this phase velocity. If we neglect ac space charge and loss, the electron velocity v for optimum gain is equal to the phase velocity, and we may write

$$v_\phi = v = 1.98 \times 10^{-3} V_0^{1/2} c. \tag{3}$$

Here and later, mks units are used. V_0 is the voltage by which the electrons are accelerated, and c is the velocity of light.

For this case of no circuit attenuation and an electron velocity equal to the phase velocity, if we neglect the effect of space charge the net gain is

$$G = -9.54 + 47.3CN \text{ db}. \tag{4}$$

$$C = (E^2/\beta^2 P)^{1/3}(I_0/8V_0)^{1/3}. \tag{5}$$

Here N is the length of the circuit in wavelengths, I_0 is the beam current in amperes, and V_0 is the accelerating voltage.

1. Phase Velocity and Group Velocity

The phase velocity is a relation between the wavelength of the traveling wave and the frequency, both of which are measurable quantities. Thus

$$v_\phi = f\lambda \tag{6}$$

where f is the frequency, and λ is the wavelength.

The phase velocity need not be the velocity with which a broad pulse travels down the circuit; a pulse travels with the group velocity v_g. The group velocity may be different in magnitude and even in sign from the phase velocity. The phase velocity does not even tell us in which direction energy is flowing.

Suppose we think of a pulse passing along a lossless transmission system. When the pulse is present in a certain portion of the system, there is stored electric and magnetic energy in that portion of the system. Later this energy will be present further along the system. Thus, the group velocity is the velocity with which stored energy travels along the system. If W is the stored energy per unit length, the power flow P is

$$P = Wv_g. \tag{7}$$

Hence,

$$(E^2/\beta^2 P)^{1/3} = (E^2/\beta^2 W v_g)^{1/3}, \tag{8}$$

and

$$C = (E^2 I_0/8\beta^2 W v_g V_0)^{1/3}. \tag{9}$$

Here W is the stored energy per unit length which produces a field of peak strength E at the position of the electron beam.

The group velocity is given by the relation

$$\frac{1}{v_g} = \frac{\partial\beta}{\partial\omega} = \frac{1}{v_\phi}\left(1 - \frac{\omega}{v_\phi}\frac{\partial v_\phi}{\partial\omega}\right). \tag{10}$$

$$v_g = v_\phi\left(1 - \frac{\omega}{v_\phi}\frac{\partial v_\phi}{\partial\omega}\right)^{-1}. \tag{11}$$

We see that the group velocity is equal to the phase velocity only if the phase velocity does not change with frequency.

2. Gain and Bandwidth in a Traveling-Wave Tube

To get high gain in a traveling-wave tube at a given frequency and voltage (the phase velocity is specified by voltage), we see from (9) that we must have either a small stored energy per unit length for unit longitudinal field, or a small group velocity v_g.

To have amplification over a broad band of frequencies, we must have the phase velocity v_ϕ substantially equal to the electron velocity over a broad band of frequencies. This means that, for very broad-band operation, v_ϕ must be substantially constant, and hence, in a

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† Bell Telephone Laboratories, Inc., New York, N. Y.

broadband tube, the group velocity will be substantially the same as the phase velocity.

If the group velocity is made smaller, so that the gain is increased, the range of frequencies over which the phase velocity is near to the electron velocity is necessarily decreased. Thus, for a given phase velocity, as the group velocity is made less the gain increases but the bandwidth decreases.

In the past, particular circuits have been compared on the basis of $(E^2/\beta^2P)^{1/3}$ and bandwidth. For instance, Field[1] has compared a coiled waveguide structure with a series of apertured disks of comparable dimensions. Both of these structures must have about the same stored energy for a given field strength. He found the coiled waveguide to have a low gain and broad bandwidth as compared with the apertured disks. We explain this by saying that the particular coiled waveguide he considered had a higher group velocity than did the apertured-disk structure. Further, if the coiled waveguide were altered so as to have the same group velocity as the apertured-disk structure, it would necessarily have substantially the same gain and bandwidth.

In another instance, O. J. Zobel of these Laboratories evaluated the effect of broad-banding a filter-type circuit for a traveling-wave tube by m derivation. He found the same gain for any combination of m and bandwidth which made $v_g = v_\phi(\partial v_\phi/\partial\omega = 0)$. We see that this is just a particular instance of a general rule.

3. A Comparison of Circuits

The group velocity, the phase velocity, and the ratio of the two are parameters which are often easily controlled, as by varying the coupling between resonators in a filter composed of a series of resonators. Moreover, these parameters can often be controlled without much affecting the stored energy per unit length. For instance, in a series of resonators coupled by loops or irises, the stored energy is not much affected by the loops or irises unless these are very large, but the phase and group velocities are greatly changed by small changes in coupling.

Let us, then, think of circuits in terms of stored energy, and regard the phase and group velocities and their ratio as adjustable parameters. When we do this, it is found that various physical configurations which promise to be useful in traveling-wave tubes can be simply compared on a generalized basis.

We will consider a case in which all electrons in the beam are acted on by fields of substantially the same strength. Let us first consider a helically conducting sheet of radius a. The upper curve of Fig. 1 shows $(E^2/\beta^2P)^{1/3}(v_\phi/c)^{1/3}$ versus βa. In obtaining this curve (see Appendix I), it was assumed that $v_\phi \ll c$. The field E is the longitudinal field at the surface of the helically conducting cylinder.[2]

The helix has a very small circumferential electric field which represents "useless" stored energy. The lower curve of Fig. 1[3] is based on the stored electric energy of an axially symmetrical sinusoidal field impressed at the radius a. This field has no circumferential component

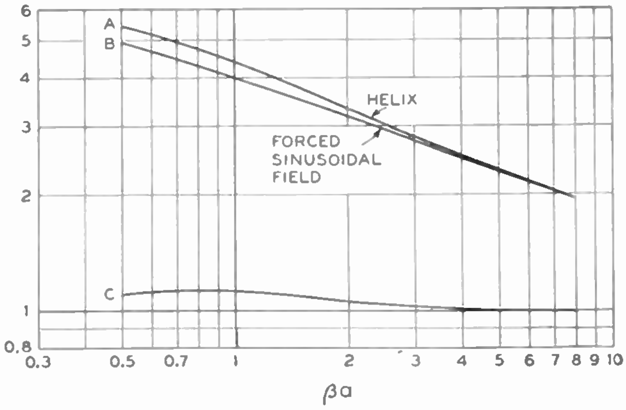

Fig. 1—A comparison of $(E^2/\beta^2P)^{1/3}$ for a helically conducting sheet (curve A) and for a sinusoidal field forced at the same radius (curve B) versus radius in radians, βa. Curve A is $(E^2/\beta^2P)^{1/3}(v_\phi/c)^{1/3}$. Curve B is $(E^2/\beta^2P)^{1/3}(v_\phi/c)^{1/3}(v_g/v_\phi)^{1/3}$. Curve A is higher than curve B because the group velocity is less than the phase velocity for the helically conducting sheet, as shown by curve C, which gives $(v_\phi/v_g)^{1/3}$ versus βa.

but is otherwise the same as the electric field of the helix (again assuming $v_\phi \ll c$). We can imagine such a field propagating because of an inductive sheet at the radius a, which provides stored magnetic energy enough to make the electric and magnetic energies equal. The quantity plotted versus βa is $(E^2/\beta^2P)^{1/3}(v_\phi/c)^{1/3}(v_g/v_\phi)^{1/3}$. For the same values of v_ϕ and v_g, we would expect a slightly larger value of $(E^2/\beta^2P)^{1/3}$ for this impressed sinusoidal field than for the helically conducting sheet. The quantity plotted for the helically conducting sheet does not contain the factor $(v_g/v_\phi)^{1/3}$, as does the quantity plotted for the impressed sinusoidal field. For small values of γa, $v_g < v_\phi$ for the helically conducting sheet. The quantity $(v_\phi/v_g)^{1/3}$ for a helically conducting sheet for which $v_\phi \ll c$ is shown in Fig. 1. This lowered group velocity accounts for the fact that the curve for the helically conducting sheet lies above the curve for the impressed sinusoidal field.

As an example, a filter-type circuit will be compared with the impressed sinusoidal field. This circuit will consist of a series of flat resonators coupled together to make a filter. Fig. 2(a) shows a series of very thin pillboxes with walls of negligible thickness. A small central hole is provided for the electron stream, and the field E is to be measured at the edge of this hole. The diameter is chosen to obtain resonance at a wavelength λ_0. Fig. 2(b) shows a similar series of flat, square resonators.

For the round resonators, it is found that[4]

$$(E^2/\beta^2P)^{1/3} = 5.36(v_\phi/c)^{1/3}(v_\phi/v_g)^{1/3}. \qquad (12)$$

[1] Lester M. Field, "Some slow-wave structures for traveling-wave tubes," Proc. I.R.E., vol. 37, pp. 34-40; January, 1949.

[2] See Appendix IV for discussion of field away from the helix.

[3] See Appendix II.

[4] See Appendix III.

For the square resonators,[4]

$$(E^2/\beta^2P)^{1/3} = 5.33(v_\phi/c)^{1/3}(v_\phi/v_g)^{1/3}. \quad (13)$$

For practical purposes, these are negligibly different.

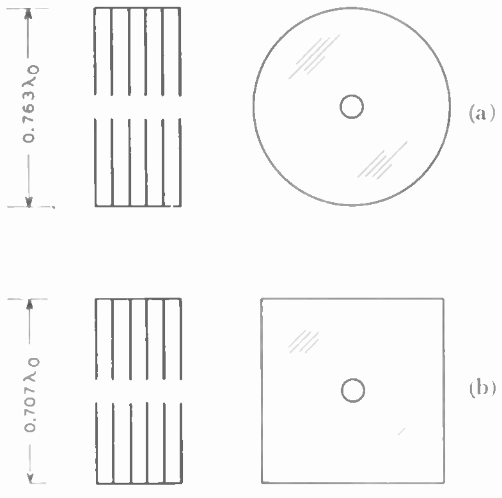

Fig. 2—A filter-type circuit made of pillbox resonators (a), or square resonators (b).

As we have noted, (v_ϕ/c), which appears in the expression for $(E^2/\beta^2P)^{1/3}$ for the sinusoidal field impressed at radius a and in (12) and (13), is a function of the accelerating voltage. Fig. 3 makes a comparison between

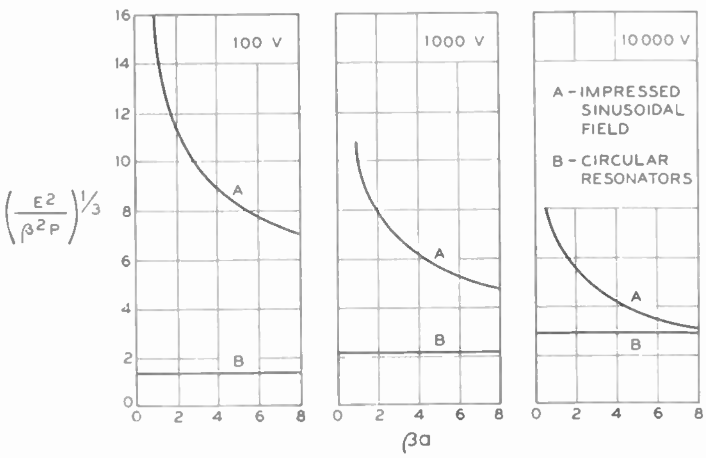

Fig. 3—A comparison of $(E^2/\beta^2P)^{1/3}$ for an impressed sinusoidal field at the radius a and for pillbox or square resonators, for 100 volts, 1,000 volts, and 10,000 volts. For each voltage, $(E^2/\beta^2P)^{1/3}$ for the forced sinusoidal field is a function of the radius expressed in radians, βa. The phase velocity is assumed to be equal to the group velocity. For low voltages, the resonators are much inferior to the impressed field, or to a helically conducting cylinder.

the sinusoidal field impressed at a radius a, curve A, and the flat resonators, either circular or square, B. In all cases, it is assumed that the coupling is so adjusted as to make $(v_g/v_\phi)=1$ (broad-band condition).

What sort of information can we get from the curves of Fig. 3? Suppose that we wish at 1,000 volts to make the gain of the resonators of Fig. 2 (or of a coiled waveguide of similar dimensions) as good as that for a helix with $\beta a=3$. For $\beta a=3$, the helix curve A is about 3.2 times as high as the resonator curve B. As $(E^2/\beta^2P)^{1/3}$ varies as $(v_\phi/v_g)^{1/3}$, we must adjust the coupling between resonators so as to make

$$v_g = v_\phi/(3.2)^3 = 0.031v_\phi$$

in order to make $(E^2/\beta^2P)^{1/3}$ the same for the resonators as for the helix. From (11) we see that this means that a change in frequency by a fraction 0.002 must change v_ϕ by a fraction 0.06. Ordinarily, a variation of v_ϕ of ± 0.03 would cause a very serious falling off in gain. At 3,000 Mc, the total frequency variation of 0.002 times would be 6 Mc. This is, then, a measure of the bandwidth of a series of resonators used in place of a helix for which $\beta a=3$ and adjusted to give the same gain.

4. Physical Limitations

In Section 3, the resonators were assumed to be very thin and to have walls of zero thickness. Of course, the walls must have finite thickness, and it is impractical to make the resonators extremely thin. The wall thickness and the finite transit time across the resonators both reduce $(E^2/\beta^2P)^{1/3}$.

4.1 *Effect of Wall Thickness*

Consider the resonators of Fig. 2. If d is the spacing between resonators ($1/d$ resonators per unit length),

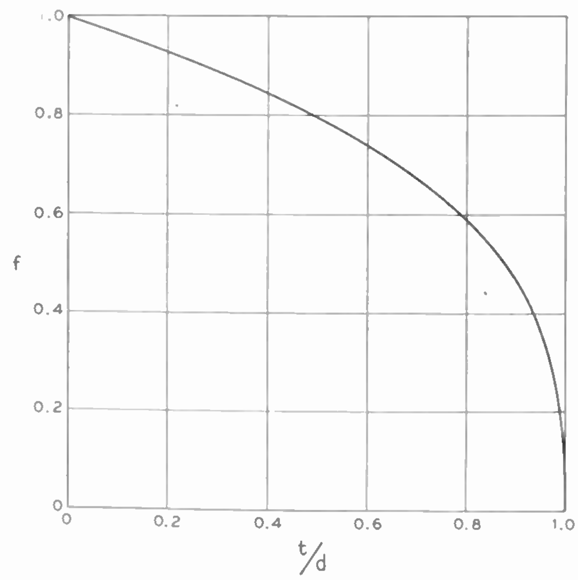

Fig. 4—In a filter-type circuit, $(E^2/\beta^2P)^{1/3}$ is reduced by a factor f as the wall thickness t is increased with respect to the resonator spacing d.

and t is the wall thickness, the stored energy will be increased over that for $t=0$ by a factor

$$(1 - t/d),$$

and $(E^2/\beta^2P)^{1/3}$ will be reduced by a factor

$$f = (1 - t/d)^{1/3}. \quad (14)$$

The factor f is plotted versus t/d in Fig. 4.

4.2 Transit Time Across Gap

As it is impractical to make the resonators infinitely thin, there will be some transit angle θ across the resonator, where

$$\theta = \beta L. \tag{15}$$

Here L is the space between resonator walls. If we assume a uniform electric field between walls, the gap factor M—that is, the ratio of peak energy gained in electron volts to peak resonator voltage, or the ratio of the magnitude of the sinusoidal field component produced to that which would be produced by the same number of infinitely thin gaps with the same voltages—will be

$$M = \frac{\sin(\theta/2)}{\theta/2}. \tag{16}$$

For a series of resonators θ long, with infinitely thin walls, $(E^2/\beta^2P)^{1/3}$ will be less than the values given by (12) and (13) by a factor $M^{2/3}$. This is plotted versus θ in Fig. 5.

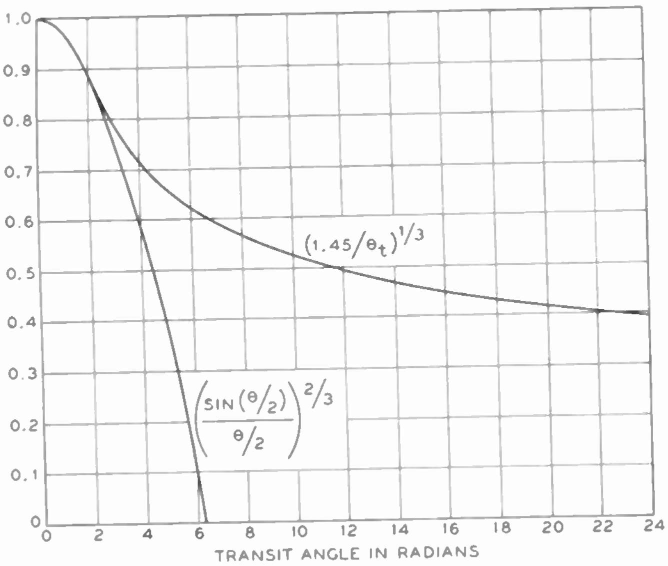

Fig. 5—In a filter-type circuit, as the transit angle θ across the gap is increased, $(E^2/\beta^2P)^{1/3}$ is reduced by a factor $[\sin(\theta/2)/(\theta/2)]^{2/3}$. If the resonators are a transit angle θ_t apart, and θ_t is less than 2.33 radians, the walls should be as thin as possible. If the resonator spacing θ_t is greater than 2.33 radians, the gap transit angle should be 2.33 radians, and $(E^2/\beta^2P)^{1/3}$ will be reduced by a factor $(1.45/\theta_t)^{1/3}$.

4.3 Fixed Gap Spacing

Suppose it is decided in advance to put only one gap in a length specified by the transit angle θ_t. How wide should the gap be made, and how much will $(E^2/\beta^2P)^{1/3}$ be reduced below the value for very thin resonators and infinitely thin walls?

Let us assume that all the stored energy is energy stored between parallel planes separated by the gap thickness, expressed in radians as θ or in distance as L.

$$\theta = \beta L$$

$$\theta_t = \beta d.$$

Here L is the gap spacing, and d is the spacing between resonators.

If we had many thin cavities with zero-thickness walls and a peak field E_0, the sum of the peak voltages across all cavities in the distance d would be

$$V_t = E_0 d.$$

The sum of the peak stored energy would be

$$W_1 = \frac{\epsilon_0}{2} E_0^2 d$$

$$W = \frac{\epsilon_0}{2} \frac{V_t^2}{d}. \tag{17}$$

If we have but one resonator with a voltage V across it, the effective voltage V_t acting on the electrons is

$$V_t = V \frac{\sin(\theta/2)}{(\theta/2)}.$$

The stored energy will be

$$W_2 = \frac{\epsilon_0}{2}\frac{V^2}{L} = \frac{\epsilon_0}{2}\frac{V^2}{d}\left(\frac{\theta_t}{\theta}\right)$$

$$W_2 = \frac{\epsilon_0}{4}\frac{(\theta/2)}{\sin^2(\theta/2)}\frac{\theta_t V_t^2}{d} \tag{18}$$

and

$$\left(\frac{W_1}{W_2}\right)^{1/3} = \left(\frac{2\sin^2(\theta/2)}{\theta_t(\theta/2)}\right)^{1/3}. \tag{19}$$

This has a maximum value for

$$\theta = 2.33 \text{ radians.} \tag{20}$$

The maximum value is

$$\left(\frac{W_1}{W_2}\right)^{1/3} = \left(\frac{1.450}{\theta_t}\right)^{1/3}. \tag{21}$$

If $\theta_t < 2.33$, it is best to make $\theta = \theta_t$. Then $(E^2/\beta^2P)^{1/3}$ is reduced by the factor $[\sin(\theta/2)/(\theta/2)]^{2/3}$, which is plotted in Fig. 5. If $\theta_t > 2.33$, it is best to make $\theta = 2.33$. Then $(E^2/\beta^2P)^{1/3}$ is reduced from the value for thin resonators with infinitely thin walls by a factor given by (21), which is plotted versus θ_t in Fig. 5.

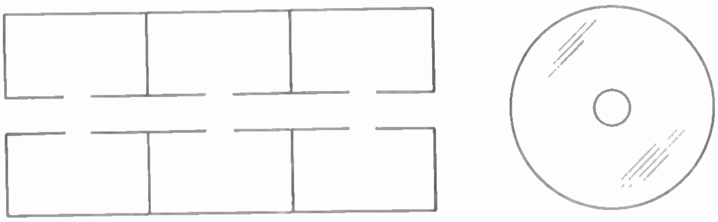

Fig. 6—When the resonators are spaced far apart, $(E^2/\beta^2P)^{1/3}$ may be made larger by the use of re-entrant resonators.

In case of wide gap separation (large θ_t), there would be some gain in using re-entrant resonators, as shown in Fig. 6.

5. Attenuation

Suppose we have a circuit made up of resonators with specified unloaded Q. The energy lost per cycle is

$$W_L = 2\pi W_s/Q. \tag{22}$$

In one cycle, however, a signal moves forward a distance L, where

$$L = v_g/f. \tag{23}$$

The fractional energy loss per unit distance, which we will call 2α, is

$$2\alpha = \frac{W_L}{W_s}\frac{1}{L} \tag{24}$$

whence

$$\alpha = \frac{\omega}{2Qv_g}. \tag{25}$$

So defined, α is the attenuation constant, and the amplitude will decay along the circuit as $\exp(-\alpha z)$.

The wavelength λ is given by

$$\lambda = v_\phi/f = 2\pi v_\phi/\omega. \tag{26}$$

The loss per wavelength in db is

$$\text{db per wavelength} = 20\log_{10}\exp(\alpha\lambda)$$

$$\text{db per wavelength} = \frac{27.3}{Q}\frac{v_\phi}{v_g}. \tag{27}$$

We see that for given values of v_ϕ and Q, decreasing the group velocity, which increased $(E^2/\beta^2P)^{1/3}$, also increases the attenuation per wavelength.

5.1 Attenuation of Circuits

For various structures, Q can be evaluated in terms of surface resistivity R, the intrinsic resistance of space, $K_0=377$ ohms, and various other parameters. For instance, Schelkunoff[5] gives for the Q of a pillbox resonator

$$Q = \frac{1.20(K_0/R)}{1 + a/h}. \tag{28}$$

Here a = the radius of the resonator, h = the height. If we express the radius in terms of the resonant wavelength λ_0 ($s = 1.2\lambda_0/\pi$), we obtain

$$Q = \frac{\pi(K_0/R)(v_\phi/c)}{n(1 + h/a)}. \tag{29}$$

Here n is the number of resonators per wavelength (assuming the walls separating the resonators to be of negligible thickness); thus,

$$n = h/\lambda = (\lambda_0/n)(v_\phi/c). \tag{30}$$

From (29) and (27) we obtain, for a series of pillbox resonators,

$$\text{db per wavelength} = 8.68(R/K_0)(c/v_g)(1 + h/a)n. \tag{31}$$

For nonmagnetic materials, surface resistance varies as the square root of the resistivity times the frequency. Table I below gives R for copper and db per wavelength for pillbox resonators for $h/a \ll 1$ (31).

TABLE I

f Mc	R Ω	(db per wavelength)/(c/v_g) Pillbox Resonators
3,000	0.0142	$3.3\times10^{-4}n$
10,000	0.0260	$6.0\times10^{-4}n$
30,000	0.0450	$10.4\times10^{-4}n$

In Section 3, a circuit made up of resonators, with a group velocity 0.031 times the phase velocity, was discussed. Suppose such a circuit were used at 1,000 volts ($c/v_\phi = 16.5$), were 40 wavelengths long, and had three copper resonators per wavelength. The total attenuation in db is given in Table II.

TABLE II

f Mc	Attenuation in db
3,000	21
10,000	38
30,000	67

Appendix I

Helically Conducting Cylinder

We have, for the helically conducting cylinder,[6]

$$(E_z^2/\rho^2P)^{1/3} = (\beta/\beta_0)^{1/3}(\gamma/\rho)^{4/3}F(\gamma a) \tag{32}$$

$$\gamma = (\beta^2 - \beta_0^2)^{1/2} \tag{33}$$

$$F(\gamma a) = \left(\frac{(\gamma a)^2}{240}\left[(I_1^2 - I_0I_2)\left(1 + \frac{I_0K_1}{I_1K_0}\right) + \left(\frac{I_0}{K_0}\right)(K_0K_2 - K_1^2)\left(1 + \frac{I_1K_0}{K_1I_0}\right)\right]\right)^{-1/3}. \tag{34}$$

Here I_0, I_1, I_2, K_0, K_1, K_2 are modified Bessel functions of the argument γa. E_z is the field on the axis. If $v_\phi \ll c$, very nearly, $\gamma = \beta$. We will make this assumption and use βa for γa in (34). Assuming this, and considering the field at the cylinder rather than on the axis, we obtain

$$(E^2/\beta^2P)^{1/3} = (c/v_\phi)^{1/3}I_0^{2/3}F(\gamma a). \tag{35}$$

Appendix II

Forced Sinusoidal Field

If $v_\phi \ll c$, the field can be very nearly represented inside the cylinder of radius a by a potential function

$$\Phi = \Phi_0\frac{I_0(\beta r)}{I_0(\beta a)}e^{-j\beta z} = \frac{E}{j\beta}\frac{I_0(\beta r)}{I_0(\beta a)}e^{-j\beta z}, \tag{36}$$

[5] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., Inc., New York, N. Y., 1943; p. 269.

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

and outside by

$$\Phi = \Phi_0 \frac{K_0(\gamma r)}{K_0(\gamma a)} e^{-j\beta z}. \tag{37}$$

Inside,

$$\frac{\partial \Phi}{\partial r} = \beta \frac{I_1(\beta r)}{I_0(\beta a)} e^{-j\beta z}\Phi_0 \tag{38}$$

$$\frac{\partial \Phi}{\partial z} = -j\beta \frac{I_0(\beta r)}{I_0(\beta a)} e^{-j\beta z}\Phi_0. \tag{39}$$

Outside,

$$\frac{\partial \Phi}{\partial r} = -\beta \frac{K_1(\beta r)}{K_0(\beta a)} e^{-j\beta z}\Phi_0 \tag{40}$$

$$\frac{\partial \Phi}{\partial z} = -j\beta \frac{K_0(\beta r)}{K_0(\beta a)} e^{-j\beta z}\Phi_0. \tag{41}$$

Taking into account the sinusoidal variation in the z direction, the average stored electric energy per unit length will be

$$W_E = \left(\frac{1}{2}\right)\left(\frac{\epsilon_0}{2}\right)\int_{r=0}^{\infty} (E_{r\max}^2 + E_{z\max}^2)(2\pi r dr). \tag{42}$$

Here $E_{r\max}$ and $E_{z\max}$ are maximum values along the z axis and are functions of r. The total electric plus magnetic stored energy will be twice this. This gives

$$W = \frac{\pi\epsilon_0(\gamma a)^2}{2\gamma^2}\left[\frac{I_0^2 - I_0 I_2}{I_0^2} + \frac{K_0K_2 - K_0^2}{K_0^2}\right]E^2$$

$$W = \frac{\pi\epsilon_0\gamma a}{\gamma^2}\left[\frac{I_1}{I_0} + \frac{K_1}{K_0}\right]E^2. \tag{43}$$

$$(E^2/\beta^2P)^{1/3} = (c/v_\phi)^{1/3}(v_\phi/v_g)^{1/3}\left[\frac{120}{\beta a\left(\frac{I_1}{I_0} + \frac{K_1}{K_0}\right)}\right]^{1/3}. \tag{44}$$

Appendix III

Pillbox Resonators

Schelkunoff[7] gives an expression for the peak electric energy stored in a pillbox resonator, which may be written as

$$0.135\pi\epsilon_0 a^2 h E^2.$$

Here a is the radius of the resonator, and h is the axial length. For a series of such resonators, the peak stored electric energy per unit length, which is also the average electric plus magnetic energy per unit length, is

$$W = 0.135\pi\epsilon_0 a^2 E^2. \tag{45}$$

For resonance,

$$a = 1.2\pi_0/\pi, \tag{46}$$

whence,

$$W = 0.0618\epsilon_0\lambda_0^2E^2, \tag{47}$$

and

$$(E^2/\beta^2P^2)^{1/3} = 5.36(v_\phi/v_g)^{1/3}(v_\phi/c)^{1/3}. \tag{48}$$

The case of square resonators is easily worked out.

[7] See p. 268 of footnote reference 5.

Appendix IV

Variation of Field over Apertures

The fields used in this paper for comparison are fields along the edge of conductors bounding an aperture, as along the surface of a helically conducting sheet. The field is, of course, weaker further from the conductors.

If β is the phase constant for a sinusoidal field component, the variation of field in the charge-free region outside of the conductors (in an aperture through resonators, or through a helix) can, in several important configurations, be expressed in terms of a parameter γ (33). For instance, in an axially symmetrical aperture of radius a,

$$E(r) = EI_0(\gamma r)/I_0(\gamma a). \tag{49}$$

Here $E(r)$ is the longitudinal field at radius r, and E is the field at a radius a (that used in the paper). Similarly, outside of an axially symmetrical system of an outside radius a,

$$E(r) = EK_0(\gamma r)/K_0(\gamma a). \tag{50}$$

In a plane-symmetrical or two-dimensional system with a slit aperture $2a$ in width,

$$E(y) = E\cosh\gamma y/\cosh\gamma a. \tag{51}$$

Here $E(y)$ is the longitudinal field a distance y from the center of the aperture, and E is the field at the center of the aperture. A distance y above a two-dimensional system

$$E(y) = E\exp -\gamma y. \tag{52}$$

Here E is the longitudinal field at the boundary of the system.

For slow waves ($v_p^2 \ll c^2$), we can replace γ in the above expressions by β.

The Effect of Pole and Zero Locations on the Transient Response of Linear Dynamic Systems*

J. H. MULLIGAN, JR.†, MEMBER, IRE

Summary—The effect of the locations of the poles and zeros of the transfer function of a linear dynamic system on the locations and the magnitudes of the maxima and minima of the transient response resulting from the application of a step-function input to the system is studied. Consideration is given to the necessary conditions for the production of a monotonic time response, expressed in terms of the pole and zero locations. In general, the results of the investigation are limited to stable low-pass systems, having only first-order poles and no poles on the $j\omega$ axis.

A method of computing the locations and magnitudes of the maxima and minima in the time response is given which allows the calculations to be made in a straightforward and efficient manner. The evaluation of the transient performance of many practical low-pass systems can be simplified considerably by the use of this method.

It is shown that, under certain conditions of pole and zero locations, the normalized time response may be well approximated by a single dominant time term. Methods of ascertaining from the pole and zero pattern whether these conditions exist are given. On the basis of the dominant-term approximation, a method is outlined for the design of pole and zero patterns to yield prescribed time-response characteristics of a certain class to step-function inputs. Constant overshoot-factor curves and charts are provided for this purpose and for rapid solution of analysis problems when applicable.

The results of this investigation are in a form which allows direct application to practical design problems in the fields of electrical networks, amplifiers, servomechanisms, and the like.

I. Introduction

Although powerful operational methods are available for the transient analysis of linear dynamic systems, the analysis of typical systems encountered in practice, such as multistage amplifiers, servomechanisms, and the like, is usually a long and tedious task, although quite straightforward. This is in contrast to the inverse problem, namely, that of transient synthesis or design, in which the method of attack is by no means as direct, or even indicated at all in some cases. As a consequence of this, a need exists for new methods for the rapid evaluation of the transient performance of a given system, and also for means of designing a system directly from prescribed requirements on the time response.

Considerable work[1-13] has been done in recent years which has provided partial satisfaction of the two needs mentioned. In most instances, this work has been given in the form of transient design information for specific circuit configurations, or generalities have been presented relating the steady-state amplitude- and phase-response characteristics of the system with the transient response to an impulse-function or step-function input.

In this paper an investigation is made of the effect that the locations of the poles and zeros of the transfer function (output/input) of a linear dynamic system have upon the location and magnitude of the maxima and minima in the transient response of the system when a step-function input is applied. Attention is also given to the necessary conditions for the production of a monotonic time-response output from the system, these conditions being expressed in terms of the pole and zero locations. The investigation is limited to stable systems whose transfer functions contain no poles on the $j\omega$ axis and no zeros at the origin, and, although the results are generally restricted to systems containing first-order poles only, under certain conditions they may be extended to systems containing multiple poles as well. Inasmuch as the results are expressed in terms of relations among pole and zero locations, the material developed is applicable in the most direct manner to systems formed by the cascading of simple structures, such as multistage nonfeedback amplifiers, for example.

II. Development and Application of Theory

1. Normalized Time Response

In order to provide sufficient generality for the work below, it will be assumed that the system under consideration has a transfer function (output/input) containing n pairs of conjugate complex poles, q real poles, g pairs of conjugate complex zeros, and m real zeros. It is also assumed that $2n+q>2g+m$, and that only first-order poles and zeros occur. It will be evident later that the restriction to first-order zeros is made only for con-

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† Formerly, Marcellus Hartley Laboratory, Columbia University, New York, N. Y.; now, Allen B. Du Mont Laboratories, Inc., Passaic, N. J.

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venience in the derivations, and that the restriction to first-order poles can be relaxed under certain conditions. The notation used is given in detail in the Appendix and is illustrated in Fig. 1.

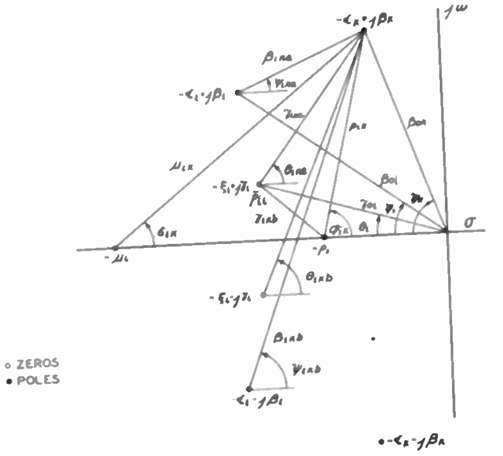


Fig. 1—Illustration of the notation employed.

If the system just described is subjected to a unit step-function input, the Laplace transform of the output time response is of the form[14]

$$F(s) = K\frac{\prod_{i=1}^{m}(s+\mu_i)\prod_{i=1}^{g}[(s+\xi_i)^2+\gamma_i^2]}{s\prod_{i=1}^{q}(s+\rho_i)\prod_{i=1}^{n}[(s+\alpha_i)^2+\beta_i^2]} \quad (1)$$

where K is a scale factor depending on the particular system considered. It will be convenient to assume that K is unity in all the work that follows.

The inverse transformation of (1) yields a time response $f(t)$ containing a constant term A_0 plus terms which are functions of time. For convenience in the comparison of the responses of systems having different values of A_0, it is desirable to consider the amplitude-normalized time response defined as

$$f(t) = A_0 f_n(t) \quad (2)$$

in which

$$A_0 = \frac{\prod_{i=1}^{m}\mu_i\prod_{i=1}^{g}\gamma_{0i}^2}{\prod_{i=1}^{q}\rho_i\prod_{i=1}^{n}\beta_{0i}^2}. \quad (3)$$

Accordingly, by inverse transformation of (1) and normalization, the following expression is obtained for the normalized time response:

$$f_n(t) = 1 + (-1)^{w+1}\sum_{k=1}^{n} M_k \csc\psi_k \epsilon^{-\alpha_k t}\sin(\beta_k t + \lambda_k + \psi_k)$$
$$+ (-1)^w \sum_{k=1}^{q}(-1)^{k+r_k}N_k\epsilon^{-\rho_k t} \quad (4)$$

in which

$$M_k = \frac{\prod_{i=1}^{m}\left|\frac{\mu_{ik}}{\mu_i}\right|\prod_{i=1}^{g}\left(\frac{\gamma_{ika}}{\gamma_{0i}}\right)\left(\frac{\gamma_{ikb}}{\gamma_{0i}}\right)}{\prod_{i=1}^{q}\left(\frac{\rho_{ik}}{\rho_i}\right)\prod_{i=1, i\neq k}^{n}\left(\frac{\beta_{ika}}{\beta_{0i}}\right)\left(\frac{\beta_{ikb}}{\beta_{0i}}\right)} \quad (5)$$

$$N_k = \frac{\prod_{i=1}^{m}\left|\frac{\mu_i-\rho_k}{\mu_i}\right|\prod_{i=1}^{g}\left(\frac{\bar\rho_{ki}}{\gamma_{0i}}\right)^2}{\prod_{i=1}^{n}\left(\frac{\rho_{ki}}{\beta_{0i}}\right)^2\prod_{i=1, i\neq k}^{q}\left|\frac{\rho_i-\rho_k}{\rho_i}\right|} \quad (6)$$

$$\lambda_k = \sum_{i=1}^{g}\theta_{ik} - \sum_{i=1, i\neq k}^{n}\psi_{ik} + \sum_{i=1}^{m}\delta_{ik} - \sum_{i=1}^{q}\phi_{ik} \quad (7)$$

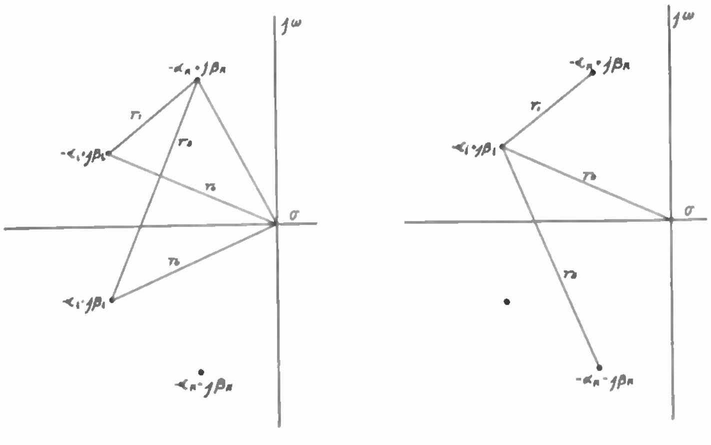
and w and r_k are the number of real zeros in the right half plane and the number of real zeros to the right of ρ_k, respectively. It is seen that the coefficients M_k and N_k are formed in the same manner. Each is equal to the quotient of the product of factors due to zeros divided by the product of factors due to poles. The factors for all poles and zeros have the same form; the typical factor for a single element (pole or zero) is equal to the ratio of two distances in the S plane; namely, the distance from the element to the pole for which the coefficient is being computed[15] divided by the distance of the element from the origin. It is now desirable to investigate this factor in greater detail.

2. Composition of Coefficients

Referring to Fig. 2(a) and (5), it is evident that the pair of complex poles at $-\alpha_i \pm j\beta_i$ contributes a multiplying factor C to the denominator of M_k of amount

$$C = \frac{r_1 r_2}{r_0^2}. \quad (8)$$

It is of interest to determine how the poles $-\alpha_i \pm j\beta_i$ must move in order that the factor C remain constant. Furthermore, it is quite desirable to consider displacement of the pole $-\alpha_i + j\beta_i$ only, inasmuch as complex elements must occur in conjugate pairs, and hence mo-



(a) (b)

Fig. 2—(a) Construction for determining the multiplying factor C, (b) Equivalent construction using a pole with positive imaginary part.

[14] The operational notation and methods of analysis used herein follow those of M. F. Gardner and J. L. Barnes, "Transients in Linear Systems," vol. I, John Wiley and Sons, Inc., New York, N. Y., 1942.

[15] This is $-\alpha_k + j\beta_k$ for M_k, and $-\rho_k$ for N_k.

tion of $-\alpha_i+j\beta_i$ causes corresponding motion of its conjugate. Consequently, in Fig. 2(b) use has been made of certain equalities of distance to refer variations in C to movement of the pole $-\alpha_i+j\beta_i$ alone. It is now necessary to find the path on which this pole must move to yield a constant value of C.

The form of Fig. 2(b) suggests solution of the locus problem by consideration of an equivalent electrostatic problem.[16] Considering the natural logarithm of C, it is found that

$$\log_\epsilon C = \log_\epsilon r_1 + \log_\epsilon r_2 - 2 \log_\epsilon r_0. \qquad (9)$$

Moreover, the electrostatic potential at the point $-\alpha_i+j\beta_i$ resulting from a line charge of $+1$ esu at the origin and line charges of $-\frac{1}{2}$ esu at $-\alpha_k \pm j\beta_k$ is[17]

$$\phi = \log_\epsilon r_1 + \log_\epsilon r_2 - 2 \log_\epsilon r_0, \qquad (10)$$

from which it is clear that the potential in the electrostatic system postulated and the factor C are related as

$$C = \epsilon^{\phi}. \qquad (11)$$

Evidently, therefore, the lines of constant C sought are the equipotentials for the system just described.

The construction of the equipotential lines for the three-line-charge system being considered is effected by the addition of the potential lines of two separate line-charge systems, each consisting of a positive and negative line charge of $\frac{1}{2}$ esu. The first system has $-\frac{1}{2}$ esu at $-\alpha_k+j\beta_k$ and $+\frac{1}{2}$ at the origin, whereas the second has $-\frac{1}{2}$ esu at $-\alpha_k-j\beta_k$ and $+\frac{1}{2}$ esu at the origin. The equipotential lines for a two-line-charge system such as is had here are circles.[18] For convenience in construction and application, it is desirable to consider that po'e and zero locations have been normalized such that β_k is equal to unity. Under such circumstances the equipotential circles for the first of the two-line-charge systems have centers at

$$-\tfrac{1}{2} \cot \psi_k(1 - \coth \phi) + j\tfrac{1}{2}(1 - \coth \phi)$$

and radii $\frac{1}{2} \csc \psi_k \operatorname{csch} \phi$. The circles for the second system have the same radii, and centers located at the conjugates of those of the first set.

Figs. 3 and 4 give equipotential lines for the upper half plane for the three-charge system for pole angles ψ_k of 40° and 60°, respectively, which have been constructed according to the method outlined above. The curves shown are for constant values of the factor C and the potential ϕ; the numerical values indicated are values of ϕ.

The above discussion has shown how one pair of complex poles affects M_k. From this it is evident how a pair

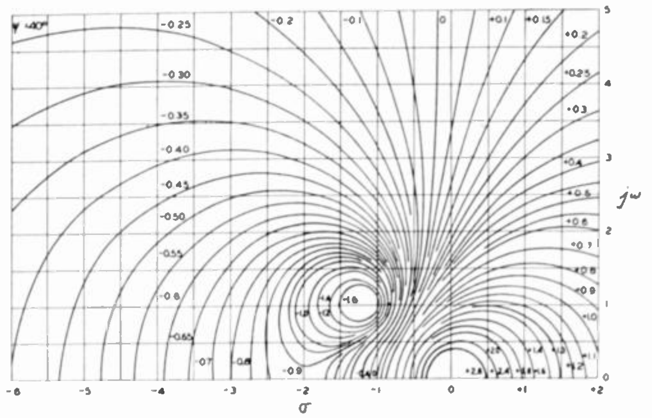


Fig. 3—Constant C (ϕ) chart for a pole angle of $\psi_k=40°$.

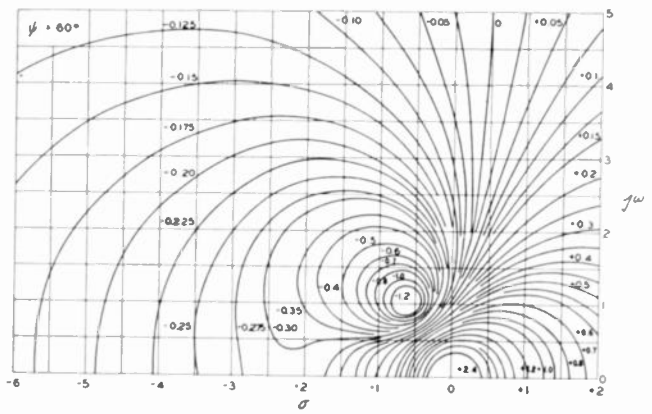


Fig. 4—Constant C (ϕ) chart for a pole angle $\psi_k=60°$.

of first-order complex zeros influences the coefficient, and a simple extension of the results will take care of the contributions of real elements and of elements of other than first order. For real poles or zeros, the logarithmic contribution to $\log_\epsilon M_k$ is one-half that shown on the curves of the figures; the logarithmic contribution of a multiple element of nth order is n times the corresponding first-order figure.

It is now apparent that if a chart, such as just discussed, is available for the angle ψ_k of the pair of poles for which the coefficient M_k is being determined, then the calculation can be performed in a very simple manner. If the pole and zero locations for elements on the σ axis and in the upper half plane are normalized and plotted on the contour chart, and if the contributions of the various elements as found from the contours are totaled, due regard being given to multiplicity and to whether pole or zero is involved (the signs shown on Figs. 3 and 4 are correct for zeros; the negative of these values are used for poles), then the total is $\log_\epsilon M_k$. By further extension of these ideas, M_k can also be expressed in terms of a potential difference in an electrostatic system constructed by placing line charges of appropriate sign and magnitude at the various pole and zero locations. This will not be discussed, however, since no application of this concept is made in the remainder of this paper.

It is quite evident that the techniques applied in the determination of M_k can be applied to the evaluation of N_k. Indeed, inasmuch as the pole $-\rho_k$ which is used in

[16] Potential theory has been applied in connection with the steady-state characteristics of amplifiers by Bradley and Hansen. See, for example, W. E. Bradley, "A theory of multistage wideband amplifier design," presented, 1947 IRE National Convention, New York, N.Y., March, 1947. Also, W. Hansen, "On maximum gain bandwidth product in amplifiers," *Jour. Appl. Phys.*, vol. 16, pp. 528–535; September, 1945.

[17] W. R. Smythe, "Static and Dynamic Electricity," first edition, McGraw-Hill Book Co., New York, N. Y., 1939; p. 62.

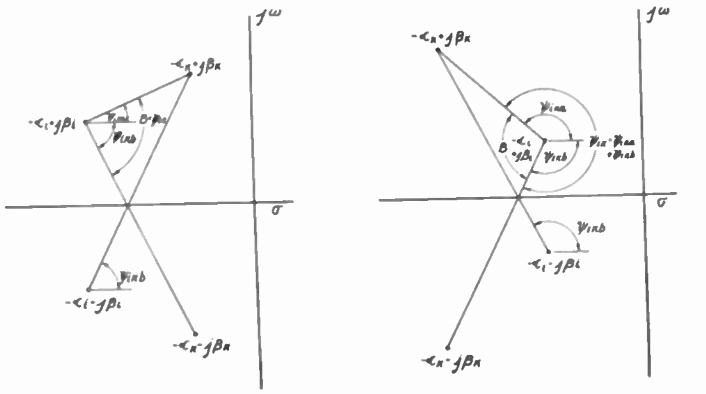
[18] See p. 75 of footnote reference 17.

he computation of the latter coefficient is located on he σ axis, the electrostatic problem used is that of a wo-line-charge system of equal and opposite charges.

3. *Composition of Phase Angles*

It is next of interest to determine the effect of pole and zero locations on the typical phase angle λ_k. An examination of Figs. 5(a) and 5(b) indicates that, if the convention is adopted that complex elements to the left of $-\alpha_k \pm j\beta_k$ produce positive angle contributions, and those to the right produce negative contributions, then the angle contribution of a pair of complex elements $-\alpha_i \pm j\beta_i$ to λ_k is equal to the angle subtended by the pair of poles $-\alpha_k \pm j\beta_k$ at the point $-\alpha_i + j\beta_i$. This is the angle B indicated in Fig. 5.

A simple calculation shows that if the pole $-\alpha_i + j\beta_i$ moves along the circle of radius $\beta_k \csc \psi_{ik}$ having its center on the σ axis at $-(\alpha_k + \beta_k \cot \psi_{ik})$, then the pair of poles $-\alpha_i \pm j\beta_i$ contributes a constant angle ψ_{ik} to λ_k. Constant-angle contours computed from this relation



(a) (b)

Fig. 5—Construction for the determination of contributions to total phase angle λ_k, (a) For an element to the left of $-\alpha_k \pm j\beta_k$; (b) For an element to the right of $-\alpha_k \pm j\beta_k$.

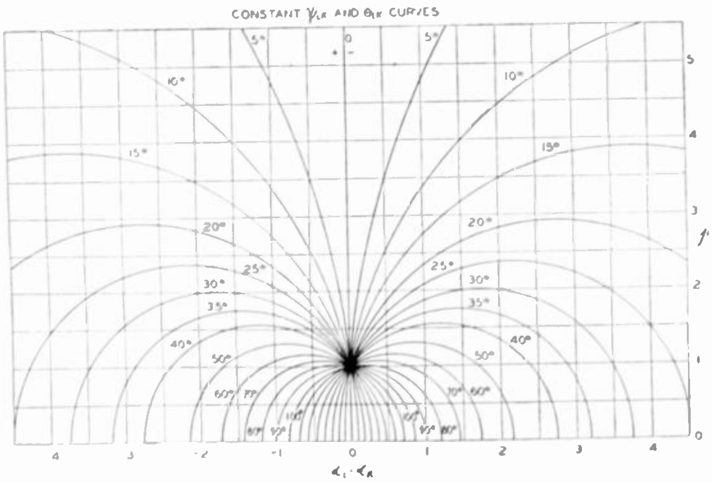


Fig. 6—Constant ψ_{ik} and θ_{ik} curves.

and taking account of the sign convention adopted are given in Fig. 6 for representative values of ψ_{ik}. The effect of complex zeros on λ_k can be determined directly from these contours, and the effect of real and multiple elements can also be evaluated without much additional effort. For example, for real poles or zeros to the left of $-\alpha_k$, the corresponding angles are one-half the values shown on the curves; for such elements located to the right of $-\alpha_k$, the angles are equal to π minus the indicated magnitudes. Similarly, the angle contributed by a multiple element of nth order is n times the corresponding first-order angle.

If the contributions of the various poles and zeros of the upper half plane are to be totaled in accordance with (7) in order to determine λ_k, then the signs shown on Fig. 6 are correct for zeros, and the negative of these should be used for poles. Evidently the curves presented furnish a very rapid means of estimating a particular λ_k, even though S plane normalization is required for the use of Fig. 6.

4. *Two Time-Response Theorems*

As a result of the condition $2n+q > 2g+m$, $f(0^+) = 0$. Therefore, the Laplace transform of the first derivative of the time response is of the form

$$F_1(s) = \frac{\prod_{i=1}^{m} (s+\mu_i) \prod_{i=1}^{g} [(s+\xi_i)^2 + \gamma_i^2]}{\prod_{i=1}^{q} (s+\rho_i) \prod_{i=1}^{n} [(s+\alpha_i)^2 + \beta_i^2]}. \tag{12}$$

Inverse transformation of (12) yields $f'(t)$, which may be written as

$$f'(t) = A_1' \epsilon^{-\alpha_1 t} g(t) \tag{13}$$

where

$$g(t) = \sin(\beta_1 t + \lambda_1) + \sum_{k=2}^{n} \frac{\beta_1}{\beta_k} m_k \epsilon^{-\alpha_k' t} \sin(\beta_k t + \lambda_k) + \sum_{k=1}^{q} (-1)^{r_k+k+1} \frac{\beta_1}{\rho_{k1}} n_k \epsilon^{-\rho_k' t} \tag{14}$$

$$A_1' = \frac{\prod_{i=1}^{m} \mu_{i1} \prod_{i=1}^{g} (\gamma_{i1a}\gamma_{i1b})}{\beta_1 \prod_{i=1}^{q} \rho_{i1} \prod_{i=2}^{n} (\beta_{i1a}\beta_{i1b})} \tag{15}$$

$$m_k = \frac{\prod_{i=1}^{m} \left(\frac{\mu_{ik}}{\mu_{i1}}\right) \prod_{i=1}^{g} \left(\frac{\gamma_{ika}}{\gamma_{i1a}}\right)\left(\frac{\gamma_{ikb}}{\gamma_{i1b}}\right)}{\prod_{i=1}^{q} \left(\frac{\rho_{ik}}{\rho_{i1}}\right) \prod_{i=2, i\neq k}^{n} \left(\frac{\beta_{ika}}{\beta_{i1a}}\right)\left(\frac{\beta_{ikb}}{\beta_{i1b}}\right)} \tag{16}$$

$$n_k = \frac{\prod_{i=1}^{m} \frac{|\mu_i - \rho_k|}{\mu_{i1}} \prod_{i=1}^{g} \left(\frac{\bar{\rho}_{ki}}{\gamma_{i1a}}\right)\left(\frac{\bar{\rho}_{ki}}{\gamma_{i1b}}\right)}{\prod_{i=2}^{n} \left(\frac{\rho_{ki}}{\beta_{i1a}}\right)\left(\frac{\rho_{ki}}{\beta_{i1b}}\right) \prod_{i=1, i\neq k}^{q} \frac{|\rho_i - \rho_k|}{\rho_{i1}}} \tag{17}$$

$$\alpha_k' = \alpha_k - \alpha_1 \tag{18}$$

$$\rho_k' = \rho_k - \alpha_1. \tag{19}$$

The definition of the function $g(t)$ is of advantage in determining the zeros of $f'(t)$, and hence the maxima and minima of $f(t)$ and $f_n(t)$. It is evident that the zeros of

$g(t)$ are those of $f'(t)$, and conversely, since the quantity $A_1'\epsilon^{-\alpha_1 t}$ is always positive.

It follows from the pole numbering that all α_k' are positive. If it is assumed that $\rho_1 > \alpha_1$, then another consequence of the numbering procedure is that all ρ_k' are positive. As a result of this, it is clear from (14) that after some value of t, all terms except the first in $g(t)$ will be less than some arbitrarily small amount, and hence $g(t)$ will exhibit zeros with the periodicity of $\sin(\beta_1 t + \lambda_1)$. From this behavior is deduced the following theorem:

THEOREM I: *The transient response to a step function input of a linear system having no real poles between the origin and the real part of the first pair of complex poles cannot be monotonic.*

It is to be emphasized that no statements have been made as to the magnitude of the maxima and minima that will occur in the time response when real poles are located outside the interval $(0, -\alpha_1)$, but rather only that such critical points will exist even though the deviations from unity in $f_n(t)$ may be small at these points under certain conditions.

It is of interest to examine next the case where one $\rho_k' = 0$, which will of necessity require that $\rho_1 = \alpha_1$. Under such conditions, $g(t)$ becomes

$$g(t) = (-1)^{r_1+2} n_1 + \sin(\beta_1 t + \lambda_1) + \sum_{k=2}^{n} \frac{\beta_1}{\beta_k} m_k \epsilon^{-\alpha_k' t} \sin(\beta_k t + \lambda_k) + \sum_{k=2}^{q} (-1)^{r_k + k + 1} \frac{\beta_1}{\rho_{k1}} n_k \epsilon^{-\rho_k' t}. \tag{20}$$

In order that the time response $f(t)$ or $f_n(t)$ not exhibit maxima nor minima, it is necessary that $f'(t)$, and hence $g(t)$, be nonnegative. It is seen from (20) that, as t becomes large, $g(t)$ can be approximated well by the first two terms. In order that $g(t)$ remain nonnegative at such time, it is necessary that r_1 be even and that n_1 be equal to or greater than unity. From these facts can be deduced the following result:

THEOREM II: *In order that the transient response to a step function input of a linear system having its first real pole at the projection of the first pair of complex poles on the real axis increase monotonically, it is necessary that the number of real zeros between the real pole and the origin be zero or even and that the quantity n_1* (equation 17) *be equal to or greater than unity.*

It is to be observed that these are necessary but not sufficient conditions for the production of a monotonic time response. A means of estimating n_1 directly from the pole and zero pattern is described in Section 5.

For the situation where $\rho_1 < \alpha_1$, it is somewhat more difficult to state in a general fashion even the necessary conditions for the production of a monotonic time response. It is noted, however, that, in order for $f'(t)$ to be positive even for large t in such cases, it is necessary that the number of real zeros between the origin and $-\rho_1$ be zero or even.

As an illustration of the use of the second theorem, the system consisting of a real pole at $-\alpha_1$ and two complex poles at $-\alpha_1 \pm j\beta_1$ will be considered. The quantity n_1 for this system is easily seen to be unity, and consequently it would be expected that no maxima nor minima exist in the time response. The normalized time-response equation for such a system is

$$f_n(t) = 1 + \csc^2 \psi_1 \epsilon^{-\beta_1 t \cot \psi_1} [\cos \psi_1 \cos(\beta_1 t + \lambda_1) - 1]. \tag{21}$$

Examination of the first derivative indicates points of inflection in the time response at $\beta_1 t = 2\pi$, 4π, etc. The normalized time response for the system having ψ_1 equal

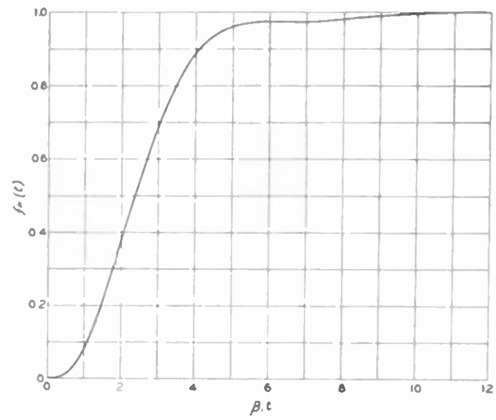


Fig. 7—Transient response to a step-function input of the illustrative system (see equation (22) of the text).

to 60° is given in Fig. 7; its appearance agrees with the theoretical predictions, showing a point of inflection in the region of $\beta_1 t = 6.28$. The normalized time response for the curve shown is

$$f_n(t) = 1 + 1.333\epsilon^{-0.5774\beta_1 t}[0.500 \cos(\beta_1 t - 60°.00) - 1]. \tag{22}$$

The above example represents the transition between time responses having maxima and minima and those not possessing such critical points. It is found by a construction to determine n_1, described in the next section, that the addition of a real zero at any point on the σ axis in the system just considered will reduce n_1 below unity, resulting in a nonmonotonic time response. This is an indication of the great importance of zeros on the time response.

5. *Dominance of the First Derivative*

The discussion of the preceding section leads to a consideration of the possibility of approximating the function $g(t)$ after a certain value of t by the first term of (14), in order to locate the zeros of $g(t)$ and $f'(t)$. This term results from the poles $-\alpha_1 \pm j\beta_1$, which are nearest the $j\omega$ axis. It has been found that such an approximation is feasible and that, in many instances, the approximation is satisfactory by the time the first maximum of $f_n(t)$ has been reached. Such an approximation, when applicable, is of considerable value in simplifying the determination of the locations and magnitudes of the maxima and minima of the time response. The condions under which the approximation is valid will now e investigated. As a result of Theorem I of the preceding section, it will be assumed in all that follows that all $\alpha_k'>0$.

Since, from (13) and (14), it is clear that for very large the zeros of $f'(t)$ and $g(t)$ will be those of sin $(\beta_1 t+\lambda_1)$, it s postulated that the zeros of $g(t)$ occur at

$$\beta_1 t = h\pi - \lambda_1 + y_h \tag{23}$$

vhere $h=b, b+1, b+2, \cdots, b$ being the integer (positive, negative, or zero) for which $b\pi-\lambda_1$ has its smallest nonnegative value. The quantity y_h is to have the nature of a correction term, and although no assumptions are made concerning its sign, the convention is to be adopted that, unless y_h has a magnitude equal to or less than $\pi/2$ radians for a particular h, no zero of $g(t)$ is considered to exist for that h. This convention is convenient under certain conditions when h is small.

Equating the left-hand side of (14) to zero, having made the substitution (23), results in the following equation from which y_h may be determined:

$$\sin y_h = (-1)^{h+1}\left\{\sum_{k=2}^{n}\frac{1}{\beta_k}m_k\epsilon^{-\alpha_k'(h\pi-\lambda_1+y_h)}\cdot\sin\left[\beta_k(h\pi-\lambda_1+y_h)+\lambda_k\right] + \sum_{k=1}^{q}(-1)^{r_k+k+1}\frac{1}{\rho_{k1}}n_k\epsilon^{-\rho_k'(h\pi-\lambda_1+y_h)}\right\} \tag{24*}$$

in which the quantities α_k', β_k, and ρ_k' are those of the normalized S plane (β_1 is equal to unity in the normalized plane). Equations in which such normalized quantities are employed will be denoted by an asterisk as in (24).

It is of interest to observe that if the relative magnitudes of $h\pi-\lambda_1$ and y_h are such that the following approximations can be made

$$\left.\begin{aligned}\epsilon^{-\alpha_k'(h\pi-\lambda_1+y_h)} &\doteq \epsilon^{-\alpha_k'(h\pi-\lambda_1)}\\ \epsilon^{-\rho_k'(h\pi-\lambda_1+y_h)} &\doteq \epsilon^{-\rho_k'(h\pi-\lambda_1)}\\ \text{and}\quad & \\ \sin\left[\beta_k(h\pi-\lambda_1+y_h)+\lambda_k\right] &\doteq \sin\left[\beta_k(h\pi-\lambda_1)+\lambda_k\right],\end{aligned}\right\} \tag{25*}$$

then (24) can be written as

$$\sin y_h \doteq (-1)^{h+1}\left\{\sum_{k=2}^{n}\frac{1}{\beta_k}m_k\epsilon^{-\alpha_k'(h\pi-\lambda_1)}\cdot\sin\left[\beta_k(h\pi-\lambda_1)+\lambda_k\right] + \sum_{k=1}^{q}(-1)^{r_k+k+1}\frac{1}{\rho_{k1}}n_k\epsilon^{-\rho_k'(h\pi-\lambda_1)}\right\}, \tag{26*}$$

which provides a direct solution for y_h. This last equation is primarily of value in estimating whether y_h is quite small, and in providing a good starting point for the systematic trial-and-error solution of (24) if y_h is not negligible.

The value of the quantity y_h is a very good measure of the approximation to $g(t)$ by the one term sin $(\beta_1 t-\lambda_1)$ at the zero of $g(t)$ represented by the value of h used. Successively small values of y_h indicate that the zeros of $g(t)$ are located very near those of the dominant term, whereas large values of y_h, i.e., near $\pi/2$, which decrease slowly with successive values of h, indicate considerable interaction due to the presence of the other terms in $g(t)$. It is clear that the values of y_h which occur for a given system are of considerable importance, and consequently attention will now be directed to what characteristics of the pole and zero pattern of a system lead to small values of this quantity.

For simplification of some of the following analysis, the quantities η_k, ζ_k, η_{kh}, and ζ_{kh} are defined as

$$\eta_k(u) = \frac{1}{\beta_k}m_k\epsilon^{-\alpha_k'u} \tag{27*}$$

$$\zeta_k(u) = \frac{1}{\rho_{k1}}n_k\epsilon^{-\rho_k'u} \tag{28*}$$

$$\eta_{kh} = \eta_k(h\pi-\lambda_1+y_h) \tag{29*}$$

$$\zeta_{kh} = \zeta_k(h\pi-\lambda_1+y_h). \tag{30*}$$

Making use of these quantities, it is now possible to express (26) in the form

$$\sin y_h \doteq (-1)^{h+1}\left\{\sum_{k=2}^{n}\eta_k(h\pi-\lambda_1)\sin\left[\beta_k(h\pi-\lambda_1)+\lambda_k\right] + \sum_{k=1}^{q}(-1)^{r_k+k+1}\zeta_k(h\pi-\lambda_1)\right\}. \tag{31*}$$

The importance of the quantities which were just introduced is now evident. If the factors $\eta_k(h\pi-\lambda_1)$ and $\zeta_k(h\pi-\lambda_1)$ are small for the value of h corresponding to the first maximum of $f_n(t)$, then the approximations (25) will be valid for this and all succeeding h, and (31) may be used for the computation of all y_h.

The actual upper limits for $\eta_k(h\pi-\lambda_1)$ and $\zeta_k(h\pi-\lambda_1)$ which will lead to a sufficiently small value of y_h such that the approximations (25) are valid depend on several factors. The limits obviously depend on the maximum value of y_h permitted; that is, on the values of h and λ_1 involved. In addition, they depend on the number of poles present (which is related to the number of terms in (31)) and on the distribution of the poles and zeros insofar as they affect the signs of the various contributions to the sum indicated by (31). For simple systems, an upper limit of 0.1 for these factors usually results in a small y_h at the first maximum of $f_n(t)$ and successive critical points; for more complicated systems, a still smaller value may be required.

The effect of β_k and ρ_{k1} on the factors η_k and ζ_k is evident from the defining equations (27) and (28). Figs. 8 and 9 have been prepared to illustrate the effect of α_k' and ρ_k' on these factors. Fig. 8 shows the variation of the factors η_k and ζ_k as a function of u for three values of α_k' and ρ_k', and two values of m_k/B_k and n_k/ρ_{k1}. Fig. 9 illustrates the variation of the factors for various α_k' and ρ_k' for a value of $u=\pi$ (this figure was chosen inas-

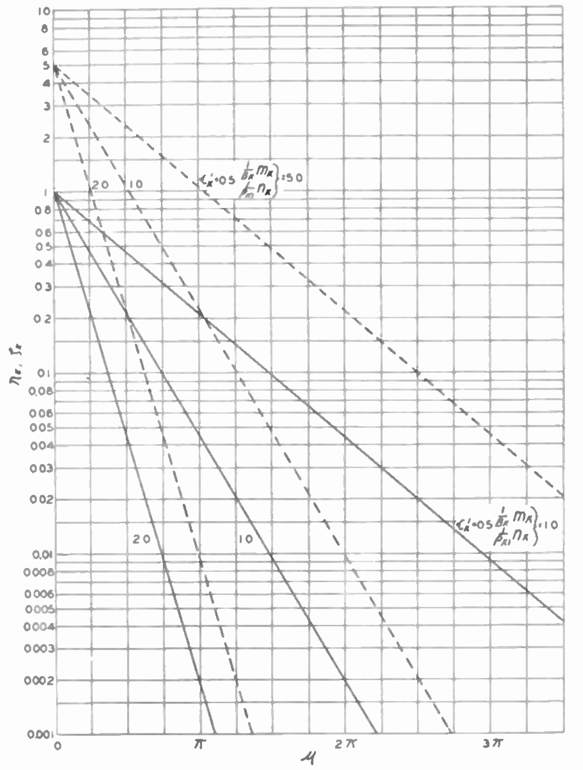


Fig. 8—The variation of the factors η_k and ζ_k with u.

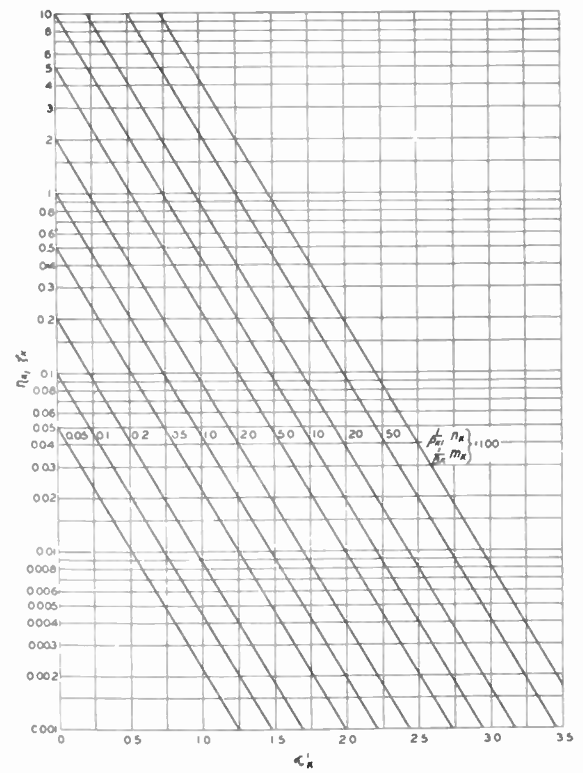
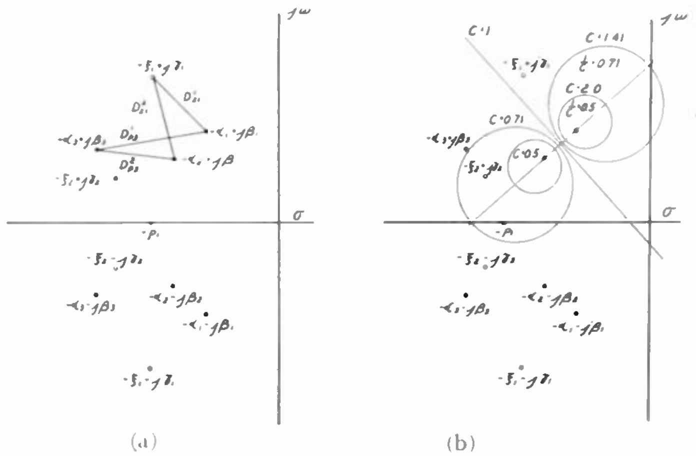


Fig. 9—The variation of the factors η_k and ζ_k with parameters α_k' and ρ_k'.

much as in most systems the first maximum is located at $\beta_1 t=\pi$ or beyond). The importance of the horizontal pole separation with respect to the dominant poles $(-\alpha_1 \pm j\beta_1)$ is immediately apparent from these figures. It is seen that the location of poles such that $\alpha_k-\alpha_1$ or $\rho_k-\alpha_1$ is less than approximately unity tends to cause relatively large values of η_k and ζ_k in the vicinity of the first maximum of $f_n(t)$, unless the quantities m_k/B_k and n_k/ρ_{k1} are quite small. The distances to which reference is made above are, of course, those in the normalized S plane.

It is next of interest to consider the magnitudes of the terms m_k and n_k. Reference to (16) and (17) indicates that m_k and n_k are both equal to the quotient of the product of factors due to zeros divided by the product of factors due to poles. For m_k, the factor for a typical element is the ratio of the distance of the element from the pole $-\alpha_k+j\beta_k$ divided by the distance of the element from the pole $-\alpha_1+j\beta_1$. For n_k, the factor is similarly defined, except that the pole at $-\rho_k$ is used instead of that at $-\alpha_k+j\beta_k$. The components of representative factors are shown for zeros (D_{z1}^1, D_{z1}^2) and poles (D_{p3}^1, D_{p3}^2) in Fig. 10(a).



(a) (b)

Fig. 10—(a) Representative components of the factors m_k and n_k. (b) Construction for the estimation of m_k.

Similarity between the defining equations of m_k and M_k and of n_k and N_k suggests application of the methods used in calculating M_k and N_k to the determination of m_k and n_k. Although several possible procedures exist for the calculation of the latter coefficients, attention will be confined here to the construction indicated in Fig. 10(b). This construction is intended to simplify the estimation of m_k (and n_k) directly from the pole and zero pattern. On the circles and straight line shown, the factor discussed above has the constant value C indicated by the figures. Evidently m_k can be estimated by multiplying together the factors for all zeros and dividing by the product of the factors for all poles except those at $-\alpha_1-j\beta_1$ and at $-\alpha_k-j\beta_k$). An estimate of n_k is made in a similar manner by the use of the pole $-\rho_k$ instead of that at $-\alpha_k+j\beta_k$.

The construction described is made in the following manner: The poles $-\alpha_k+j\beta_k$ and $-\alpha_1+j\beta_1$ are connected by a straight line. At one-half the distance d_k between the poles, a perpendicular is drawn establishing the line $C=1$. The circles $C=0.71$, $C=1.41$ have radii equal to 1.41 d_k and centers displaced an amount d_k from each pole on the extension of the line connecting them. The circles $C=0.5$, $C=2.0$ have centers on the same line at a distance of $d_k/3$ from each pole and radii of 2/3 d_k.

The construction provided yields a considerably good estimate of the quantities m_k and n_k in a simple fashion. It is of interest to note that quite an amount of information often can be obtained merely by drawing the line $\zeta=1$. Zeros to the right and poles to the left of this line contribute multiplying factors to m_k which are greater than unity, whereas zeros to the left and poles to the right produce factors less than unity.

On the basis of the preceding development, several conclusions can be drawn relating the pole and zero pattern for a system with the quality of the approximation to $g(t)$ by the term $\sin(\beta_1 t+\lambda_1)$. It is evident that, in general, conditions which lead to very small values of the factors η_k and ζ_k for the value of h corresponding to the first maximum of $f_n(t)$ will result in a very good approximation. These factors decrease rapidly with horizontal separation from the dominant pair of poles $(-\alpha_1 \pm j\beta_1)$, and they vary inversely as the quantities β_k and ρ_{k1}. Moreover, from comments concerning m_k and n_k, it follows that the location of zeros quite close to poles other than the pair $-\alpha_1 \pm j\beta_1$ and the occurrence of poles such that no dense pole concentration is formed, are conditions favorable for producing small values of η_k and ζ_k. It follows, therefore, that for systems having appreciable separation between poles, particularly along the σ axis relative to the first pair of complex poles, and having no complex poles very close to the σ axis and no complex zeros very near to the first pair of complex poles, it is quite likely that the term $\sin(\beta_1 t+\lambda_1)$ will provide a very good approximation to $g(t)$ at the time corresponding to the first maximum and beyond.

6. *Dominance of the Normalized Time Response*

Inasmuch as the locations of the maxima and minima of $f_n(t)$ have been discussed in some detail, it is now appropriate to consider the magnitude of $f_n(t)$ at these critical points. It will be convenient to consider the deviation of $f_n(t)$ from unity at these maxima and minima, rather than the absolute value of the normalized time response itself. For this purpose the quantity γ_h will be introduced, being defined as the deviation of $f_n(t)$ from unity at the critical point corresponding to the value of h used; that is,

$$\gamma_h = f_n\left(\frac{h\pi - \lambda_1 + y_h}{\beta_1}\right) - 1. \tag{32}$$

By evaluating (4) at the critical points defined by (23) and applying (32), it is found that

$$\begin{aligned}\gamma_h = {} & (-1)^{\omega+h+1} M_1 \csc\psi_1 \sin(\psi_1 + y_h)\epsilon^{-\alpha_1(h\pi-\lambda_1+y_h)} \\ & + (-1)^{\omega+1}\sum_{k=2}^{n} M_k \csc\psi_k \epsilon^{-\alpha_k(h\pi-\lambda_1+y_h)} \\ & \cdot \sin\left[\beta_k(h\pi-\lambda_1+y_h)+\lambda_k+\psi_k\right] \\ & + (-1)^{\omega}\sum_{k=1}^{q}(-1)^{r_k+k+1} N_k \epsilon^{-\rho_k(h\pi-\lambda_1+y_h)}.\end{aligned} \tag{33}^*$$

It is desirable to introduce two factors at this time which will serve as measures of the dominance of γ_h by the first term of (33). These are k_{Mkh} and k_{Nkh}, which are defined as

$$k_{Mkh} = \frac{M_1 \csc\psi_1 \sin(\psi_1 + y_h)\epsilon^{-\alpha_1(h\pi-\lambda_1+y_h)}}{M_k \csc\psi_k \epsilon^{-\alpha_k(h\pi-\lambda_1+y_h)}} \tag{34}^*$$

and

$$k_{Nkh} = \frac{M_1 \csc\psi_1 \sin(\psi_1 + y_h)\epsilon^{-\alpha_1(h\pi-\lambda_1+y_h)}}{N_k \epsilon^{-\rho_k(h\pi-\lambda_1+y_h)}}. \tag{35}^*$$

Simpler forms for these factors which are obtained by making use of certain previously defined quantities are

$$k_{Mkh} = \frac{\beta_{0k}}{\eta_{kh}} \sin\psi_1 \sin(\psi_1 + y_h) \tag{36}^*$$

and

$$k_{Nkh} = \frac{\rho_k}{\zeta_{kh}} \sin\psi_1 \sin(\psi_1 + y_h). \tag{37}^*$$

In terms of these new factors, (33) can be written as

$$\begin{aligned}\gamma_h = \gamma_{1h}\Bigg\{1 & + (-1)^h \sum_{k=2}^{n} \frac{1}{k_{Mkh}} \\ & \cdot \sin\left[\beta_k(h\pi - \lambda_1 + y_h) + \lambda_k + \psi_k\right] \\ & + (-1)^h \sum_{k=1}^{q} (-1)^{r_k+k} \frac{1}{k_{Nkh}}\Bigg\}\end{aligned} \tag{38}^*$$

where γ_{1h} is the first term in (33).

The importance of the factors k_{Mkh} and k_{Nkh} is now apparent. If these quantities are sufficiently large, then the contributions resulting from terms other than the first in (33) will cause only a small correction, resulting in a dominance of the deviation by the first term. It is seen from (36) and (37) that k_{Mkh} and k_{Nkh} will be large, providing the quantities η_{kh} and ζ_{kh} are small, unless the terms $\beta_{0k} \sin\psi_1 \sin(\psi_1+y_h)$ and $\rho_k \sin\psi_1 \sin(\psi_1+y_h)$ are appreciably less than unity. Clearly, the larger β_{0k} and ρ_k are, that is, the greater the distance of the pair of complex poles or real pole from the origin, the larger will be the factors k_{Mkh} and k_{Nkh}, other things being equal. In Section 5 it was shown, however, that large separation of the various real and complex poles from the first pair of complex poles tends to produce small values of the factors η_{kh} and ζ_{kh}, and hence a small y_h. Evidently, therefore, the conditions for dominance of the function $g(t)$ by one term and the conditions for dominance of the time response $f_n(t)$ are compatible, and thus it follows that small values of y_h will be accompanied by the dominance of $f_n(t)$ by the quantity $1-M_1 \csc\psi_1 \epsilon^{-\alpha_1 t} \sin(\beta_1 t+\lambda_1+\psi_1)$ for systems in which the pole separation is not small. The terms "large" and "small" have been used here only for convenience; quantitative conclusions depend on the actual magnitudes involved, and can be obtained directly from the curves and equations already presented.

Before presenting illustrations in which the dominant-term approximation is quite good, some attention will be given to variations in the factors k_{Mkh} and k_{Nkh} when the quantities η_{kh} and ζ_{kh} are not small for the lowest value of h. By examination of (27) to (30), (36), and (37), it is found that the factors k_{Mkh} and k_{Nkh} are equal to a constant which depends on the pole and zero pattern in a manner previously discussed, multiplied by ϵ raised to a positive power equal to α_k' or ρ_k' times the quantity $(h\pi-\lambda_1+y_h)$. Consequently, regardless of the value of the constant term, the magnitude of the quantities k_{Mkh} and k_{Nkh} increases by a factor approximately equal to $\epsilon^{\alpha_k'\pi}$ or $\epsilon^{\rho_k'\pi}$ for each unit increase in h. The implication of this is that, although the constant quantity may be very small, such that the factors are small at the value of h corresponding to the first maximum of $f_n(t)$, they will increase with successive maxima and minima, bringing the dominant term into appearance. This behavior is typical in many instances.

A somewhat different situation exists if α_2' is very small; that is, if the first and second pairs of complex poles have practically the same abscissa. For such a condition, the factor $\epsilon^{\pi\alpha_2'}$ will not be appreciably greater than unity. As a consequence, if the constant term in the factor k_{M2h} is near unity, the first term in $f_n(t)$ will not dominate the second, but instead they will both contribute more or less equal amounts to the time response. A more extreme condition occurs, however, when α_2' is very small, and where $\beta_2<\beta_1$; that is, the second pair of poles is closer to the σ axis than the first. Under such circumstances the constant may be appreciably less than unity, and the factor will increase at a slow rate. This results in dominance of $f_n(t)$, not by the term due to the first pair of poles, but by that due to the second pair. By the time that the factor k_{M2h} can increase sufficiently so that the first term is dominant, γ_h is reduced to a very small value, and the transition is concluded. An example of this behavior is given below.

Figs. 11–14 illustrate the types of time-response behavior that have been discussed. These curves are plotted to a normalized time $\beta_1 t$. In these figures the solid

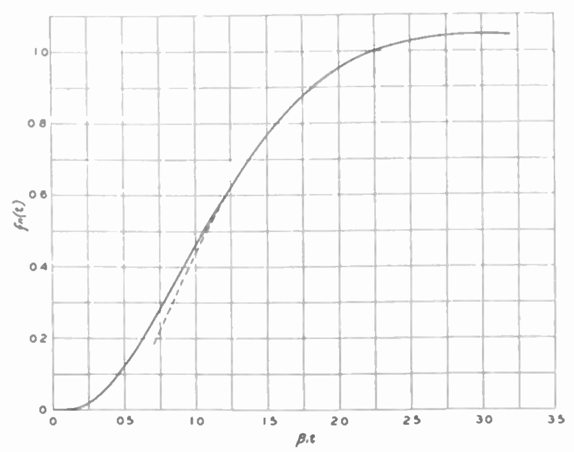


Fig. 11—A typical transient response in which the dominant-term approximation is valid. Solid curve, actual response; dash curve, dominant-term approximation.

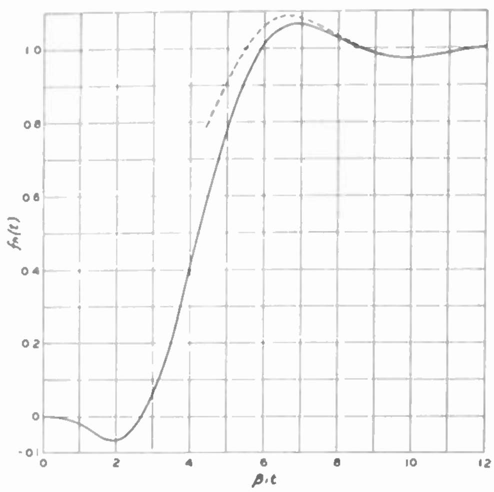


Fig. 12—A typical transient response for a system containing an all-pass section for which the dominant-term approximation is valid. Solid curve, actual response; dash curve, dominant term approximation.

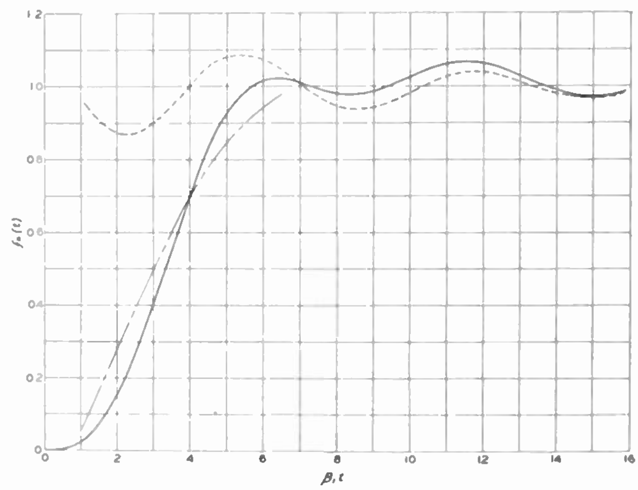


Fig. 13—Transient responses in which the dominant-term approximation is not satisfactory until after the second maximum of $f_n(t)$. Solid curve, actual response; dash curve, dominant-term approximation using damped sinusoid term due to first pair of poles; long-dash-short-dash curve, approximation using damped sinusoid due to second pair of poles.

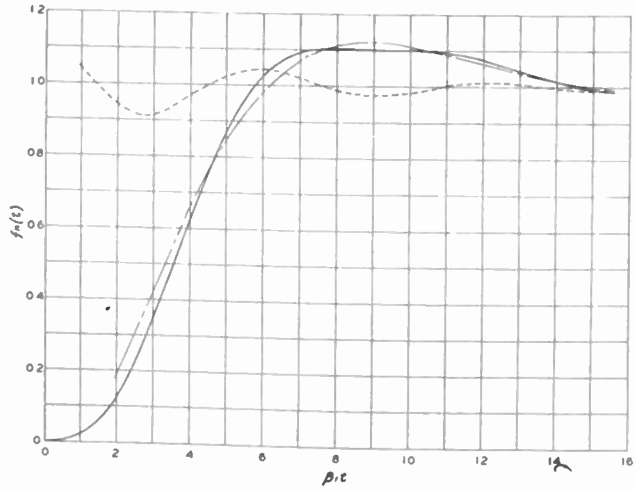


Fig 14—Transient response in which the dominant-term approximation is quite poor. Solid curve, actual response; dash curve, approximation using damped sinusoid due to the first pair of poles; short-dash–long-dash curve, approximation using damped sinusoid due to the second pair of poles.

curves are the actual time responses, while the short-dash curves are the approximation $1-M_1 \csc \psi_1 \epsilon^{-\alpha_1 t} \sin (\beta_1 t+\lambda_1+\psi_1)$, and the short-dash–long-dash curves are the approximation $1-M_2 \csc \psi_2 \epsilon^{-\alpha_2 t} \sin (\beta_2 t+\lambda_2+\psi_2)$. Figs. 11 and 12 are examples in which the dominant-term approximation to $f_n(t)$ at the first maximum and beyond is very good; Figs. 13 and 14 indicate the form of time response produced when there is interaction between the first and second damped sinusoids in $f_n(t)$. The latter two figures illustrate the form of behavior discussed immediately above. In Fig. 13 the approximation using the term due to the first pair of poles (which would ordinarily be the dominant ones) is poor at the first maximum of $f_n(t)$, but it has become quite good by the time the second maximum is reached. Fig. 14 indicates the form of time response obtained for the last situation discussed above; namely, when α_2' is quite near zero, and $\beta_2<\beta_1$. It is seen that the approximation using the term caused by the second pair of poles is very good over the whole transition, and that the ordinary "dominant-term" approximation is of little value in this instance. The normalized separation α_2' for this last example is 0.037.

7. Constant-Overshoot Charts

The simplification of analysis problems by the use of the dominant-term approximation represents a good reason in itself for the further consideration of the first term in (33) for γ_h. Of much greater importance, however, are the possibilities afforded by the use of the dominant-term theory in design or synthesis problems. In the latter class of problems, if there are more than two or three poles involved in the system, it is usually difficult to design directly for a prescribed time response to a step-function input simply because of the large number of variables involved. In many practical design problems, it is permissible to arrange the pole and zero pattern such that the dominant-term approximation will be a good one, and if such is the case, then the method of design to be outlined will be of great value.

In this section, consideration will be given to the graphical determination of the magnitude of the term γ_{1h} in the equation for the deviation γ_h at the critical points of $f_n(t)$. Charts will be constructed for this purpose, and the contours on these charts may be used for analysis or synthesis problems.

By making use of the fact that $\alpha_1=\cot \psi_1$ in the normalized S plane, it is possible to write the expression for γ_{1h} as

$$\gamma_{1h} = (-1)^{w+h+1}\epsilon^{-h\pi \cot \psi_1} \cdot M_1 \epsilon^{\lambda_1 \cot \psi_1} \frac{\sin (\psi_1 + y_h)}{\sin \psi_1} \epsilon^{-y_h \cot \psi_1}. \qquad (39)$$

The term

$$\frac{\sin (\psi_1 + y_h)}{\sin \psi_1} \epsilon^{-y_h \cot \psi_1}$$

in (39) can be considered to be due to a fictitious real pole introduced at $-(\alpha_1+\cot y_h)$. Therefore, if M_1' denotes the equivalent of M_1 and λ_1' denotes the equivalent of λ_1 in the system containing all the original poles and zeros, plus the real pole added due to y_h, then (29) can be written

$$\gamma_{1h} = (-1)^{w+h+1}\epsilon^{-h\pi \cot \psi_1} M_1' \epsilon^{\lambda_1' \cot \psi_1} \qquad (40)$$

and the natural logarithm of the magnitude of π_{1h} is

$$\log_\epsilon | \gamma_{1h} = - h\pi \cos \psi_1 + \log_\epsilon M_1' + \lambda_1' \cot \psi_1. \qquad (41)$$

The terms $\log_\epsilon M_1'$ and λ_1' cot ψ_1 are very similar in form, since each term consists of the sum of factors for all zeros minus the sum of factors due to all poles. Moreover, each pair of complex poles or complex zeros can be replaced by the member of the pair having positive imaginary part, and the contribution of the pair to $\log_\epsilon M_1'$ can be found by using charts such as are illustrated in Figs. 3 and 4, whereas the contribution of the pair to λ_1' can be found using a chart such as shown in Fig. 6.

The form of (41) therefore suggests the following construction: For a given pole angle ψ_1, let a chart of the same form as Fig. 6 be prepared on which lines of constant ψ_{ik} cot ψ_1, instead of constant ψ_{ik} be drawn. Let the coefficient chart (of the type shown in Figs. 3 and 4) for the same pole angle be combined with the constant ψ_{ik} cot ψ_1 contour chart. Then, any point on a contour of the resulting chart will contribute a constant amount to the sum $\log_\epsilon M_1' +\lambda_1'$ cot ψ_1 in (41). Such a chart is of very great value, for if the complex poles and zeros of a system having positive imaginary parts and the real poles and zeros (including the one due to y_h) are plotted on such a chart, then it is only necessary to observe the value of the curve on which each element lies, and to total these values with proper regard to sign in order to determine the quantity $\log_\epsilon M_1' +\lambda_1'$ cot ψ_1. The addition of this quantity to the term $-h\pi \cot \psi_1$ as indicated by (41) yields γ_{1h}. Moreover, it will be observed from (41) that, except for the real pole due to y_h, the quantity obtained from the chart is independent of h. Because of

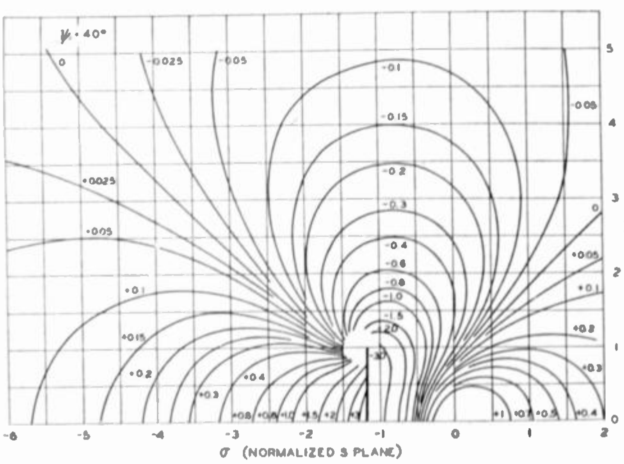


Fig. 15—Constant-overshoot-factor chart for a dominant-pole angle of 40°.

the nature of the calculations, the name "constant-overshoot-factor" contours will be given to the lines in such a chart.

Two such constant-overshoot-factor charts are shown in Figs. 15 and 16 for dominant-pole angles (ψ_1) of 40° and 60°, respectively. The signs shown on the contours are correct for zeros; the sign for poles is the negative of that shown. The magnitudes indicated are the natural logarithms of the constant-overshoot factor for a pair of complex zeros (poles). In order to determine the factor for a real pole or zero, one-half the magnitudes indicated should be used. It is important to notice that the use of such charts carries no restrictions as to multiple poles and zeros, except the obvious one that the poles $-\alpha_1 \pm j\beta_1$ must be of first order. The effect of multiple complex poles or zeros of nth order is found by using n times the factor indicated on the curves, whereas if an nth-order real pole or zero is involved, $n/2$ times the magnitude shown must be used.

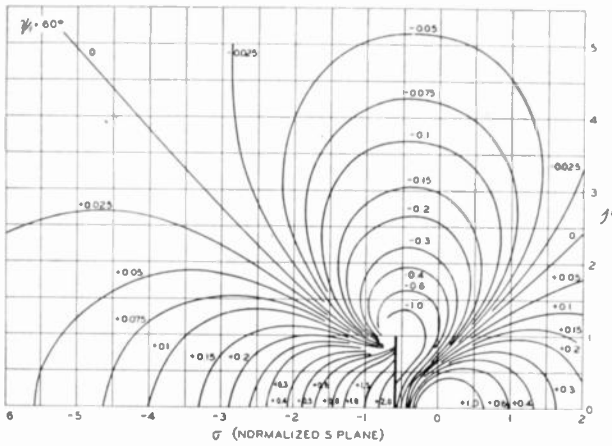


Fig. 16—Constant-overshoot-factor chart for a dominant-pole angle of 60°.

It is to be noted that the constant-overshoot-factor charts are of great value even when the dominant-term approximation is not valid. By means of the charts, γ_{1h} can be determined and γ_h can be computed by application of (38), taking due regard of terms that are not negligible. The computation of a maximum or minimum of $f_n(t)$ using (38), or an equivalent, results in an advantage in that a minimum number of terms is required for an approximation to γ_h of a desired accuracy.

Equation (40) shows that, except for a correction due to interaction (y_h), the deviation γ_{1h} depends on the quantity $\epsilon^{-h\pi}\cot\psi_1$ multiplied by a factor, due to the location of the poles and zeros in the system other than the first pair of poles. Moreover, it has been shown[19] that the deviation (γ) from unity at the first maximum of the normalized time response resulting from the application of a step-function input to a system having only one pair of complex poles is (in the notation of this paper)

$$\gamma = \epsilon^{-\pi \cot \psi_1}. \tag{42}$$

[19] See eq. (20) of footnote reference 13.

Furthermore, from the derivation of this result in the work cited, it is clear that the deviation (γ_a) at the ath zero of $f'(t)$ (not counting that at $t=0$) is

$$\gamma_a = (-1)^{a+1}\epsilon^{-a\pi \cot \psi_1}. \tag{43}$$

Returning now to (40), and replacing h by $b+a$ ($a=0, 1, 2 \cdots$), it is possible to write

$$\gamma_{1h} = (-1)^{w+b}\epsilon^{-b\pi \cot \psi_1}(-1)^{a+1}\epsilon^{-a\pi \cot \psi_1'}M_1'\epsilon^{\lambda_1' \cot \psi_1}. \tag{44}$$

The quantity b represents a reduction factor caused by the value of λ_1. For example, if $-\pi<\lambda_1<0$, b is equal to zero; if $0<\lambda_1<\pi$, b is equal to unity. Thus the result shown by (44) is that, except for a correction due to the necessary reduction of λ_1 and one to y_h, the deviation

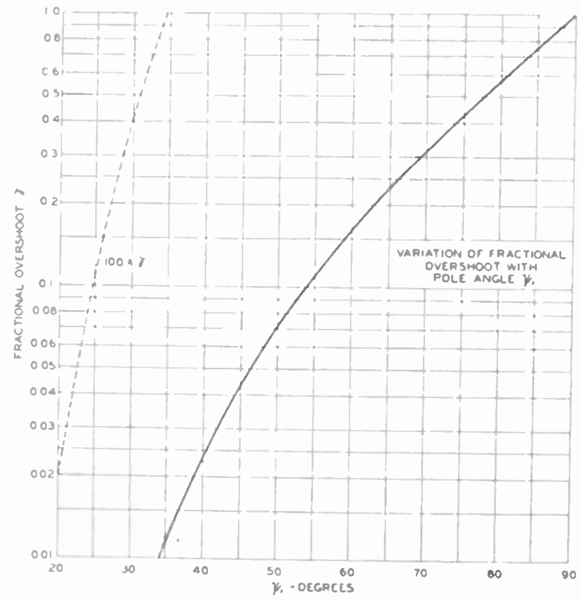


Fig. 17—Variation of fractional overshoot produced by the application of a step-function input to a system having one pair of complex poles with pole angle ψ_1.

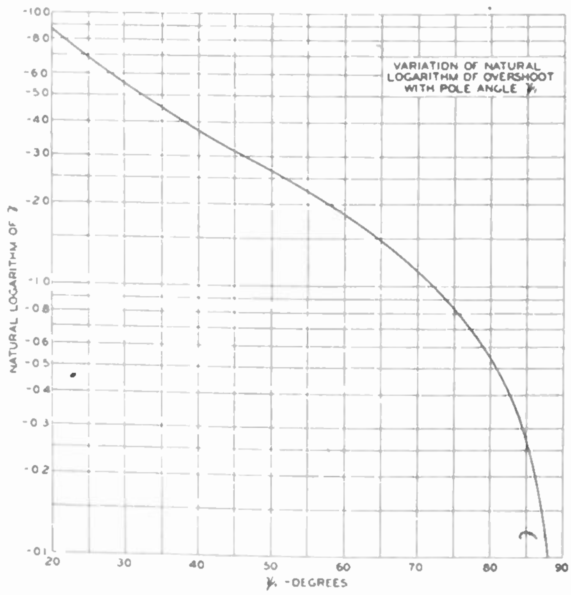


Fig. 18—Variation of the natural logarithm of the fractional overshoot produced by the application of a step-function input to a system having one pair of complex poles with pole angle ψ_1.

from unity γ_{1h} is that of the system containing only the dominant pair of poles $(-\alpha_1 \pm j\beta_1)$ multiplied by a factor, previously defined as a constant-overshoot factor, due to every other pole and zero in the system. For convenience in calculation, Figs. 17 and 18 give the variation of the factors $\epsilon^{-\pi \cot \psi_1}$ and $-\pi \cot \psi_1$ as functions of the dominant-pole angle ψ_1.

The foregoing material has not been developed merely for the treatment of specific analysis problems, but rather in order to determine the effect of pole and zero locations on the deviation γ_h (as approximated by γ_{1h}), so that a class of synthesis problems can be solved. Evidently, in the design problem, one is usually free to pick a convenient dominant-pole angle as a starting point. As a result of this, a relatively small number of charts, such as Figs. 15 and 16, would be required for design work, whereas to make feasible the use of charts for analysis, a much greater number would have to be available. The methods of analysis developed in this paper should be extremely useful in determining the locations and the magnitudes of the maxima and minima of $f_n(t)$ in a perfectly general analysis problem, even when suitable coefficient and overshoot-factor charts are not available. It is to be noted, for example, that Fig. 6 provides a rapid means of determining the angles λ_k required in various steps of the calculation, and that the figure is quite general, requiring only S-plane normalization by the appropriate β_k for its use.

In concluding this section, a problem of designing a complete pole and zero pattern to meet prescribed requirements on the time response to a step function input will be solved. Inasmuch as the purpose of the example is merely to illustrate the use of the charts and the procedures developed, a comparatively simple problem has been chosen. Evidently, many additional factors can be treated by extension of the methods shown here. For example, no attempt has been made to maximize the various quantities involved, in order to achieve an overall optimum design.

The problem to be solved is as follows: Given that four pairs of complex poles and three pairs of complex zeros are available for a given system, it is asked how they shall be located in order that the time response of the system to a step-function input have a 5 per cent overshoot at the first maximum which is to occur at a time of 1 microsecond after the application of the step function, and in order that the ratio of the deviation at the first maximum to the deviation at the first minimum be 6.

As a first step in the solution, it is recalled that the ratio between the deviations at two successive critical points is equal to $\epsilon^{\pi\alpha_1/\beta_1}$. To achieve a ratio of 6, it is necessary that $\alpha_1/\beta_1 = 0.570$. This corresponds to a pole angle of 60.25°. Inasmuch as a constant-overshoot-factor chart is available for $\psi_1 = 60°$, this will be chosen as the dominant-pole angle. This corresponds to a ratio of 6.12 instead of 6.00. From Fig. 18 it is found that the logarithm of the overshoot at the first maximum for the dominant poles of angle 60° is -1.82. The total overshoot factor that the system must possess in order to yield an overshoot of 5 per cent at the first maximum is $\log_\epsilon 0.05 = -3.00$. Therefore, the six poles and six zeros remaining must have factors totaling the difference between these two figures, namely, -1.18. Inasmuch as no other conditions have been imposed on the time response, it is convenient to divide this amount approximately equally among all the elements. Consequently, it is postulated that the poles $-\alpha_2 \pm j\beta_2$, $-\alpha_3 \pm j\beta_3$, $-\alpha_4 \pm j\beta_4$ and the zeros $-\xi_1 \pm j\gamma_1$, $-\xi_2 \pm j\gamma_2$ have overshoot factors of -0.20, whereas the zeros $-\xi_3 \pm j\gamma_3$ have an overshoot factor of -0.18. Reference is then made to Fig. 16 for the actual choice of the locations. Since the signs shown on the chart are for zeros, the poles must be put on the $+0.2$ contour. The lack of additional design constraints allows the pole and zero locations to be chosen arbitrarily, other than being located on the proper contours. From considerations which have been discussed concerning the location of poles relative to the dominant pair and the value of y_h, however, it is desirable to make α_2' greater than unity, and to locate the other two pairs of poles not too close to the second pair. No such considerations apply to the zeros.

Using the contours of Fig. 16, the poles and zeros are located as follows in the normalized S plane:

$$
\begin{aligned}
-\alpha_1 \pm j\beta_1 &= -0.5774 \pm j1.000\\
-\alpha_2 \pm j\beta_2 &= -1.750 \pm j1.000\\
-\alpha_3 \pm j\beta_3 &= -2.100 \pm j0.950\\
-\alpha_4 \pm j\beta_4 &= -2.500 \pm j0.750\\
-\xi_1 \pm j\gamma_1 &= -1.250 \pm j2.000\\
-\xi_2 \pm j\gamma_2 &= -1.270 \pm j1.700\\
-\xi_3 \pm j\gamma_3 &= -1.250 \pm j1.500.
\end{aligned}
$$

By the use of Fig. 6, λ_1 for this system is estimated to be $-75°$, and considering the pole and zero pattern in the manner described in Section 5, it is estimated that y_1 is very small. Therefore, the location of the first maximum is predicted at $\beta_1 t = \pi - \lambda_1 = 4.45$ radians. Inasmuch as this is to occur at 1 microsecond, $\beta_1 = 4.45 \times 10^{-6}$. The actual locations of the poles and zeros are then found to be

$$
\begin{aligned}
-\alpha_1 \pm j\beta_1 &= (-2.569 \pm j4.450) \times 10^6\\
&= (-5.563 \pm j8.900) \times 10^6\\
-\alpha_2 \pm j\beta_2 &= (-7.788 \pm j4.450) \times 10^6\\
&= (-5.652 \pm j7.565) \times 10^6\\
-\alpha_3 \pm j\beta_3 &= (-9.345 \pm j4.228) \times 10^6\\
&= (-5.563 \pm j6.675) \times 10^6\\
-\alpha_4 \pm j\beta_4 &= (-11.125 \pm j3.338) \times 10^6.
\end{aligned}
$$

The normalized time response for the system designed is (t in microseconds)

$$
\begin{aligned}
f_n(t) = 1 &- 0.760\epsilon^{-2.569t} \sin (4.450t - 0.2324)\\
&- 5.071\epsilon^{-7.888t} \sin (4.450t - 2.7758)
\end{aligned}
$$

$$- 15.381\epsilon^{-9.345t} \sin (4.228t - 0.8472)$$
$$- 15.109\epsilon^{-11.125t} \sin (3.338t + 1.2910). \quad (45)$$

This response is shown in Fig. 19. The data for the curve indicate that the ratio of overshoot to undershoot is 5.84. The design objectives are thus realized, inasmuch as the overshoot at the first maximum is 4.9 per cent.

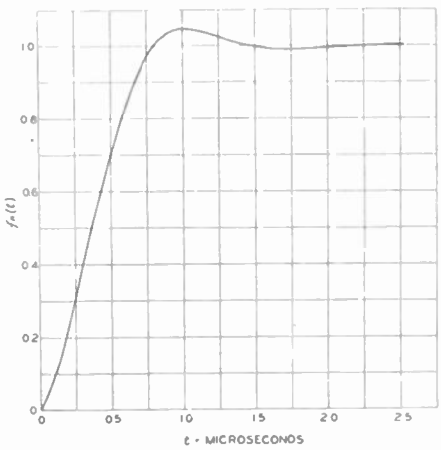


Fig. 19—Transient response to step-function input of a system obtained from the design example in the text.

8. *Constant-Overshoot-Factor Curves*

Systems having only one pair of complex poles and the remainder real poles and zeros are simpler to analyze than the systems already considered, inasmuch as only a single curve is required to show the variation of overshoot factor (with σ co-ordinate), rather than the plane previously required. Furthermore, since such an overshoot-factor curve represents the variation along the line $\omega = 0$ in the S plane, some idea is obtained of the density of the factor lines for complex poles and zeros from such a plot.

By definition, the overshoot factor for a real zero is

$$g_2 = \frac{\mu_{ik}}{\mu_i} \epsilon^{\delta_{ik} \cot \psi_1}, \quad (46)$$

and from Fig. 1, for the zero located σ units from the origin (in the normalized S plane), the natural logarithm of the factor is

$$\log_\epsilon g_2 = \tfrac{1}{2} \log_\epsilon \frac{1 + (\sigma + \cot \psi_1)^2}{\sigma^2}$$
$$+ \cot \psi_1 \cot^{-1} [- (\sigma + \cot \psi_1)]. \quad (47)^*$$

Curves of the natural logarithm of the overshoot factor (for a real zero) as a function of σ are given in Fig. 20 for pole angles ψ_1 of 20°, 40°, 60°, and 80°. Since the range of variation of δ_{ik} and ϕ_{ik} is from 0° to 180°, and the effect of complex elements is not to be considered, no advantage is gained in using the barrier on the σ axis at the abscissa of the pair of complex poles as was done in Figs. 6, 15, and 16; instead, the angles δ_{ik} and ϕ_{ik} can be allowed to vary continuously. The curves of Fig. 20 have been prepared accordingly, and as a result care must be exercised when applying these curves to systems having real zeros in the right half plane, in order that proper reduction of λ_1 be effected.

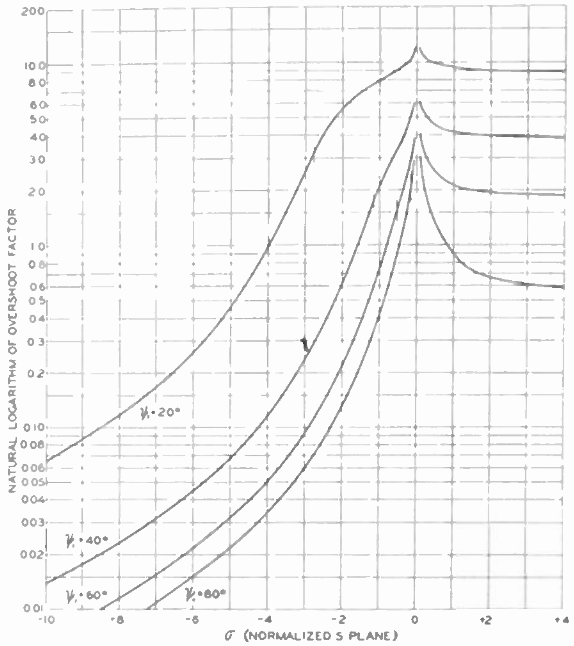


Fig. 20—Constant-overshoot-factor curves for various values of the dominant-pole angle.

The curves presented are used in the same manner as those of Figs. 15 and 16; that is, a base factor for the proper angle ψ_1 is read from Fig. 18, multiplied by the proper integer depending on the value of h being used, and then added to the factor or factors taken from Fig. 20. The magnitude of the deviation γ_{1h} is then equal to ϵ^{-x} where x is the total overshoot factor determined as above. The short vertical lines on the curves of Fig. 20 indicate the abscissa of the complex poles having pole angle ψ_1. When real poles are being considered, a negative sign must be used with the factors of Fig. 20.

III. Conclusions

From the material developed in the preceding sections, it is evident that considerable information concerning the transient response to a step-function input can be obtained merely from an examination of the pole and zero pattern of the system plotted in the S plane. By the application of Theorems I and II, it is often possible to determine whether a nonmonotonic time response exists. If it does, a qualitative idea of the shape of $f_n(t)$, that is, whether two or more terms interact, can be obtained by observing the relative separation of the poles. The location of the first maximum of $f_n(t)$ can be estimated by normalization of the pole and zero pattern and the use of Fig. 6.

The analytical procedures that have been outlined allow the determination of the locations and magnitudes of the maxima and the minima to any desired degree of approximation in a direct and relatively simple manner. By computing the approximate value of y_h for the first maximum, making use of (26), it is found which poles produce terms of importance and which, if any, can be neglected for the accuracy required. After determination of this quantity, the calculation of the factors k_{Mkh} and k_{Nkh} serves to indicate the importance of the various terms in (38) for the deviation at the critical point. Computation of overshoot and undershoot can thus be effected in an efficient, straightforward manner. It is to be noted that most of the factors required in the calculations just described can be determined graphically from the normalized pole and zero plot.

It has been shown in Sections 5 and 6 that, if sufficient separation exists between the poles and zeros in the normalized S-plane plot, then the time response can be well approximated at the first maximum and beyond by using only a single time term. Methods were developed and presented in those two sections which enable one to determine whether this behavior exists. When the dominant-term theory is applicable, computation of the location and magnitude of the maxima and minima is simplified considerably.

In addition to use in analysis problems, certain of the methods presented are of great use in the solution of design problems to yield prescribed time responses to step-function inputs. The design method illustrated depends on the use of the dominant-term approximation, and hence elements must be placed so that the approximation is valid. The constant-overshoot-factor charts developed make the solution of such problems a simple matter. These charts are also of use in problems where it is desired to correct certain characteristics of the time response.

Finally, even though only two coefficient charts and two constant-overshoot-factor charts have been constructed, they yield much qualitative information concerning the effect of poles and zeros on the coefficients and the amount of overshoot produced by a step-function input for systems having pole angles different than those for which they were computed. It is found that complex poles and complex zeros can either increase or decrease the overshoot of the dominant system, depending on the location of the elements. Complex zeros high in the plane and complex poles low in the plane both cause reduction of transient overshoot. Conversely, poles high in the plane and zeros low in the plane both cause increase of overshoot. Real poles and zeros to the left of the dominant pair cause decrease and increase of overshoot, respectively, the effect being greater the closer the real elements are to the dominant poles. In addition, in general, the overshoot is affected much more by the movements of elements in the region very near the dominant poles than by corresponding movement of elements in more removed regions of the plane. An appreciation of these ideas is often quite helpful in the modification of existing designs and the design of completely new systems.

Acknowledgment

Appreciation is expressed to the International Business Machines Corporation for a Watson Computing Laboratory Fellowship which made possible the major part of the work reported in this paper. The co-operation of the Watson Laboratory staff and the assistance of John B. Russell and John R. Ragazzini of the electrical engineering department of Columbia University are also gratefully acknowledged.

APPENDIX

Explanation of Transform and Time-Response Notation

Symbols used (illustrated in Fig. 1).

$-\alpha_k \pm j\beta_k$	Location of kth pair of complex poles
$-\xi_k \pm j\gamma_k$	Location of kth pair of complex zeros
$-\rho_k$	Location of kth real pole
$-\mu_k$	Location of kth real zero
β_{0k}	Distance from origin to $-\alpha_k \pm j\beta_k$
β_{ika}	Distance from $-\alpha_i + j\beta_i$ to $-\alpha_k + j\beta_k$
β_{ikb}	Distance from $-\alpha_i - j\beta_i$ to $-\alpha_k + j\beta_k$
γ_{0k}	Distance from origin to $-\xi_k \pm j\gamma_k$
γ_{ika}	Distance from $-\xi_i + j\gamma_i$ to $-\alpha_k + j\beta_k$
γ_{ikb}	Distance from $-\xi_i - j\gamma_i$ to $-\alpha_k + j\beta_k$
ρ_{ik}	Distance from $-\rho_i$ to $-\alpha_k + j\beta_k$
μ_{ik}	Distance from $-\mu_i$ to $-\alpha_k + j\beta_k$
$\bar{\rho}_{ik}$	Distance from $-\rho_i$ to $-\xi_k + j\gamma_k$

$$\theta_{ik} = \theta_{ika} + \theta_{ikb} = \tan^{-1}\frac{\beta_k - \gamma_i}{\xi_i - \alpha_k} + \tan^{-1}\frac{\beta_k + \gamma_i}{\xi_i - \alpha_k}$$

$$\psi_{ik} = \psi_{ika} + \psi_{ikb} = \tan^{-1}\frac{\beta_k - \beta_i}{\alpha_i - \alpha_k} + \tan^{-1}\frac{\beta_k + \beta_i}{\alpha_i - \alpha_k}$$

$$\psi_k = \tan^{-1}\frac{\beta_k}{\alpha_k}; \qquad \delta_{ik} = \tan^{-1}\frac{\beta_k}{\mu_i - \alpha_k};$$

$$\phi_{ik} = \tan^{-1}\frac{\beta_k}{\rho_i - \alpha_k}.$$

Pole and Zero Numbering

The poles and zeros of a particular type are all numbered consecutively, starting from the right-hand side of the pole pattern. For example, the real pole nearest the $j\omega$ axis is ρ_1, the next nearest ρ_2, and so on. The pair of conjugate poles nearest the $j\omega$ axis is denoted as $-\alpha_1 \pm j\beta_1$, the next nearest $-\alpha_2 \pm j\beta_2$, etc. If two or more pairs of conjugate poles or zeros have the same abscissa, then numbering for such an abscissa starts with the pair nearest the σ axis.

Contributors to the Proceedings of the I.R.E.

Henri Busignies (M'42–SM'43–F'45) was born in Sceaux, France, on December 29, 1905. He received the E.E. degree from the Paris University in 1926.

Henri Busignies

On completion of his military service, he entered the Paris Laboratories of the International Telephone and Telegraph Corp. in 1928. Until about 1940, Mr. Busignies traveled to Italy, Spain, Switzerland, England, Africa in the interests of the company. Since 1940, he has been an executive engineer, and is now technical director of Federal Telecommunication Laboratories in Nutley, N. J. His first patent was issued in 1926 and has been followed by more than one hundred others in the field of radio aids to navigation and automatic direction finding, with particular attention to air navigation. He is the chairman of the IRE Radio Aids to Navigation Committee.

In 1948 Mr. Busignies received the Presidential Certificate of Merit for his work with the National Defense Research Council during the war. He also received the Lakhovsky award of the Radio Club of France in 1926.

❖

E. L. Clark (A'31–SM'48) was born in Aberdeen, Wash., on December 4, 1901. He was graduated from the State College of Washington in 1926, with the B.S. degree in electrical engineering, and entered the test course of the General Electric Co. in September of that year. At the end of the training, he became a member of their radio engineering department. In 1930 he transferred to RCA Victor, engaged in the development of electronic equipment. He joined Philco Radio and Television Corporation in 1934, and worked on the development of radio receivers, returning to RCA Victor in 1939. During World II Mr. Clark was engaged in development on a government television project, and at present he is supervising television receiver design. He is a member of Phi Kappa Phi, Sigma Tau, and Tau Beta Pi.

E. L. Clark

❖

James H. Mulligan, Jr., whose complete biography and photograph appear on page 633 of the May, 1948 issue of the Proceedings of the I.R.E., received the Ph.D. degree from Columbia University in 1948, after which he rejoined the Allen B. Du Mont Laboratories as head of the advanced development group. He is now chief engineer of the television transmitter division.

Milton Dishal (A'41–SM'45) was born on March 20, 1918, in Philadelphia, Pa. He received the B.S. degree from Temple University in 1939, and the M.A. degree in physics in 1941. From 1939 to 1941, Mr. Dishal was a teaching fellow in physics at Temple University.

Milton Dishal

In 1941 he entered the employ of Federal Telecommunications Laboratories, where he is now a project engineer, engaged in the development of radio receivers having special characteristics.

❖

Herbert A. Glucksman (A'47) was born in New York, N. Y., on March 26, 1917. He received the B.S. degree, from the College of the City of New York in 1939. This was followed by the M.A. degree from Columbia University in 1940. Military service, starting in 1942, turned his interest toward electronics, and he served as radar countermeasures officer for the Eleventh Air Force in the air campaign against the Northern Kuriles. After leaving the service in 1947, he obtained the degree of M.A. in electronics from Harvard University, and was engaged as an electronic engineer at Cambridge Field Station, Air Matériel Command, where he is now employed.

H. A. Glucksman

❖

H. L. Kirke (M'25–SM'43–F'45) was born on January 7, 1895, in London, England. He served in the first World War in Army Signals as a radio instructor. In 1920 he joined Marconi's Wireless Company, and with others built and operated the Writtle Broadcasting Station. In 1924 he joined the British Broadcasting Corporation as development engineer, becoming head of the development department in 1925. Later, this department was called the research department.

H. L. Kirke

Mr. Kirke was chairman of the Radio Section of the Institution of Electrical Engineers from 1944 to 1945, and is now a member of Council of that body. He was awarded the C. B. E. in June, 1947.

Millett Granger Morgan (S'42–A'43–SM'48) was born in Hanover, N. H., on January 25, 1915. He attended Cornell University from 1933 to 1938 as a George W. Lefevre Scholar, and was at Stanford University from 1938 to 1940. He received the A.B. degree with honors in physics from Cornell University in 1937, the M.Sc. in engineering from Cornell in 1938; Engineer from Stanford in 1939; and Ph.D. from Stanford in 1946. He was an instructor in electrical engineering at Dartmouth College from 1940 to 1941, and spent the following year teaching at the Massachusetts Institute of Technology. From 1942–1946 he was associated with the Submarine Signal Company in Boston, Mass., as a research and development engineer.

M. G. Morgan

Dr. Morgan was a staff engineer at the United States Navy Electronics Research Group at the University of California during 1946–1947, also lecturing for this period. He has been assistant professor of electrical engineering and assistant dean of the Thayer School of Engineering, Dartmouth College, since 1947. He is a member of the American Society for Engineering Education, Sigma Xi, and past president of the Intercollegiate Ski Union.

❖

William G. Tuller (S'37–A'40–M'45–SM'48) was born in Rutherford, N. J., on September 8, 1918. From 1939 to 1944 he was a staff member of the electrical engineering department of the Massachusetts Institute of Technology, attached to the Ultra-High Frequency Laboratory, and later the Radiation Laboratory. From 1941 to 1947 he worked for the Raytheon Manufacturing Company as development engineer and consultant.

William G. Tuller

In June, 1948, Dr. Tuller received the Sc.D. degree from MIT, having been a member of the staff of the Research Laboratory of Electronics from 1945 to 1947. Since July, 1947, Dr. Tuller has been a project engineer at Melpar, Inc., Alexandria, Va., in charge of several communication and navigation system projects. He is a member of Sigma Xi.

❖

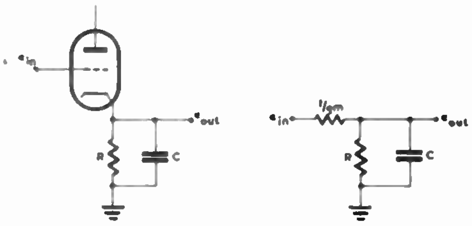
For a photograph and biography of J. R. Pierce, see page 1003 of the August, 1948, issue of the Proceedings of the I.R.E.

Correspondence

Nonlinearity in Feedback Amplifiers*

The recent paper by Mulligan and Mautner[1] forms a valuable contribution to the literature of feedback amplifiers. The theory they describe is, however, restricted to linear amplifiers, and there is a danger of its being used with disappointing results in cases where the amplifier becomes overloaded. In some cases the feedback itself contributes to such overloading.

A common example of a feedback circuit which is easily overloaded is a cathode follower with a capacitive load (Fig. 1(a)). The equivalent circuit given by linear feedback theory is shown in Fig. 1(b).



(a) (b)

Fig. 1

If R is large compared with $1/gm$, the cathode-circuit time constant given by the equivalent circuit is C/gm. The tube itself will conduct current only one way, however, so that at no time can the cathode potential fall at a rate exceeding that given by the time constant RC and the initial dc voltage on the cathode. The explanation is, of course, that the tube can be grossly overloaded by a sharp negative edge and the equivalent circuit no longer applies, even with modified values for the transconductance gm.

A similar case arises in the output stage of a video amplifier for A-scope presentation on a cathode-ray tube. The capacitance of the deflection plates shunts the anode load, producing a slow rate of rise on a positive output signal. The rate of rise cannot exceed that given by the time constant of R (the anode load)$\times C$ (the total shunt capacitance). The value of R is accordingly kept low, which restricts the maximum signal output available. It might be thought that feedback could improve the performance, and, in fact, the theory, as outlined in the paper referred to, would predict a definite improvement. Again it is necessary to look for nonlinearity in the amplifier, and it may easily be seen that, if large signals with a rate of rise exceeding that given by the time constant RC are applied to the input, overloading occurs, and the performance suffers. In fact, overloading, if it takes the form of grid current, may lead to serious distortion of another kind.

Analysis along these lines is necessary, supplementing the normal feedback analysis, to determine whether any improvement of the transient response on large signals may be expected. Careful design may inhibit the possibility of overloading due either to grid current or to cutting off, but at the expense of a much larger and more complex amplifier than might otherwise be considered necessary.

On smaller signals, considerable improvement is effected by the use of feedback and, in any event, the improved performance at low frequencies and the greater range of range of amplification would justify its use in many cases.

ADIN B. THOMAS
Radiophysics Laboratory
Council for Scientific and Industrial Research
Chippendale
N.S.W., Australia

* Received by the Institute, October 14, 194.8

[1] J. H. Mulligan, Jr., and L. Mautner, "The steady-state and transient analysis of a feedback video amplifier," PROC. I.R.E., vol. 36, pp. 595–610; May 1948.

Conformal Mapping Transformations*

In a recent letter,[1] it was concluded from Maxwell's equations that the coplanar scalar components of E and H can be mapped conformally only for very low frequencies. In general, this is so. However, in the important case of the plane wave, or TEM wave, it will be shown below that they can be mapped conformally for all frequencies.

A plane wave is one in which $E_z = H_z = 0$.

Maxwell's equations for a homogeneous, isotropic, nonconducting medium, free of electric or magnetic charge, are (in customary rationalized mks units):

$$\text{curl } E = -\mu \frac{\partial H}{\partial t} \tag{1}$$

$$\text{div. } E = 0 \tag{2}$$

$$\text{curl } H = \epsilon \frac{\partial E}{\partial t} \tag{3}$$

$$\text{div. } H = 0. \tag{4}$$

From (1),

$$\frac{\partial Ez}{\partial y} - \frac{\partial Ey}{\partial z} = -\mu \frac{\partial Hx}{\partial t} \tag{5}$$

$$\frac{\partial Ex}{\partial z} - \frac{\partial Ez}{\partial x} = -\mu \frac{\partial Hy}{\partial t} \tag{6}$$

$$\frac{\partial Ey}{\partial x} - \frac{\partial Ex}{\partial y} = -\mu \frac{\partial Hz}{\partial t}. \tag{7}$$

Since $Hz = 0$, (7) becomes

$$\frac{\partial Ex}{\partial y} = \frac{\partial Ey}{\partial x}. \tag{8}$$

From (2),

$$\frac{\partial Ex}{\partial x} + \frac{\partial Ey}{\partial y} + \frac{\partial Ez}{\partial z} = 0. \tag{9}$$

Since $Ez = 0$, (9) becomes

$$\frac{\partial Ex}{\partial x} = -\frac{\partial Ey}{\partial y}. \tag{10}$$

But (8) and (10) are the Cauchy-Riemann differential equations, which show that the function $f(w) = Ex - j\,Ey$ is analytic where $w = x + jy$. Therefore, $Ex - j\,Ey$ can be conformally mapped in the xy plane. From (3) and (4) it can be shown in a similar manner that $Hx - j\,Hy$ is analytic and can be mapped conformally.

We can tell from this how these components can vary in the x and y directions. To see how they vary with z and t, set $Ez = Hz = 0$ in (5), (6), and (3), and combine. This results in

$$\frac{\partial^2 Ey}{\partial z^2} = \mu\epsilon \frac{\partial^2 Ey}{\partial t^2},$$

which is the wave equation in one dimension. Its general solution is

$$Ey = M_1\left(x, y, t - \frac{z}{v}\right) + M_2\left(x, y, t + \frac{z}{v}\right),$$

which represents waves traveling with unchanging form in the positive and negative z directions, with a velocity

$$v = \frac{1}{\sqrt{\mu\epsilon}} = C$$

where C is the velocity of light in the medium. The same applies to Ex, Hx, and Hy.

Although we have proved that the individual field components can be mapped conformally, we are not so much interested in this as in whether the electric and magnetic potential and the stream function representing the lines of force can be mapped conformally.

Since $f(w) = Ey - jEx$ has been proved analytic,

$$F(w) = \int^w f(w)dw \text{ is analytic and } \frac{dF}{dw} = f(w)$$

$$F = V + j\phi$$

$$\frac{dF}{dw} = f(w) = \frac{\partial V}{\partial x} + j\frac{\partial \phi}{\partial x} = -j\frac{\partial V}{\partial y} + \frac{\partial \phi}{\partial y}$$

$$= \frac{\partial V}{\partial x} - j\frac{\partial V}{\partial y}$$

but

$$f(w) = Ex - jEy,$$

so

$$Ex = \frac{\partial V}{\partial x}, \qquad Ey = \frac{\partial V}{\partial y}.$$

* Received by the Institute, October 1, 1948.

[1] D. R. Rhodes, "Conformal mapping transformations," PROC. I.R.E., vol. 36, p. 632; May, 1948.

Correspondence

V is, therefore, the potential, and the mapping of $V+j\phi$ in the x, y plane is conformal, where ϕ is the stream function. V and ϕ are conjugate harmonic, and the contours of constant V (equipotentials) and the contours of constant ϕ (lines of force) are orthogonal and form curvilinear squares in the small.

By a similar procedure, it may be shown that the magnetic potential field is also conformally mapped.

The TEM wave, which we have discussed, is the principal mode of transmission on the outside of a perfect conductor, or on a multiconductor, straight, lossless transmission line, and the map is on a plane moving in the direction of propagation with the speed of the wave. This wave is impossible inside a hollow waveguide.

It is interesting to note that Willoughby[2] plots fields in complicated high-frequency transmission-line structures which are conformal maps of equipotentials and lines of force, although he does not derive his procedure directly from Maxwell's equations.

MILTON D. RUBIN
George A. Philbrick Researches, Inc.
Boston, Mass.

[2] E. O. Willoughby, "Some applications of field plotting," *Jour. IEE* (London), vol. 93, part III, pp. 275–293; July, 1946.

The Use of Conformal Transformations in Ultra-High-Frequency Transmission-Line Problems*

The question is often raised as to the validity of using conformal transformations in problems which depend upon solutions of the wave equation.[1] While it is true that, in general, conformal transformations cannot be used to obtain solutions of the wave equation except approximately in limited regions of dimensions, small compared to a wavelength, there is an important special case in which conformal transformations can be used *exactly*. Hence, applications in this case do not invalidate results based upon the use of conformal transformations.[2]

The exception to the general rule referred to is the case of guided waves which are made up of a summation of plane waves in such a way that the equiphase surface is a plane normal to the direction of propagation and that only transverse components of electric and magnetic field exist. For rectangular co-ordinates, the wave equation has the well-known form:

$$\frac{\partial^2 \bar{E}}{\partial x^2}+\frac{\partial^2 \bar{E}}{\partial y^2}+\frac{\partial^2 \bar{E}}{\partial z^2}=\mu\epsilon\frac{\partial^2 \bar{E}}{\partial t^2}$$

where the notation is the usual one. For the particular case cited, a component of the electric field may be written in the form:

$$x \text{ component of } \bar{E} = E_x(x, y)\epsilon^{j\omega(t-z/v)}.$$

The corresponding scalar wave equation for this component has the form:

$$\frac{\partial^2 E_x}{\partial x^2}+\frac{\partial^2 E_x}{\partial y^2}-\frac{\omega^2}{v^2}E_x=-\mu\epsilon\omega^2 E_x,$$

which, under the condition that $v=1/\sqrt{\mu\epsilon}$, reduces to the Laplace equation

$$\frac{\partial^2 E_x{}^2}{\partial x^2}+\frac{\partial^2 E_x{}^2}{\partial y^2}=0.$$

The same will apply to other transverse components of electric and magnetic field. Accordingly, conformal transformations may be used to investigate and design two-conductor transmission-line structures.

The assumptions implicit in the above relations are (1) inappreciable conductor losses, (2) all the energy of the transmitted wave resides in the fundamental TEM mode, (3) the cross section of the system does not vary with length, and (4) there is inappreciable radiation from the system. The above conditions do not apply to waveguides, nor to transmission lines with losses, nor to transmission lines carrying energy in their higher-order modes.

The above result is not astonishing. We know that the field configuration of a transmission line under the conditions specified is the same from zero to infinite frequency. We also know that characteristic-impedance formulas based upon low-frequency inductance and capacitance values are valid at ultra high frequencies if the above assumptions are met.

KARL SPANGENBERG
Office of Naval Research
Navy Department
Washington 25, D. C.

* Received by the Institute, January 18, 1949.

[1] D. R. Rhodes, "Conformal mapping transformations," Correspondence, PROC. I.R.E., vol. 36, p. 632; May, 1948.

[2] W. Bruce Wholey and W. Noel Eldred, "A new type slotted line section," Presented, National Electronics Conference, Chicago, Ill., November 4, 1948.

Resonances in Capacitors*

It is well known that capacitors show reversals in the sign of their reactance at sufficiently high frequencies, and that this effect can be troublesome or useful, depending on how well it can be taken into account.

In a parallel-plate capacitor, rolled or not, if both leads come out at the same end, we have an open-ended transmission line; and, utilizing the well-known equation for input current and voltage, the impedance is

$$Z=\frac{E_0}{I_0}=\sqrt{\frac{z}{y}}\coth L\sqrt{zy} \tag{1}$$

where z and y are impedance and admittance per unit length, and L is the length. If resistance and leakance are negligible, $z=j\omega l$ and $y=j\omega c$ where l and c are inductance and capacitance per unit length. Substituting and changing to trigonometric form,

$$Z=-i\sqrt{\frac{l}{c}}\cot\omega L\sqrt{lc}$$
$$=-i\sqrt{\frac{l}{c}}\cot\omega\sqrt{LC} \tag{2}$$

where L and C are total inductance and capacitance. This equation shows infinities where $\omega\sqrt{LC}=0$, π, 2π, etc., and zeroes for $\pi/2$, $3\pi/2$, $5\pi/2$, etc. Hence, antiresonant frequencies occur at

$$f_A=\frac{n}{2\sqrt{LC}} \qquad n=1, 2, 3, \text{etc.} \tag{3}$$

Resonant frequencies are at

$$f_R=\frac{n}{4\sqrt{LC}} \qquad n=1, 2, 3 \text{ etc.} \tag{4}$$

When the capacitor leads come out at opposite ends, a very different situation arises.

The currents at a given distance from one end are now not the same on the two conductors. Take the right end of the lower conductor as zero potential. Using 1 and 2 to denote the upper and lower conductor parameters, respectively,

$$\frac{de_1}{dw}=i_1z_1 \tag{5}$$

$$\frac{de_2}{dw}=i_2z_2 \tag{6}$$

$$\frac{di_1}{dw}=(e_1-e_2)y=-\frac{di_2}{dw}. \tag{7}$$

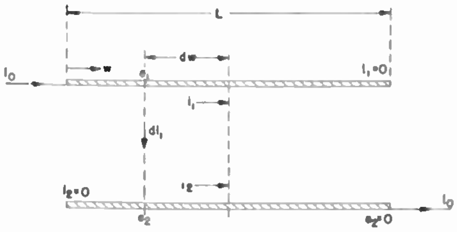


Fig. 1

If the two conductors are similar, $z_1=z_2$. Calling $e_1-e_2=e$ and $z_1+z_2=z$,

$$\frac{d^2e}{dw^2}=zye. \tag{8}$$

* Received by the Institute, December 15, 1948.

Correspondence

If, as usual, we call $\sqrt{zy}=\alpha$, (8) gives the well-known solution

$$e = c_1\epsilon^{\alpha w} + c_2\epsilon^{-\alpha w}. \tag{9}$$

Utilizing (5) and (6) and the boundary conditions shown on Fig. 1,

$$i_1 = \frac{I_0}{2\sinh\alpha L}\left[\sinh\alpha L(1+\cosh\alpha w) - \sinh\alpha w(1+\cosh\alpha L)\right]. \tag{10}$$

Integrating this and using (5), remembering that

$$z_1 = \frac{z}{2}, \tag{11}$$

$$e_1 = -\frac{I_0 z}{4\alpha\sinh\alpha L}\left[\alpha(L-w)\sinh\alpha L + \cosh(\alpha L - \alpha w) + \cosh\alpha L + \cosh\alpha w + 1\right] \tag{12}$$

Now the impedance of the capacitor is the drop between leads divided by I_0. Since the one lead is at zero potential, we have

$$Z = \frac{(e_1)_{w=0}}{I_0} = \sqrt{\frac{z}{4y}}\left(\operatorname{csch}\alpha L + \coth\alpha L + \frac{\alpha L}{2}\right). \tag{13}$$

If, as before, we neglect resistance and leakance,

$$Z = -j\sqrt{\frac{l}{4c}}\left(\csc\omega\sqrt{LC} + \cot\omega\sqrt{LC} - \frac{\omega\sqrt{LC}}{2}\right). \tag{14}$$

The antiresonant points now occur only for $\omega\sqrt{LC}=2\pi$, 4π, etc., since the odd multiples give

$$Z = j\frac{2n-1}{2}\pi\sqrt{\frac{l}{4c}}, \quad n = 1, 2, 3, \text{etc.}$$

Hence, the antiresonant frequencies are

$$f_A = \frac{n}{\sqrt{LC}} \quad n = 1, 2, 3, \text{etc.} \tag{15}$$

Resonant points occur for $Z=0$. Calling $\omega\sqrt{LC}=V$, from (14)

$$\frac{V}{2} = \frac{1+\cos V}{\sin V}. \tag{16}$$

The first seven roots are at $V=1.72$, 5.60, 6.85, 12.24, 12.87, 18.64, and 19.06, approximately, and higher-valued ones approach closely to integral multiples of 2π. Since these multiple 2π points are those of antiresonance, the impedance changes very rapidly in their vicinity and relatively slowly in between. Thus, for resonance,

$$f_R = \frac{1.72}{2\pi\sqrt{LC}}, \frac{5.60}{2\pi\sqrt{LC}}, \frac{6.85}{2\pi\sqrt{LC}}, \text{ etc.} \tag{17}$$

This peculiar phenomenon can easily cause apparently anomalous effects if one happens to be using a capacitor of this type near a 2π point, where it may be acutely frequency-sensitive.

Carl F. Muckenhoupt
Office of Naval Research
Boston Branch Office
495 Summer St.
Boston 10, Mass.

Electronics in Industry*

I have just finished reading R. R. Batcher's thought-stimulating editorial, "Electronics in Industry," in the November, 1948, issue of the Proceedings of the I.R.E.[1] It brings to my mind another and unjustified paradox.

Through some defect in our educational system, which I shall not attempt to uncover, the average undergraduate is misled into believing that his training in electronics and/or communications can *only* lead to a vocation in the communications industry. At its best, information concerning the industrial electronic engineer is colored in such a manner as to convince the student that such a life would be, indeed, dull.

Someone has aptly stated that in most schools of electrical engineering the students are divided, sharply, into two catagories "10^6 times apart." There are the power engineers who work with kilovolts, kilowatts, and kiloamperes, as against the electronic engineers who work with millivolts, milliwatts, and milliamperes. But where are the industrial electronic engineers in this picture? They are the select few who, largely by accident, master both of these schools of thought and can, therefore, make milliwatts control kilowatts.

This, then, is the paradox of electronic engineering. When graduation day comes, 99 per cent of the students rush off to the overcrowded communications industry to become "dime a dozen" engineers, while the remaining 1 per cent go into industrial electronics—leaving many positions still unfilled. The net result is that, usually, the former is underpaid for his efforts, while ordinary industry, in order to get qualified men, must pay the latter a premium wage.

Unfortunately, the communications industry is just now awaking to the fact that this self-inflicted divorce has cost them money and reputation. They are awaking to find that compressed air has taken over and it will be terribly difficult to push them out. Nearly all of the shortcomings of the contacts the communications industry makes with other forms of industry are ascribable to the inbred propaganda previously discussed.

Richard G. Devaney
2013 Sherwood Rd.
Kingsport, Tenn.

* Received by the Institute, December 15, 1948.
[1] Vol. 36, p. 1323; November, 1948.

Credit for "Low-Noise Amplifiers"*

The following is a comment on the paper by Harold Goldberg, "Some Notes on Noise Figures," which appeared in the October, 1948, issue of the Proceedings of the I.R.E.[1]

It was naturally a source of satisfaction to note the favorable comments made in this paper[1] to a low-noise circuit described in a paper of which I was one of the authors.[2] However, as the authorship of the latter paper indicates, the development of the low-noise circuit in question, to which Dr. Goldberg attaches my name, represented joint work, and exclusive credit for its development should not be assigned to one of the three authors.

Henry Wallman
Chalmers Institute of Technology
Gothenburg, Sweden
(On leave of absence from
Massachusetts Institute of Technology)

* Received by the Institute, January 10, 1949.
[1] Harold Goldberg, "Some notes on noise figures," Proc. I.R.E., vol. 36, pp. 1205–1215; October, 1948.
[2] Henry Wallman, Alan B. Mecnee, and C. P. Gadsden, "Low-noise amplifier," Proc. I.R.E., vol. 36, pp. 700–708; June, 1948.

British Television in Denmark*

In the January, 1949, issue of the Proceedings of the I.R.E., page 63, under the heading of Industrial Engineering Notes, there is published an article on "British Television in Denmark."

In reading this article we find that the only manufacturer mentioned is the Electrical and Musical Industries Company under the heading of "Film Scanner," and anyone reading this article would naturally assume that the whole of the television equipment was provided by the E.M.I.

This was not so, since the whole of the transmitter, including cameras and radio transmitters, were provided by this Company.

I should therefore be extremely grateful to you if you could arrange to publish this letter in an early issue of your journal.

B. J. Edwards
Technical Director
Pye Ltd.
Cambridge, England

* Received by the Institute, February 21, 1949.

Institute News and Radio Notes

Dayton to Hold Technical Conference

The Dayton Section of the IRE is planning a series of annual technical conferences, the first of which will be held on June 3 and 4 at the Biltmore Hotel in Dayton, Ohio.

In view of the widespread interest in that area in the field of aeronautics, the keynote of the conference will be "Development and Future Trends in Airborne Electronics." The technical program will include several papers presented by experts in their respective fields. Discussions are scheduled of aircraft communications equipment and theory, air navigation and traffic control systems, antenna developments for high-speed aircraft, air-to-air radio-frequency propagation, design trends in airborne electronic systems, and physiological aspects in the design of airborne electronic equipment.

W. L. Everitt, Dean of Engineering at the University of Illinois, will be toastmaster at a banquet to be held on June 3. The principal speaker is tentatively scheduled to be W. Stuart Symington, Secretary of the Air Force.

The program also includes displays and exhibits by various manufacturers in the electronic field, and a visit to the laboratories of the Engineering Division of the Air Matériel Command at the Wright-Patterson Air Force Base.

AIEE Holds Electron-Tube Conference in Buffalo

To provide an opportunity to discuss the application of electron tubes in the industrial field, the AIEE subcommittee on electronic control of the industrial control committee sponsored, in conjunction with the Niagara frontier section of the AIEE, a conference on the industrial application of electron tubes. The conference was held in Buffalo, N. Y., on April 11 and 12.

A program planned to cover experiences of the users and designers of equipment utilizing electron tubes, as well as by those responsible for designing and rating industrial electron tubes, was offered. Sessions were held on industrial electronic control applications, problems of electronic equipment design, and problems of the tube designer.

Conference on Ionospheric Research

A three-day conference and symposium will be held at The Pennsylvania State College on June 27, 28, and 29 to acquaint scientists in the field with the latest theoretical and experimental developments in ionospheric research. Eight to twelve papers following the theme, "Present-Day Research on Radio Wave Propagation via the Ionosphere," will be presented during the three-day session. Discussion-conference sessions will be planned after each series of papers to allow free intermingling of the latest ideas and thoughts among the participants.

This conference is being sponsored jointly by The Pennsylvania State College and the Geophysical Research Directorate of the U. S. Air Forces. Further details may be secured from A. H. Waynick, Radio Propagation Laboratories, The Pennsylvania State College, State College, Pa.

AIEE Southwest District Meeting

The Dallas and Fort Worth Sections of the AIEE held a Southwest District Meeting from April 19 to 21 in Dallas, Tex. Under the theme of "What Electrical Engineers Are Doing to Meet the Needs of the Growing Southwest," a technical program, including sessions on Transmission and Distribution, Communication, Usage of Electric Power by Heavy Industry, Relays and Protective Devices, Power Generation, Transmission and Distribution, and Nucleonics, was presented.

Section Chairmen to be Listed Bimonthly

In order to conserve much-needed space, the PROCEEDINGS henceforth will print the listing of section officers every other month, rather than each month as in the past.

Acoustical Society Meets

The theme of acoustics as the servant of mankind is being emphasized in an unusual manner by the Acoustical Society of America in its twentieth anniversary meeting, from May 5 to 7, at the Statler Hotel in New York City. Researches reported at this meeting will be grouped according to their function in relation to man, instead of by technical fields, as is usual in a scientific meeting.

Invited speakers will summarize developments for each of the sessions, which are grouped as follows: acoustics in comfort and safety, acoustics as a tool in physics, acoustics in the arts, and acoustics in communication. One day of talks and demonstration is scheduled at Murray Hill, N. J., where the Bell Telephone Laboratories will act as host.

The meeting is open to the public.

Technical Committee Notes

A meeting of the members of the **Audio Techniques Committee**, the **Video Techniques Committee**, and the **Sound Recording and Reproducing Committee** was held on March 9, during the National Convention. The problems dealt with at the meeting were those of organization, and were discussed with a view toward defining the scope of the three new committees. Following the combined meeting, each committee met separately to discuss membership and to set up subcommittees. . . . The **Circuits Committee** also met on March 9. Subcommittees preparing definitions reported on the progress of the work to date. A petition was drawn up and signed by the members present to initiate formation of a Professional Group on Circuit Theory. The committee is also making plans to sponsor a network symposium during the 1950 National Convention. . . . On March 7 the **Technical Committee on Electron Tubes and Solid-State Devices** met. A review was made of the definitions and methods of testing under preparation by the Committee. This work, now nearing completion, is to be presented as standards proposals. Under the chairmanship of J. A. Morton, the group which is preparing for the 1949 Electron Tube Conference to be held at Princeton University during the week of June 20 reported to the Committee. . . . **The Modulation Systems Committee** met at IRE Headquarters on March 14 to consider definitions prepared by its subcommittees on Pulse Code Modulation and Modulation Theory. . . . A meeting of the **Navigation Aids Committee** was held during the Convention, on March 9. A few definitions were rewritten in order to bring the Committee's standards proposal into shape for final approval and publication. The membership of the Committee for the next committee year was discussed, and three new subcommittees were formed. . . . At the **Nuclear Studies Committee's** meeting on March 8, W. R. G. Baker, the Standards Co-ordinator, addressed the members and explained the purpose and functions of the Institute's Professional Group System. At the suggestion of Chairman L. R. Hafstad, the Committee signed a petition for the formation of a Professional Group in Nuclear Science. Dr. Hafstad was elected chairman of the Group. William Geogehan analyzed responses received from the questionnaires sent to guests who attended the Conference on Nucleonics on November 29 and 30 and December 1. . . . The **Committee on Receivers** met on March 10. Two subcommittees reported: one on proposed "Italian Standards for Broadcast Receivers" and the other on Standards Definitions for Effects of Mistuning and for Downward Modulation. . . . The **Committee on Railroad and Vehicular Communications** met on March 10 at the National Convention. A few changes were suggested for the *Standards on Railroad and Vehicular Communications, Methods of Testing, 1949*. The petition for a Railroad and Vehicular Professional Group was discussed informally, and copies of it were circulated to the members of the Committee.

The following committees also held meetings during the National Convention: **Electronic Computers, Standards, Wave Propagation, Symbols, and Research.**

Left to right: R. L. Smith-Rose, Karl Spangenberg, Ralph Bown and S. L. Bailey, and R. F. Guy and Frank Stanton.

IRE Convention Draws Record Attendance

A record attendance of over 16,000 made the 1949 IRE Convention the largest in the thirty-seven years of the Institute's existence. Thousands of engineers, physicists, and technicians from thirty countries participated in the four-day conclave.

A total of 170 papers on every aspect of radio and electronics was presented at the twenty-seven sessions and six symposia which made up the technical program. Although none of the papers presented are available in reprint form, it is hoped that many of them will appear in subsequent issues of the PROCEEDINGS.

At the annual meeting which opened the Convention on Monday morning, I. S. Coggeshall, Traffic Manager of the Western Union Co., spoke on "Perpetual Youth—and the IRE," emphasizing the fact that radio is still a young man's profession. Monday afternoon the technical program opened with four sessions: Modulation Systems, Antennas and Waveguides, Instruments and Measurements, and Audio, under the respective chairmanships of Raymond F. Guy, L. J. Chu, W. R. Hewlett, and O. L. Angevine, Jr.

A Symposium on Network Theory also was held at this time. Led by J. G. Brainerd, it reviewed all of the fields of network theory in which there has been important new activity. In the evening a cocktail party provided an opportunity for the members to get together.

Sessions on Antennas, Synthesis of Passive Networks, Oscillographic Instruments and Measurements, and Electronic Computers preceded the luncheon on Tuesday to honor President Bailey. Lester C. Van Atta, Ernst Weber, T. T. Goldsmith, Jr., and E. W. Cannon headed these groups.

At the luncheon, the Junior Past President, B. E. Shackelford, acted as toastmaster to introduce Delos W. Rentzel, Civil Aeronautics Administrator, who spoke on "All-Weather Flying," and described the new devices in air navigation which will "open the way for a whole new era of aviation." After this came sessions on Television Wave Propagation, Analysis of Passive Networks, Components and Materials, and Nucleonic Instrumentation, headed by A. Earl Cullum, Jr., John B. Coleman, W. B. Anspacher, and L. J. Haworth; and a Symposium on Electronic Computing Machines, led by E. U. Condon.

Tuesday evening the Symposium on Nuclear Sciences was offered, with L. R. Hafstad as chairman. Nucleonics has received more attention from radio men this year than ever before, for from the radio laboratories are coming newer and better instruments indispensable to the production, control, and utilization of fissionable materials. Leading authorities discussed this vast new world of the subatomic, the by-products of which are expected to help recast much of our medical and other scientific knowledge.

Wednesday morning featured Television, Active Circuits, Instruments and Measurement, and Tube Design and Engineering Sessions, headed respectively by Axel G. Jensen, John R. Ragazzini, R. M. Bowie, and Dayton Ulrey. H. Busignies was chaireman of the Symposium on Radio Aids to Navigation.

More Television, Wave Propagation, Active Circuits, and Instruments and Measurements sessions were held on Wednesday afternoon, plus a session on Electron-Tube Cathodes. Robert Shelby, Charles R. Burrows, R. W. Hickman, Scott Helt, and E. A. Lederer were chairmen.

The annual IRE Banquet was held Wednesday evening. Raymond F. Guy acted as toastmaster to introduce Frank Stanton, president of the Columbia Broadcasting System, who spoke on "Television and People." At this time Ralph Bown was presented with the Medal of Honor, Claude E. Shannon with the Morris Liebmann Memorial Prize, and R. V. Pound with the Browder J. Thompson Memorial Award. Thirty-one Institute Members were given Fellow Awards.

Sessions on Relay Systems, Navigation Aids, and Electron-Tube Theory, headed by Raymond F. Guy, W. L. Everitt, and W. G. Dow, were held Thursday morning, plus a Symposium on Marketing, guided by E. H. Vogel. This symposium was an innovation designed to acquaint engineers with merchandising practicalities.

The Convention concluded with sessions on Information Transmission and Noise, Navigation Aids, Oscillators, and New Forms of Tubes, and a Symposium on Germanium and Silicon Semiconductors. These were headed by Claude E. Shannon, Lloyd T. DeVore, Robert Adler, R. R. Law, and H. A. Zahl.

Two hundred and twenty-five exhibitors, including the U. S. Air Force, Army, and Navy, representing an increase of 25 per cent over last year's 180, displayed an estimated $7,000,000 worth of the newest miracles of science, many shown for the first time.

Dissolving tumors without surgery is only one of the many potential uses of the ultrasonic fountain, a device employing a powerful crystal vibrator which focuses so much ultrasonic energy at a given point that water can be spurted two or three feet upward from a small reservoir. It can, moreover, force the mixing of nonmiscible liquids, and thus homogenize such previously uncombinable substances as oil and water.

Nucleation of a steam-cloud with dry ice produced man-made miniature snow in a cabinet containing a scale model of a U. S. Army Signal Corps Arctic installation. The Signal Corps also displayed, as part of its "miniaturization" program, a radio receiver and transmitter so small that it fits into a king-size cigarette package.

New pickups, playing 33⅓, 45, and 78 rpm records were shown, one without changing needle pressure, and another phonograph was shown that played under water. Unique television equipment was demonstrated, including rotatable antennas, a guest television system for hospitals, and a complete and operating 30-tube television receiver spread out on an upright panel to present a giant "operating blueprint" of the components and circuits of a television receiver. The world's smallest microphone not only was shown, but was used at all Convention sessions. Weighing less than one quarter of an ounce, it is supposed to have as good sound quality as any microphone in existence. Facsimile recorders receiving weather maps from Washington, Tokyo, and the Rhine Main; printed circuits for television receivers; dynamic relay testers; and new germanium triodes in dynamic applications were but a few of the innumerable scientific developments to be shown and explained to the public.

A program of women's activities was also held for members' wives. Monday afternoon, following a trip to a television program at the Du Mont studios, a "get-acquainted" group gathered at the Commodore, with Mrs. Reginald Smith-Rose acting as hostess. On Tuesday a luncheon and bridge at the Engineering Women's Club was followed by tea at IRE Headquarters, at which time the ladies were taken on tours of the building. On Wednesday the New York *Times* building was toured, after which many of the ladies attended an IRE theater-party matinee. A tour of the Brooklyn Museum of Fine Arts was offered on the final day of the Convention.

Left to right: S. L. Bailey, D. W. Rentzel, R. V. Pound and C. E. Shannon with S. L. Bailey, and B. E. Shackelford.



Industrial Engineering Notes[1]

NBS Establishes Radio Station; Develops Subminiature Devices

The National Bureau of Standards has established **radio station WWVH** on the island of Maui, Territory of Hawaii; and it is now broadcasting, on an experimental basis, continuous time and frequency standards on 5, 10, and 15 Mc. The station extends four useful technical services to the Pacific area: standard radio frequencies, time announcements, standard time intervals, and standard musical pitch.

A complete description of the NBS' work on **subminiature electronic devices,** which is being performed by the Bureau for the Navy Bureau of Aeronautics, has been published in the April issue of the *Technical News Bulletin*. Copies may be obtained from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C., for 10 cents each.

A major task of the Bureau's program is the adaptation of new techniques in the mass production of more complicated subminiature electronic devices, such as broad-band, high-gain, intermediate-frequency amplifiers for aircraft and missiles. The Bureau report describes two methods of fabrication employed in the construction of the miniaturized amplifiers. One method uses a maximum of miniature component parts based on standard design, while a second assembly uses a maximum of printed circuits.

The miniaturized amplifier is designed to have eight stagger-tuned if stages, a detector, a video amplifier, and a cathode-follower output circuit; more than 95 db gain from the if input to output detectors; manual and automatic gain control; a 60-Mc center frequency and a bandwidth of 10 Mc; and an assembly readily adaptable to mass-production methods.

Tin Conservation Urged Although Supplies Ample

Although tin allocations have been increased ten per cent by the Department of Commerce from the 1948 quota, thus making current industry supplies ample for radio and television production, government officials continue to urge tin conservation and possible substitutions, in order to avert future shortages of the medal.

Discontinuance of some 60-40 solder, and general use of 50-50 solder is recommended by the government allocation authority, which also suggests substituting aluminum for tin foil as far as possible in the manufacture of capacitors. It is believed that such substitution can be made satisfactorily for both television and radio set production in 85 to 90 per cent of capacitors.

The reduced use of tin, with consequent reduced consumption, in other industries, such as electrical appliances, motors, automotive parts, etc., has eased the tin situation, decreasing nonradio demand, and hence making the increased supply of tin available to the radio industry.

[1] The data on which these Notes are based was selected, by permission, from the Radio Manufacturers Association's "Industry Reports," issues of February 18, 25, and March 4, 11, and 18.

Calendar of COMING EVENTS

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

Twentieth Anniversary Meeting, Acoustical Society of America, New York, N. Y., May 5–7

Technical Conference on Airborne Electronics, Dayton Section, IRE, Dayton, Ohio, June 3–4

Central Section Regional Meeting, Society of Motion Picture Engineers, Toledo, Ohio, June 10

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

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

1949 IRE West Coast Convention, San Francisco, Calif., August 30–September 2

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

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

1950 IRE National Convention, New York, N.Y., March 6–9

FCC Actions

The FCC authorized two concerns, the Seismograph Service Corp., and the Frost Geophysical Corp., both of Tulsa, Okla., to **use radio equipment to locate oil deposits** in the Gulf of Mexico. The grant is on an experimental basis and permits the temporary use of frequencies in the 1,750–1,800-kc band. Copies of the decision and orders (Mimeograph Numbers 3270, 31543, and 31544) may be obtained from the Secretary of the FCC, Washington 25, D. C. . . . A request made by the Academy of Model Aeronautics to relax the **amateur regulations** in order to accommodate the radio-frequency needs of model-plane enthusiasts was denied. The Commission pointed out that the amateur regulations are subject to international agreement, and the prime purpose of these licenses is to develop radio techniques. The FCC's proposed Citizens Radio Service in the 460–470-Mc band, the Commission suggested, may prove useful for the type of communication needed in model-aircraft operation. . . . The FCC also granted **certificates of type approval for arc-stabilizer units** employing radio-frequency energy as used in the "Inert Gas Arc Welding" process to the Glenn-Roberts Co., of Indianapolis, and the National Cylinder Gas Co., of Chicago. The new equipment utilizes a tube oscillator. . . . The FCC has issued its first experimental **authorization to construct a nonmilitary shore-based radar station** for installation and operation at a major U. S. harbor. The city of Long Beach, Calif., requested the authority to study the value of shore-based radar in association with radiotelephone communication as an aid in the movement of ships in periods of reduced visibility.

New FM Stations

A total of 755 FM stations are on the air, including 31 noncommercial educational outlets. New FM stations beginning operations recently are WGAU-FM, Athens, Ga.; KPFA, Berkeley, Calif.; WDUN-FM, Gainesville, and WDBO-FM, Orlando, Fla.; WMBO-FM, Auburn, N. Y.; and WTWO, Dayton, Ohio.

Television News

Four representative plans for television station allocations, studied from two angles, one employing phase synchronization on the present vhf band only, and the other phase synchronization on vhf and uhf, were presented to the Federal Communications Commission by the Joint Technical Advisory Committee of the IRE and the RMA Engineering Department.

The first plan is based on a minimum of two stations in each of the largest 140 metropolitan centers, the second on a minimum of three stations, the third on a minimum of four stations, and the fourth on a minimum of five stations. Any larger minimum number of outlets would require more channels than the 69 available between 675 and 890 Mc.

In each of these cases there was adopted a co-channel minimum separation of 150 miles, and an adjacent-channel minimum separation of 75 miles for the vhf stations. The JTAC believes that it is not practical to assign vhf channels only to primary cities, and uhf channels only to secondary cities. As more channels are made available for television, the demand for additional channels in the large metropolitan centers must be accomodated. Therefore, of necessity, both vhf and uhf channels will have to be employed in these cities. To do otherwise in other cities would require those cities to wait for television service for a period of one to three years, even though a uhf channel or channels were currently available for allocation to it.

The National Broadcasting Co. and the Radio Corporation of America have both filed applications with the FCC for permission to construct **experimental uhf television stations.** NBC proposes to operate an experimental uhf television station in the area of Bridgeport, Conn. Planned to operate on 529–535 Mc, it would continue the uhf television experiments which NBC has been conducting in Washington.

RCA asked the FCC for permission to construct an experimental television broadcasting station at Princeton, N. J., to be operated on 846–854 Mc with 100 watts power. The purpose of the station is to obtain propagation data which compares vertical and horizontal polarization with respect to shadows and multipath phenomena in the uhf region under summer and winter conditions.

The American Telephone and Telegraph Co. and a Bell System Co. have filed applications with the FCC to **extend relay facilities** for television programs. The AT&T has requested construction permits for two experimental radio stations to extend its radio-relay system from Boston, Mass., to Providence, R. I. The Pacific Telephone and Telegraph Co. has asked permission to establish a microwave radio-relay system between Los Angeles and San Francisco, Calif.

Officials of the District of Columbia and of Maryland have issued orders which **prohibit television receivers in automobiles** where the set can be viewed by the driver while the car is in motion. The measure does not ban television sets that can be viewed by passengers only.

A **public-relations program** to give the public, also government, trade, and other interests, complete and accurate information regarding television broadcasting service and receiving sets is being undertaken by the Radio Manufacturers Association.

By the end of March, 1949, **60 television stations were on the air** in over 30 cities, and, when all the 65 stations with authorized construction permits begin broadcasting, a total of 71 cities will have one or more television stations. In addition, there are 322 applications pending for television stations in more than 94 other cities. Of television licensees, permittees, and applications, approximately three-quarters are affiliated with AM or FM stations.

Stations are operating in the following cities, as of the first of this year: Los Angeles, San Francisco, Calif.; New Haven, Conn.; Washington, D. C.; Atlanta, Ga.; Chicago, Ill.; Louisville, Ky.; New Orleans, La.; Baltimore, Md.; Boston Mass.; Detroit, Mich.; Minneapolis, Minn.; St. Louis, Mo.; Newark, N. J.; Albuquerque, N. M.; Buffalo, N. Y., Schenectady, Syracuse, N. Y.; Cincinnati, Cleveland, Toledo, Ohio; Erie, Philadelphia, Pittsburgh, Pa.; Memphis, Tenn.; Fort Worth, Tex.; Salt Lake City, Utah; Richmond, Va.; Seattle, Wash.; and Milwaukee, Wis.

Radio Set Production Declines as Television Rises

The January production of radio sets declined sharply, while the output of television sets decreased moderately in a seasonal post-holiday dip. Total RMA set production was 830,871, the lowest total since July, 1948. Production of all types of radio and television receivers in February dropped even below the January output, with a total of 716,538 units. AM and FM radio production has been declining since last fall and last winter, owing to the increasing public interest in television.

Television set shipments during the fourth quarter of 1948 increased 88 per cent over those of the third quarter.

Sets have been shipped to 42 states plus the District of Columbia. Manufacturers' radio and television set sales rose considerably in 1948, breaking all previous annual records. Sales of both types of sets exceeded $750,000,000, as compared with approximately $700,000,000 in 1947, in spite of the 20 per cent decline in radio receiver sales. Almost a third of set manufacturers' dollar volume in 1948 was of television receivers, and by December nearly half of the dollar sales came from television.

U. S. imports of radio apparatus, including receivers, parts, tubes, and cathode-ray tubes, have been increasing steadily. Imports for 1948 totalled $645,493, as compared to $295,189 in 1947. The largest shipments came from the Netherlands, and consisted chiefly of cathode-ray tubes.

Canadian Radio News

The Canadian Government has issued a decree removing a number of products, including radio receivers from the list of commodities which require export permits.

Radio set sales in Canada have been steadily declining. Sales of radio receivers during November, 1948, totalled 79,426, valued at $6,539,102, compared with 114,933 sets valued at $8,618,094 sold in November, 1947. December set sales amounted to 80,447 units valued at $6,242,389, compared with 86,946 sets valued at $8,838,545 in December, 1947. Total sales in 1948 numbered 596,467 receivers valued at $49,351,338, against 836,419 units valued at $60,399,221 in 1947.

Sets imported into Canada during November numbered 179 valued at $20,457, and brought the total for the year to 2,879 sets valued at $307,521. Exported receivers in November totalled 2,722, valued at $81,739. A total of 27,815 sets valued at $893,388 were exported by Canadian manufacturers in 1948.

Production of receiving tubes amounted to 272,697, valued at $130,674, in November, and 3,748,294 during the eleven months. Imported receiving tubes numbered 43,194 and 866,547 for November and the eleven months of 1948, respectively. Radio tube parts imported totalled 19,794 units in November, and 249,293 for the first eleven months of last year.

Radio and Television Abroad

In August, 1950, the Danish Government is sponsoring a radio exhibition at Copenhagen, in connection with the celebration of the State Radio's forthcoming twenty-fifth anniversary. As of January 1, 1949, there were 1,163,272 licensed radio receivers in **Denmark**, an increase of nearly 15 per cent over the total of March 15, 1948. . . . An International Television Congress and an International Television Fair will be held simultaneously at Milan, **Italy**, next fall. The National Research Council of the Italian Government will organize the congress and the National Association of Electrotechnical Industries will act as hosts for the fair. Foreign television experts will be invited to serve on the executive committee, and the U. S. television industry will be asked to participate via private channels. The U. S. Government is also expected to take part in the congress. . . . The Radio Industry Council of **Great Britain** recently announced the adoption by leading manufacturers of television standards to apply on television equipment for export to the continent. The manufacturers agreed to standardize equipment for 625-line pictures with positive modulation for the visual signal, in contrast to the negative modulation used in the United States. Other standards adopted in England include 25 frames per second interlaced two to one, and vestigial sideband operation and six megacycle channel width. Television receiver licenses in force in Great Britain at the end of November, 1948, totalled 80,850. Radio licenses numbered 11,343,320. . . . Manufacture of radio sets in **Austria** during 1948 totalled 70,000 receivers, less than ten per cent of which were exported, although before the war exports accounted for more than a third of Austrian radio production. . . . In **Japan** radio receiver production during the first ten months of 1948 totalled 1,392,403. Transmitter production amounted to 1,746 units and vacuum tubes totalled 9,700,824. . . . The **Polish** state-owned radio factories produced 30,000 sets, 140,000 loud speakers, and 736 amplifiers in 1948, compared to 7,287 receivers, 85,000 loud speakers, and 293 amplifiers manufactured in **Poland** in 1947. . . . The Government of **Iceland** recently issued a decree subjecting electrical household appliances, including radios, to a fee of 100 per cent of import value of the product in addition to the regular import duty. In 1947 U. S. exports of radio receivers to Iceland amounted to approximately $48,000. . . . Top user of radio in **Latin America** is Brazil, with 1,700,000 sets estimated to be in use. Argentina follows close behind with 1,600,000, and Mexico with 1,113,500.

IRE People

Three members of the Bell Telephone Laboratories have been advanced to new positions: **William H. Martin** (SM'46) has been made a vice-president, **Donald A. Quarles** (M'41–SM'43) has added to his duties as vice-president the charge of staff functions, and **James W. McRae** (A'37–F'47) has succeeded Mr. Quarles as head of the development of transmission, switching, and electronic apparatus.

Mr. Martin has been engaged in telephone research and development for the Bell System for nearly forty years, and he has made distinguished contributions to the development of telephone apparatus.

Mr. Quarles has been associated with the Bell System for thirty years, and has previously served as outside plant development director and transmission development director. He has been vice-chairman of the committee on electronics of the Joint Research and Development Board of the federal government, and a member of the AIEE board of directors.

Mr. McRae, a member of the laboratories since 1937, was director of electronic and television research. He is a member of the AIEE, and of Sigma Xi, and in 1943 received honorable mention for the Eta Kappa Nu awards.

John H. Barron (M'29–SM'43), consulting radio engineer of Washington, D. C., died recently.

Mr. Barron was born on November 16, 1900, in Baltimore, Md., and received a technical education from the Baltimore Polytechnic Institute and The Johns Hopkins University, both in Baltimore.

Becoming interested in radio as an amateur in 1914, he received a first-grade commercial radio operator's license in 1918, and was employed as a ship radio operator for several years. In 1925 he joined the Department of Commerce as a radio inspector, leaving in 1930 to become a radio engineer in the Broadcast Section of the Federal Radio Commission and later the Federal Communications Commission.

In November, 1935, he became engaged in private practice as a consulting engineer, and continued this practice until his death.

Arthur E. Newlon (A'38–SM'44) has left the Stromberg-Carlson Co. in order to work in the Ordnance Research Laboratory of the National Bureau of Standards' Electronics Division.

Mr. Newlon was born in New Lexington, Ohio, on November 15, 1911, and attended Ohio State University, receiving the B.S. in electrical engineering in 1933. Shortly after graduation he joined the staff of the U. S. Coast and Geodetic Survey in Washington as an assistant scientific aide, leaving in 1935 to join the Muzak Corp. of Ohio.

In 1936 he transferred to RCA, where he did research and development work for the license division laboratory on broadcast, television, frequency-modulation, short-wave, and facsimile receivers. After five years there, he entered the employment of Stromberg Carlson as senior radio engineer, where he began by working on communication systems above 100 Mc, television receivers, quartz crystals, radar, and color television; and was advanced to take charge of all broadcast receiver research, including circuits and manufacturing techniques. During this time he also acted as representative of Stromberg Carlson on thirteen RMA committees concerned with television and frequency modulation.

Mr. Newlon is a past chairman of the Rochester Section of the IRE, has been a member of the Rochester Engineering Society's Board of Directors, and is a member of Phi Eta Sigma, Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

Roger L. Merrill (A'46), formerly an engineer with the Curtiss-Wright Corp., has been appointed to the staff of the Battelle Institute at Columbus, Ohio, where he will be engaged in electrical engineering research.

Mr. Merrill holds the bachelor's and master's degrees in electrical engineering from Ohio State University, and is a member of the AIEE, Tau Beta Pi, and Sigma Xi.

Edwin Henry Colpitts (A'14–F'26), whose pioneer achievements led to the development of practical systems of long-distance wire and radio communication and to the establishment of transatlantic telephone service, died recently.

Dr. Colpitts was born in Point de Bute, New Brunswick, Canada, on January 19, 1872, and received the A.B. degree from Mount Allison University in 1893 and from Harvard in 1896. The following year he received the M.A. degree from Harvard, and in 1926 the honorary LL.D. degree from Mount Allison.

From 1897 to 1899 Dr. Colpitts was an assistant in physics at Harvard. In the latter year he joined the American Bell Telephone Co. in Boston as an engineer in the mechanical engineering department, which later was merged with the engineering department of the American Telephone and Telegraph Co. At that time long-distance telephone conversation was limited to 1,000 miles, and in his early years of research Dr. Colpitts helped apply Pupin's theory of electrical loading, and worked on repeater tubes.

In 1907 Dr. Colpitts was transferred to the engineering department of the Western Electric Co. in New York, N. Y., and became head of the physical laboratory. Four years later he was put in charge of the research branch of the engineering department, and in 1917 he was named assistant chief engineer. During the first World War he served on the Chief Signal Officer's staff, AEF.

He returned to the AT&T in 1924 as assistant vice-president in the department of development and research, and, ten years later, when this department was merged with the Bell Telephone Laboratories, he became vice-president of the latter organization. In 1937 he retired and, during his retirement, delivered the Idware lectures sponsored by the Japanese Institute of Electrical Engineers. For this service Emperor Hirohito decorated him with the Fourth Order of Merit of the Sacred Treasure.

Shortly before World War II, Dr. Colpitts came out of retirement to serve with the Anti-Submarine Warfare Division of the National Defense Research Committee. He received the Medal of Merit for his services in this connection, and in 1948 was awarded the Elliott Cresson Medal of the Franklin Institute in Philadelphia for his contribution to long-distance telephony.

Dr. Colpitts was a director of the Engineering Foundation, a fellow of the AIEE, the American Physical Society, the Acoustical Society of America, the American Chemical Society, and the American Association for the Advancement of Science.

Homer R. Oldfield, Jr. (M'46–SM'47) has been appointed sales manager for the government division of the General Electric Co.'s electronics department at Electronics Park, and he has named **James W. Nelson, Jr.**, (A'46–SM'47) to succeed him as head of the Air Force sales section in the division.

Born in Mount Vernon, N. Y., Mr. Oldfield was graduated from the Massachusetts Institute of Technology in 1938 with the B.S. degree in aeronautical engineering. After serving briefly as a draftsman with the Glenn L. Martin Co. in Baltimore, Md., he returned to MIT as a research associate, earning the M.S. degree in 1939.

Mr. Oldfield remained with MIT as an instructor until the outbreak of war in 1941, at which time he went on active duty with the U. S. Army Coast Artillery. In 1945, having attained the rank of major, he was demobilized, and joined the General Electric Co. at Syracuse as head of the Air Force sales section. He is also now serving as commanding officer of the Air National Guard radar unit at Syracuse.

Mr. Nelson is a native of Berkeley, Calif., and received the B.S. degree in electrical engineering from the University of California in 1941. During that year and the next he served as a research associate in the Radiation Laboratory at the Massachusetts Institute of Technology and was engaged in microwave research there.

In 1942 Mr. Nelson joined the U. S. Air Force and served until 1946 as a development engineering officer. He then joined General Electric as a development engineer in the government division, becoming a sales engineer the following year.

Edwin Reginald Love (M'46), associate professor of electrical engineering at the University of Manitoba, died recently.

Born in England on January 18, 1912, Professor Love subsequently came to Canada and was educated at the Winnipeg schools and the University of Manitoba, where he was graduated with the B.S. degree in electrical engineering in 1934. For one semester after his graduation he was a demonstrator in electrical engineering at the University of Manitoba, and then joined the Canadian Westinghouse Co., remaining on their engineering staff until 1940.

In 1940 Professor Love received a commission in the Royal Canadian Corps of Signals, and rose to the rank of captain. Owing to a serious shortage of qualified instructors in electrical engineering, he was lent by the Army to the University of Manitoba for three successive semesters, and on demobilization in 1945 he was taken on the permanent staff as an assistant professor of electrical engineering, and was promoted to the rank of associate professor in 1946.

Keith L. Maurer (A'43), inventor and electrical engineer for the Bell Telephone Laboratories, died of a heart attack recently.

Born in Iowa, Mr. Maurer was educated at Yale University where he received the B.S. degree in 1917 and the Ph.D. four years later. During the first World War he served in the U. S. Navy.

In 1925 he joined the Bell Telephone Co., where he specialized in methods for protecting telephone transmission systems from inductive interference by electrified railways and power systems. There he also invented apparatus for preventing such interference, and served as a consultant to various Bell System operating companies in their studies of this problem.

During the recent war he turned to the development of countermeasures in the field of radio navigation. Later he added to his duties the study of the prevention of electrolytic corrosion of cables. He also helped set up the civilian defense communication system for Nassau County.

Mr. Maurer was a member of the AIEE, the Association of American Railroads, the National Association of Corrosion Engineers, the American Association for the Advancement of Science, the Yale Engineering Society, Phi Delta Theta, and Sigma Xi.

Maxwell K. Goldstein (A'30–SM'48), head of the Office of Naval Research's programs research department, has been selected by a special naval board as its recommended senior technical staff member on the Air Navigation Development Board, a new organization established jointly by the Secretary of Defense and the Secretary of Commerce to formulate and carry out the research and development program for their common air navigation problems. Dr. Goldstein has been granted leave of absence from the Office of Naval Research to assume his duties with the Board.

Dr. Goldstein received the B.S. and Ph.D. degrees in electrical engineering from The Johns Hopkins University. From 1935 to 1937 he participated actively in the development of radio navigation aids at Wright Field. For the next two years he continued these developments and their applications to civil aviation with the Civil Aeronautics Authority. He was later placed in charge of the Radio Direction Finder Section at the Naval Research Laboratory, for which work he was given the Distinguished Civilian Service Award, the Navy's highest civilian award.

In 1944 Dr. Goldstein was made head of the Avigation Section of the Naval Research Laboratory, a position which he held until he joined the ONR staff early in 1948 to organize and head the programs research work of the Naval Science Division. In 1947 he received the Distinguished Civilian Service Award for distinguished contributions to the Naval Service in developing high-frequency direction finding as a vital weapon for combating the German submarine menace during the crucial battle of the Atlantic.

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George C. Schleter (A'38–VA'39) formerly on the research staff of the Naval Research Laboratory, has been appointed to the staff of the National Bureau of Standards, where he will conduct an engineering development program on guided missiles, including missile systems and components, in the Bureau's Electronics Laboratories.

Born in Seymour, Ind., Mr. Schleter attended Purdue University, where he received the B.S. degree in 1922. Upon graduation he joined the faculty of New York University, from which he received the M.S. degree three years later.

In 1925 Mr. Schleter was appointed head of the mathematics and physics departments of Broaddus College in Philippi, W. Va., leaving in 1928 to become a civilian physicist on the staff of the U. S. Army Air Corps at Wright Field. From 1931 to 1934 Mr. Schleter worked on the development of blind landing systems at the National Bureau of Standards, and left in the latter year to work on the development of antennas for ASB radar, radio-transmitters, and confidential microwave research, at the Naval Research Bureau.

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Daniel E. Noble (A'25–SM'44–F'47) and **Angus C. Tregidda** (A'40–SM'44) have been appointed to the Motorola Co's new research laboratory in Phoenix, Arizona. Mr. Noble, director of research and vice-president in charge of communications at the communications and electronics division of Motorola, will be in charge of the laboratory. Dr. Tregidda has been named chief engineer and general manager.

Mr. Noble was born on October 4, 1901. After training in radio engineering at Connecticut State College and the Massachusetts Institute of Technology, he joined the faculty of the University of Connecticut's School of Engineering. In the year 1937–1938 he developed a 100-mc frequency-modulation broadcast system which was used to relay programs from Storrs to Hartford, Conn., for rebroadcast on WDRC.

From 1937 through 1939 Mr. Noble was one of the developers of Connecticut's first FM commercial broadcast system, and also developed the first frequency-modulation police system of this type for Connecticut. In 1940 Mr. Noble joined Galvin Manufacturing Corp.

Dr. Tregidda was born on August 28, 1910. He attended the University of British Columbia from 1928 to 1935, receiving the B.A., B.A.Sc., and M.A. in 1932, 1933, and 1935 respectively. In 1935 he became a teaching fellow at the Massachusetts Institute of Technology, and remained there until he received the Ph.D. in 1939. In that year he became an instructor in electrical measurements and electronics at Kansas State College, leaving in 1941 to work for the U. S. Government at research and development engineering. Several years later he joined the Galvin Manufacturing Corp., where he became a senior research engineer.

Timothy E. Shea (SM'46), assistant engineer of manufacture of the Western Electric Co., has been elected president and a director of the Teletype Corp., a subsidiary of Western Electric.

Mr. Shea received the B.S. and M.A. degrees in electrical engineering from the Massachusetts Institute of Technology, and a B.S. as well from Harvard University. After graduation he joined the Bell Telephone Co., and there, early in his career, he developed filters and networks used for transatlantic radio, carrier telephony, and television. In 1929 he was placed in charge of acoustical, optical, and electrical development for sound motion pictures.

In 1939 Mr. Shea was elected vice-president of Electrical Research Products, Inc., which later became a division of Western Electric. When war was declared in 1941, he was granted leave of absence in order to serve as Director of War Research at Columbia University, and as a member of Division Six, Underseas Warfare Committee, National Defense Research Committee. He directed the New London, Conn., Submarine Research Laboratory and supervised a fleet of engineers and scientists working with the Pacific Fleet at Pearl Harbor.

The Medal for Merit, highest civilian award, and a Presidential citation for exceptional services to the submarine forces of the Navy, were presented him in 1946. Columbia University awarded him the honorary degree of doctor of science. A former treasurer of the Society of Motion Picture Engineers and a Fellow of the Acoustical Society of America, Mr. Shea is the author of "Transmission Networks and Wave Filters."

Frederic W. Townsley (A'32–VA'39), staff engineer for military radio and radar systems at the Signal Corps Engineering Laboratories in Fort Monmouth, N. J., died early this year at the Holloman Air Force Base, Alamogorde, N. M., where he was performing interference surveys of radar and radio systems for the Signal Corps.

Born on April 12, 1905, Mr. Townsley attended St. Lawrence University and Union College from 1923 to 1930, receiving the B.S. degree in physics from St. Lawrence in 1928. In 1930 he joined the General Electric Co. in Schenectady, and subsequently the Radio Corporation of America, leaving in 1938 to become a radio engineer with the Civil Aeronautics Administration in Washington, D. C. From 1939 until 1942 he was on the staff of the Signal Corps Engineering Laboratories in Fort Monmouth, N. J., and in the latter year was commissioned as a lieutenant in the Signal Corps of the U. S. Army. There he served as battalion radio officer and contracting officer.

Resuming civilian status in 1946, he returned to the Signal Corps Engineering Laboratories as staff engineer for military radar and radio systems.

Sections

Chairman		Secretary
H. I. Metz C.A.A. 84 Marietta St. N.W. Atlanta, Ga.	ATLANTA May 20–June 17	M. S. Alexander 2289 Memorial Dr., S.E. Atlanta, Ga.
G. P. Houston 3000 Manhatten Ave. Baltimore, Md.	BALTIMORE	J. W. Hammond 4 Alabama Ct. Baltimore 28, Md.
T. B. Lawrence 1833 Grand Beaumont, Texas	BEAUMONT— PORT ARTHUR	C. E. Laughlin 1292 Liberty Beaumont, Texas
R. W. Hickman Cruft Laboratory Harvard University Cambridge, Mass.	BOSTON	A. F. Coleman Mass. Inst. of Technology 77 Massachusetts Ave. Cambridge, Mass.
G. E. Van Spankeren San Martin 379 Buenos Aires, Arg.	BUENOS AIRES	A. C. Cambre San Martin 379 Buenos Aires, Arg.
J. F. Myers 249 Linwood Ave. Buffalo 9, N. Y.	BUFFALO-NIAGARA May 18–June 15	R. F. Blinzler 76 Woodward Ave. Buffalo 14, N. Y.
M. S. Smith 1701 10th St. Marion, Iowa	CEDAR RAPIDS	V. R. Hudek Collins Radio Co. Cedar Rapids, Iowa
K. W. Jarvis 6058 W. Fullerton Ave. Chicago 39, Ill.	CHICAGO May 20–June 17	Kipling Adams General Radio Co. 920 S. Michigan Ave. Chicago 5, Ill.
C. K. Gieringer 3016 Lischer Ave. Cincinnati, Ohio	CINCINNATI May 17–June 14	F. W. King RR 9 Box 263 College Hill Cincinnati 24, Ohio
F. B. Schramm 2403 Channing Way Cleveland 18, Ohio	CLEVELAND May 26–June 23	J. B. Epperson Box 228 Berea, Ohio
Warren Bauer 376 Crestview Rd. Columbus 2, Ohio	COLUMBUS May 13–June 10	George Mueller Electrical Eng. Dept. Ohio State University Columbus, Ohio
S. E. Warner Aircraft Electronics Assoc. 1031 New Britain Ave. Hartford 10, Conn.	CONNECTICUT VALLEY May 17–June 16	H. L. Krauss Dunham Laboratory Yale University New Haven, Conn.
A. S. Levelle 801 Telephone Bldg. Dallas 2, Texas	DALLAS-FT. WORTH	E. A. Hegar 802 Telephone Bldg. Dallas 2, Texas
George Rappaport 132 East Court Harshman Homes Dayton 3, Ohio	DAYTON May 12	C. J. Marshall 1 Twain Place Dayton 10, Ohio
T. G. Morrissey Radio Station KFEL Albany Hotel Denver, Colo.	DENVER	Hubert Sharp Box 960 Denver 1, Colo.
F. E. Bartlett Radio Station KSO Old Colony Bldg. Des Moines, Iowa	DES MOINES- AMES	O. A. Tennant 3515 Sixth Ave. Des Moines, Iowa
C. F. Kocher 17186 Sioux Rd. Detroit 24, Mich.	DETROIT May 20–June 17	P. L. Gundy 519 N. Wilson Royal Oak, Mich.
R. W. Slinkman Sylvania Electric Products Emporium, Pa.	EMPORIUM	T. M. Woodward 203 E. Fifth St. Emporium, Pa.
J. C. Ferguson Farnsworth Tel. & Radio Co. 3700 E. Pontiac St. Fort Wayne, Ind.	FORT WAYNE	S. I. Harris Farnsworth Tel. & Radio Co. 3702 E. Pontiac Fort Wayne, Ind.
W. H. Carter 1309 Marshall Ave. Houston 6, Texas	HOUSTON	J. C. Robinson 1422 San Jacinto St. Houston 2, Texas
R. E. McCormick 3466 Carrollton Ave. Indianapolis, Ind.	INDIANAPOLIS	Eugene Pulliam 931 N. Parker Ave. Indianapolis, Ind.
E. R. Toporeck Naval Ordnance Test Sta. Inyokern, Calif.	INYOKERN	R. W. Johnson 303 B. Langley China Lake, Calif.
Karl Troeglen KCMO Broadcasting Co. Commerce Bldg. Kansas City 6, Mo.	KANSAS CITY	Mrs. G. L. Curtis 6005 El Monte Mission, Kan.
R. W. Wilton 71 Carling St. London, Ont., Canada	LONDON, ONTARIO	G. H. Hadden 35 Becher St. London, Ont., Canada
Berhard Walley Radio Corp. of America 420 So. San Pedro St. Los Angeles 13, Calif.	LOS ANGELES May 17–June 20	J. J. Fiske Westinghouse Electric Corp. 600 St. Paul Ave. Los Angeles 14, Calif.
O. W. Towner Radio Station WHAS Third & Liberty Louisville, Ky.	LOUISVILLE	D. C. Summerford Radio Station WKLO Henry Clay Hotel Louisville, Ky.
F. J. Van Zeeland Milwaukee School of Eng. 1020 N. Broadway Milwaukee, Wis.	MILWAUKEE	H. F. Loeffler Wisconsin Telephone Co. 722 N. Broadway Milwaukee 1, Wis.
K. R. Patrick RCA Victor Div. 1001 Lenoir St. Montreal, Canada	MONTREAL, QUEBEC May 11–June 8	S. F. Knights Canadian Marconi Co. P.O. Box 1690 Montreal, P. Q., Canada
L. A. Hopkins, Jr. 1711 17th Loop Sandia Base Branch Albuquerque, N. M.	NEW MEXICO	T. S. Church 3079-Q-34th Street Sandia Base Branch Albuquerque, N. M.
J. W. McRae Bell Telephone Labs. Murray Hill, N. J.	NEW YORK May 4–June 1	R. D. Chipp DuMont Telev. Lab. 515 Madison Ave. New York, N. Y.
J. T. Orth 4101 Fort Ave. Lynchburg, Va.	NORTH CAROLINA- VIRGINIA	C. E. Hastings 117 Hampton Rds. Hampton, Va.
W. L. Haney 117 Bourque St. Hull, P. Q.	OTTAWA, ONTARIO May 19–June 16	G. A. Davis 78 Holland Ave. Ottawa, Canada
M. W. Bullock Capital Broadcasting Co. 501 Federal Securities Bldg. Lincoln 8, Nebraska	OMAHA-LINCOLN	B. L. Dunbar Radio Station WOW Omaha, Nebraska
A. N. Curtiss Radio Corp. of America Camden, N. J.	PHILADELPHIA May 5–June 2	C. A. Gunther Radio Corp. of America Front & Cooper Sts. Camden, N. J.
M. A. Schultz 635 Cascade Rd. Forest Hills Borough Pittsburgh, Pa.	PITTSBURGH May 9–June 13	E. W. Marlowe Union Switch & Sig. Co. Swissvale P.O. Pittsburgh 18, Pa.
A. E. Richmond Box 441 Portland 7, Ore.	PORTLAND	Henry Sturtevant Rt. 6, Box 1160 Portland 1, Oregon
A. V. Bedford RCA Laboratories Princeton, N. J.	PRINCETON	L. J. Giacoletto 9 Villa Pl. Eatontown, N. J.
K. J. Gardner 111 East Ave. Rochester 4, N. Y.	ROCHESTER May 19	Gerrard Mountjoy Stromberg-Carlson Co. 100 Carlton Rd. Rochester, N. Y.
E. S. Naschke 1073-57 St. Sacramento 16, Calif.	SACRAMENTO	W. F. Koch 1340 33rd St. Sacramento 14, Calif.
G. M. Cummings 7200 Delta Ave. Richmond Height 17, Mo.	ST. LOUIS	C. E. Harrison 818 S. Kings Highway Blvd. St. Louis 10, Mo.
O. C. Haycock Dept. of Elec. Eng. University of Utah Salt Lake City, Utah	SALT LAKE	M. E. Van Valkenburg Dept. of Elec. Eng. University of Utah Salt Lake City, Utah
C. L. Jeffers Radio Station WOAI 1031 Navarro St. San Antonio, Texas	SAN ANTONIO	L. K. Jonas 267 E. Mayfield Blvd. San Antonio, Texas
L. G. Trolese U. S. Navy Elect. Lab. San Diego 52, Calif.	SAN DIEGO May 3–June 7	Jack Jacoby U. S. Navy Elect. Lab. Point Loma, Calif.
F. R. Brace 955 Jones St. San Francisco 9, Calif.	SAN FRANCISCO	R. A. Isberg Radio Station KRON 901 Mission St. San Francisco 19, Calif.
J. M. Patterson 2009 Nipsic Bremerton Wash.	SEATTLE May 12–June 9	J. E. Hogg General Electric Co. 710 Second Ave. Seattle 1, Wash.

Sections

Chairman		Secretary
F. M. Deerhake 600 Oakwood St. Fayetteville, N. Y.	SYRACUSE	S. E. Clements Dept. of Electrical Eng. Syracuse University Syracuse 10, N. Y.
A. R. Bitter 4292 Monroe St. Toledo 6, Ohio	TOLEDO	J. K. Beins 435 Kenilworth Ave. Toledo 10, Ohio
C. J. Bridgland 266 S. Kingsway Toronto, Ont., Canada	TORONTO, ONTARIO	T. I. Millen 263 Winona Dr. Toronto 9, Ont., Canada
D. A. Murray Fed. Comm. Comm. 208 Uptown P.O. & Federal Cts. Bldg. Saint Paul, Minn.	TWIN CITIES	C. I. Rice Northwest Airlines, Inc. Holman Field Saint Paul 1, Minn.
T. J. Carroll National Bureau of Stand. Washington, D. C.	WASHINGTON May 9–June 13	P. DeF. Mekeel 9203 Sligo Creek Parkway Silver Spring, Md.
J. C. Starks Box 307 Sunbury, Pa.	WILLIAMSPORT May 4–June 1	R. G. Petts Sylvania Electric Products, Inc. 1004 Cherry St. Montoursville, Pa.

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A. R. Kahn Electro-Voice, Inc. Buchanan, Mich.	SOUTH BEND (Chicago Subsection) January 20	A. M. Wiggins Electro-Voice, Inc. Buchanan, Mich.
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S. S. Stevens Trans Canada Airlines Box 2973 Winnipeg, Manit., Canada	WINNIPEG (Toronto Subsection)	S. G. L. Horner Hudson's Bay Co. Brandon Ave. Winnipeg, Manit., Canada

Books

Frequency Analysis, Modulation, and Noise, by Stanford A. Goldman

Published (1948) by the McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 18, N. Y. 427 pages, 6-page index, xiv pages, 180 figures. $6\frac{1}{4} \times 9\frac{1}{4}$. $6.00.

This book will be found useful by all engineers engaged in the development of radio communication systems and equipments. It deals with three main subjects of great importance to them: one is the analysis of the effect of transmission circuits on the kind of signals utilized; the second is a discussion of the fundamental theoretical properties of the various types of carrier modulation; the third is the limitation in the range and quality obtainable in radio communication systems which is due to noise, and, more particularly, fluctuation noise. A considerable part of the material included in the book cannot be found in any existing textbook, at least in such detailed treatment. The present work will thus serve as a very good reference text and a clear introduction to a number of subjects, among which may be cited the relation between bandwidth and reproduction fidelity in video and pulse amplifiers, the distortion due to nonlinearity in phase characteristics, the protection against interference in the case of frequency modulation, the evaluation of noise figures in receiver networks, and the signal-to-noise ratio improvements due to frequency as well as pulse modulation systems.

The book is the outcome of out-of-hour courses which the author has taught to the engineers of the Electronics Receiver Division of the General Electric Co. Though mainly theoretical, it is written in view of practical applications and aims at limiting the amount of mathematics involved in the problems considered to its usefulness for experimental techniques. The derivation of formulas is given in detail, and all definitions and symbols are explained with great care and precision. If the reader has a good knowledge of calculus and general radio engineering, he will find the text easy to read and remarkably clear.

It is perhaps to be regretted that the author should occasionally have lapsed into philosophical considerations which will be considered as superfluous by many of his prospective readers. Such is his introduction on the "Phenomena of Mathematics" and the choice of the existence of complex quantities that obey the laws of the algebra of real quantities as an instance of such phenomena. One would think that the operations on complex quantities had been defined to achieve this result intentionally. A certain number of loose statements are to be found in the text, such as in footnote 3, page 33, where it is said that "e^{jx} is the only function whose derivative, integral, and absolute value are all independent of x." These are minor points, however, and will be easily amended by the type of readers for whom the book is written.

The references are always pertinent without any attempt at being complete. Those to a companion volume, so far unpublished, by the same author on *Transformation Calculus and Radio Transients* might perhaps have been avoided in all cases where equivalent information can be found in available technical literature. The allusions to the outstanding achievements of the great pioneers in mathematics and applied physics are well made to place the works of their followers in correct perspective. A number of interesting exercises add to the value of the book, and they are well chosen to facilitate the understanding of the theories and their possible scope of application. Mathematical tables are included at the end and a good index is provided.

On the whole, this is a very good reference book, which gives a clear and up-to-date treatment of questions not to be easily found elsewhere, as well as some original contributions of the author himself. It is to be recommended to all engineers working in the field of radio communications, and will be of particular interest to those engaged in problems where it is essential to reduce the effect of noise sources to the utmost possible extent.

A. G. CLAVIER
Federal Telecommunication Laboratories
Nutley, N. J.

Techniques in Experimental Electronics, by C. H. Bachman

Published (1948) by John Wiley and Sons, 440 Fourth Ave., New York 16, N. Y. 243 pages+8-page index+vii pages. 128 figures. 5¾×8¾. $3.50.

As its title aptly implies, this book discusses the techniques, the processes, and the equipment employed in development and research work on electronic devices. Here are brought together for the first time many up-to-date details of the production and measurement of vacua, design and assembly of vacuum systems, preparation and processing of cathodes and other electrodes, utilization of phosphors, and other processes and techniques which will help physicists, engineers, and others engaged in production or research work in this field. As a result of his broad experience, the author has been able to include many items of detailed "know-how" previously available only by personal contact with experienced workers in a few industrial or university laboratories.

By concentrating on that branch of electronics relating to conduction of electricity in moderately high vacuum, the author has been able to give a reasonably adequate presentation of a specialized field in this small volume. The table of contents may appear to many readers to be oversimplified, but this deficiency is made up in part by descriptive section headings in the text. The utility of the book might be improved if the table of contents included these section headings.

The book is very readable, the concise style and the many excellent illustrations combining to give a clear presentation of the subject. The difficult choice of a dividing line between techniques for studying a subject, and the study itself, or again, differentiation between experimental and production techniques, is generally well made.

At first glance the treatment may seem elementary to the experienced worker; however, the inclusion of many simple but pertinent details is the very feature that should make this book most valuable. As he reads further, the experienced worker will admire the treatment accorded such controversial subjects as thermionic emission from oxide-coated cathodes. The discussion of the practical aspects of "preparation," "processing," "poisoning," and "reactivation" of such cathodes is very well done, and should prove invaluable to those who do not already have this information.

This book should be required reading for every physicist or engineer entering the field. In that it provides new viewpoints and affords a convenient starting point for developing the techniques required by new problems, it should also prove to be a valuable addition to the library of the experienced worker.

R. R. Law
RCA Laboratories Division
Princeton, N. J.

Elektrische Wellen (Electric Waves), by Winifried Otto Schumann

Published (1948) in German by Carl Hansser Verlag, Munich, Germany. 337 pages+3-page index. 248 figures. 6¾×9¼.

The propagation of electromagnetic waves forms the central theme of this book, which is an outgrowth of lectures given by Professor Schumann at the Technische Hochschule in Munich. Intended for those wishing to acquire a basic understanding of wave propagation, the book starts at an intermediate level, the notation of vector analysis being used throughout, and a knowledge of elementary field theory being assumed.

The fundamental relations of field theory are reviewed briefly in the first twenty-three pages, including discussions of Faraday's law, Gauss's law, Stokes' theorem, boundary relations, and the vector operations of divergence and curl. From these, Maxwell's equations are developed in both integral and differential form. Gradient is then introduced, and with it higher-order differential equations of the potential functions are derived.

Several pages dealing with energy relations and the Poynting vector are followed by fifteen pages on plane waves in dielectric and conducting media, and twenty-six pages on the reflection and transmission of waves at normal and oblique incidence. Wave propagation in ionized media is also treated; fifty-seven pages are devoted to waves of higher order, such as occur in waveguides and resonators; and fifty-eight pages to a comprehensive treatment of wave propagation along plane surfaces. The book concludes with a forty-page discussion of spherical waves, as generated by dipoles and other fundamental forms of antennas. Interspersed with these longer sections are a number of short chapters on such topics as waves in crystals, dielectric waveguides, and the reciprocity theorem.

The centimeter, second, volt, ampere system of units of G. Mie is used, giving the equations an appearance similar to that in the rationalized mks system. The illustrations to the text are numerous and clear.

"Elektrische Wellen" should be of great interest as a reference book for advanced students and research workers.

John D. Kraus
Ohio State University
Columbus 10, Ohio

Practical Spectroscopy, by George R. Harrison, Richard C. Lord, and John F. Loofbourow

Published (1948) by Prentice-Hall, Inc., New York, N. Y. 586 pages+19-page index+xiv pages. 240 figures. 5¾×8¾. $6.65.

To the constantly growing number of scientific and technical men who use spectroscopy as a tool, this book by three members of the Spectroscopy Laboratory at the Massachusetts Institute of Technology will be a very useful guide. Representing as it does the contribution of three scientists—a physician, a chemist, and a biophysicist—the book is triply valuable. There has been a very smooth blending of three viewpoints so that it appeals to a very wide audience in the scientific and technical world. The technologist, little concerned with mathematical analysis and derivation of formulas, cannot complain that there is a preponderance of theoretical material; there is just enough to satisfy the needs of almost all of those who worked in applied spectroscopy.

Since the books deals with the practical aspect of spectroscopy, there is considerable detail given to the description of physical apparatus. Here the profuse illustrations, many of the latest type of equipment and installations, will prove of real value. Techniques, especially as they relate to the production and interpretation of spectra, are fully and clearly discussed. Separate chapters are devoted to the discussion of raman, infrared, ultraviolet (vacuum), and interferometric spectroscopy. The various types of spectrophotometers deservedly receive full treatment, along with considerable discussion of the techniques associated with their use.

Tables summarizing procedures, and listing references, for example, to charts and atlases of spectra, appear throughout the text. At the end of each chapter there is given a list of general references. Two appendices contain a tabulation of the sensitive lines of the elements, arranged in both according to element and to wavelength. One might expect that in what is a very practical treatment of spectroscopy there might be more extensive appendices; that this is not the case is explained by the liberal inclusion of much reference material throughout the main body of the book.

The book is a long one, however; so ramified has the field become that the authors necessarily had to confine themselves and gloss over specialized subjects. It seems that the emphasis on physical apparatus precluded more extensive treatment of applications of spectroscopy, especially in the field of the biological sciences.

With the exception of a few advanced topics, all points developed by the authors are described in direct and easily understood language. The technical worker lacking a rigorous formal training need have no hesitancy in delving into this work. There is little doubt that this book will stand for some time as the standard book to which the practical spectroscopist will make constant reference.

R. E. Lapp
Research and Development Board
Washington, D. C.

Electron-Optics, by Paul Hatschek

Translated by Arthur Palme. Published (1948) by the American Photographic Publishing Co., 353 Newbury St., Boston 15, Mass. 180 pages+3-page index+v pages. 176 figures. 6×9. $3.50.

This is a second edition of the translation of Hatschek's book *Optik des Unsichtbaren*, written in 1935–1936, the first edition having appeared in 1944. A brief section on the electron microscope, electron diffraction, electron multipliers, and television, added by the translater to the first edition, has been retained, and a second supplement, by Dr. Walter Hausz, on particle accelerators, radar, television, and the phasitron, seeks to bring the account up to date.

The main portion of the book is largely concerned with a very elementary presentation of the operation of light and electron lenses—utilizing, as far as possible, examples from everyday experience. The exposition of the functioning of amplifier tubes, cathode-ray tubes, magnetrons, and electron multipliers, as well as the account of the relation between resolution and diffraction, proceed in the same vein. At many points the treatment is decidedly clever and original. Unfortunately, the value of the presentation for the technically untrained reader, for whom the book is obviously intended, is largely nullified by the excessive number of errors—errors in fact, errors in logic, numerical errors, and erroneous illustrations. The employment of unconventional terminology and symbols at various points further detracts from the usefulness of the book as an introduction to electron optics.

These adverse criticisms do not apply to the two supplements, taking up the last thirty-three pages of the book. It is unfortunate, however that the publisher did not utilize the opportunity of a new edition to revise the first supplement, in particular with reference to the electron microscope. As a consequence, this portion also must be regarded as out-of-date. It seems unfortunate that this book was not submitted to a competent reviewer before publication.

V. K. Zworykin
RCA Laboratories Division
Princeton, N. J.

Mathematics for Radio Engineers, by Leonard Mautner

Published (1947) by the Pitman Publishing Corp., 2 W. 45 St., New York 19, N. Y. 319 pages+vii pages +7-page index. 136 figures. 6¼×9¼. $5.00.

Mr. Mautner's book is designed for the so-called practical engineer who wishes to broaden or renew his background in undergraduate mathematics, and in this he is entirely successful. The level of the book, however, is hardly high enough to enable the reader to remove the "difficulty in obtaining maximum benefit" from articles in the technical journals, a further aim stated in the preface.

Well-organized, and clearly presented in an agreeable style, the topics covered include a study of the elementary functions, complex algebra, determinants, an introduction to differential and integral calculus, trigonometric and power series, and elementary linear differential equations. Throughout the text, examples from the field of communication engineering are used to introduce and demonstrate the mathematical principles; thus the book is as much a radio engineering text as a mathematical one. Although this does lead to a certain superficiality in the mathematical approach—nowhere in the chapter on determinants is there given the useful and fundamental definition of a determinant as a sum of specified products—it very clearly achieves the practical objectives specified by the author.

Certain portions of the book are particularly well done; for example, the section which introduces the integral calculus, and the portion of the chapter on Fourier series which very clearly summarizes and tabulates functional symmetries. The chapter on power series is not so well done, for the introduction to this subject is somewhat obscure. It is also unfortunate that the author does not devote more space to the justification of complex algebra in the analyses of elementary alternating-current problems. The simple concept behind this somehow escapes a great many engineering students, and they would have benefited by the clear and logical presentation generally typical of the material in this book.

All in all, Mr. Mautner is to be congratulated on producing a text which will be of greater service to a large body of readers, and which, in general, avoids the extremes of pedantry or oversimplification.

Herbert J. Carlin
Microwave Research Institute
Brooklyn 1, N. Y.

Radar Primer, by J. L. Hornung

Published (1948) by the McGraw-Hill Book Co., Inc., 330 W. 42 St., New York, N. Y. 210 pages+3-page bibliography+5-page index+vi pages. 123 figures. 5¼×8¼. $3.50.

"Radar Primer" presents in nonmathematical terms the fundamental principles upon which the operation of pulsed-radar equipment is based, and discusses its uses and limitations. The related topics—radar altimeter, beacons, loran, pulse communication, microwave relay, television, and sonar—are also briefly covered, and a short history of radar development is included.

The presentation is such that it may easily be understood by a reader having little or no background in the elements of radio and electricity, and the book should, therefore, be very helpful to the user of radar equipment and to the beginning student. Block diagrams of simple radar equipments are given, but no circuit details, and the functions of the principal components are explained.

The text is concise and, for the most part, clear, with copious diagrams and photographs used to augment the explanation. Although most of the illustrations are well chosen, a few are not well scaled. The reader may be confused by the figures on page 42 and the accompanying text, where a $\frac{1}{4}$-microsecond pulse is illustrated, but referred to as a $\frac{1}{2}$-microsecond pulse.

Some topics have been oversimplified in order to keep the material brief and elementary. Although only a few features of military equipment are discussed, the principal aspects of modern merchant-marine radar and ground-controlled-approach (GCA) radar are fully covered. The reader interested chiefly in radar may wish that the ten per cent of the book devoted to television had been used instead for additional material on radar. A section might better have been included discussing the factors that affect radar performance, and why the performance of a given installation may vary considerably from time to time. A bibliography lists thirty-nine other books, many of them elementary, to guide the student who wishes to pursue the subject further.

David B. Hoisington
U. S. Naval School
Monterey, Calif.

Standard Handbook for Electrical Engineers, edited by Archer E. Knowlton

Published (1949) by the McGraw-Hill Book Co., 330 W. 42 St., New York 19, N. Y. 2280 pages+31-page index+xi pages. Over 1,000 figures. 6×9¼. $12.00.

This is the eighth edition of a work which first appeared in 1907, and has been revised from time to time to take account of new developments in electrical engineering.

Since the appearance of the seventh edition in 1941, the war period has seen accelerated developments of new materials, new apparatus, new fields of communication, and new standards and methods of testing. Accordingly, this new edition includes up-to-date sections on magnetic materials, synthetic rubber, and plastics. The recently expanded use of servomechanisms receives attention, and concisely written sections on the essentials of loran, radar, microwave techniques, and fundamental knowledge of the physics of nuclear energy have been added.

The policy of limiting the physical size of the handbook has been so successfully carried out that, without increasing the size of the page, the present edition numbers only two pages more than the seventh edition, and that out of a total of nearly 2,300 pages. This has been accomplished by skillful revision and rearrangement of pages and without the omission of useful material.

The general plan of previous editions has been retained. The whole body of the text has been grouped in twenty-six sections, each written by one or more experts on the subject. Reference to any main section is instantly possible by reason of the use of page indentations, and each subject heading in the section is numbered and printed with bold-faced type. The index lists each of the subject headings throughout the book, giving section and paragraph numbers. Wherever information on a given subject appears in more than one connection, this is at once apparent from the index.

The subject matter is concisely written, giving an outline of essential theory, a statement of fundamental equations, tables of constants, and design data, together with information regarding standard practice.

The primary purpose of the handbook is, of course, to provide for the engineer a compact and convenient source of reference to useful information, formulas, and numerical data. However, although not intended as a textbook on any single branch of electrical engineering, the book's pages and extensive bibliographies offer much of interest to the engineer in fields other than his own specialty, and should be valuable to the engineering teacher also as a source of illustrative material for his courses.

Frederick W. Grover
Union College
Schenectady, N. Y

Understanding Television: What It Is and How It Works, by Orrin E. Dunlap, Jr.

Published (1948) by Greenberg, Publisher, 201 E. 57 St., New York 22, N. Y. 110 pages+11-page glossary+2-page list of television stations+2-page bibliography+3-page index+vi pages. 40 figures. 5¼×8. $2.50.

This book, although of minor value to radio engineers and technicians, provides a useful and informative introduction to television for the layman. The scientific origins of television are described in simple language, and the mechanics of the new art are outlined for the general reader. A chapter on television techniques for the performer is also included, and, finally, a "Television I.Q. Test," which condenses a few of the essential facts of television, principally from the potential set-buyer's point of view, into handy question-and-answer form.

The Specialist Writer*

P. P. ECKERSLEY†, FELLOW, IRE

IF, AS SOME SAY, the Man of Tomorrow is bound to be a specialist, then it is hoped that, in fact as well as in hyperbole, "Tomorrow never comes."

The specialist is prone to refine his competence to so narrow a point as to have, in common with the Euclidian one, position but no magnitude. A belief is current which holds that the world's maladies could be alike diagnosed and cured by the analytical methods common to science and technology; in other words, some believe that the technician must take a greater interest in politics. Mankind has lost faith, and no amount of technical ingenuity or any abstract postulate will restore it; by all means let men of good will enter the political field, but they will not find a logical way to virtue. Science is a complex of changing abstractions, moral values are sempiternal. Plato's republic demanded that a king should be a philosopher, but this cannot be taken to mean that a technician should be a king.

Technicians, specialists, men of science, or what-would-you tend to exaggerate the value, much as they may demonstrate the power, of the intellect. The intellect, unrefreshed by sensibility, develops into a wizened bureaucrat, busy framing schemes to teach the community the art of being ruled or waxing indignant with human beings for being human.

These generalizations, exaggerated and intolerant as they may seem, are the outcome of a contact more with the writings than with the persons of a new generation of technicians and specialists.

In an age of belligerent mediocrity it is understandable that style should be lost; an absence of clarity is, in the circumstances, a more serious matter, since it rebukes utilitarian standards. Clarity of thought, precisely set down, may result in no more than a flat desert of undistinguished sentences, but at least there are no obscurities.

Too many modern articles are obscured by luxuriant branches of mathematical expressions which, spreading across the page, force the puzzled student to grope in an umbrageous text for hints of knowledge. Sylvanus Thompson prefaced his little book about the differential calculus with the words, "What one fool can do another can," and so led us to follow him to be shown the elegant simplicities of the subject. In technical articles, at any rate, the shorthand of mathematics ought to be used sparingly, and then chiefly to reveal the relationships between important quantities. As it is, mathematics appears to exist chiefly to sidestep the effort of clear writing.

As to grammar, the rules, if not forgotten, are certainly disobeyed. Small blame, maybe, to the specialist that he copies the slackness of popular publication, be it by radio, press, or film, but it might be supposed that the habits of care in the handling of things and the meticulous accuracy in statements about quantity, which are habitual to the technologist, would encourage him to an equal nicety in the use of words and to obey rules concerning their arrangement.

A number of us still remember the stern classroom of the early years of the century which, if compared with, would be found to be very different from the sunlit "lazy fair" of the modern school.

The anthropomorphic method of explanation is often as convenient as it is picturesque, but, in common with the decibel, it wants watching. A peculiar aspect of a circuit may be made clearer if viewed through the eyes of a transformer which "sees an inductive load"; nevertheless, it is doubtfully correct to continue to personify the component so that it is a "transformer whose secondary winding" has this or that peculiarity or function. Incidentally, and in this connection, we were for a long time deceived by the term "a long-tailed pair," believing it to have some zoological association. Unless technicians take a pull, they will come to endow the bits and pieces of their trade with sentience as did primitive man the trees and stones of his environment.

Thus an overspecialization in technicalities has resulted in a neglect of the humanities, and this neglect, in its turn, robs the specialist of the power to explain his ideas and discoveries, except to those who already have a specialized knowledge of the subject. The danger is also that the use of jargon may even confuse meaning.

These strictures would be cowardly unless supported by at least a few examples. Thus, to give one, the formal definition of a sideband as "all the frequencies created by modulation" distorts meaning by compression. The statement that "Ohm's Law" (sic) "for alternating current" is that this "current equals the voltage divided by the impedance" is not likely to be found anywhere but in a popular science publication. However, in ratio, is it any more damnable than the publication, in an official glossary, as well as in a widely circulated textbook, that transmission gain in a passive network can be expressed in decibels? Incidentally, how many technicians state Ohm's law correctly? A writer on magnetic amplifiers may be excused on the basis of "you-know-what-I-mean" for saying that "a magnetic amplifier is a choke with a dc winding" (through which, a little later on, a "dc current flows"), but the student is not much helped by such a definition. These few examples exist to illustrate not so much a lack of knowledge as "sloppy thinking," avoidance of difficulty, and consequent ugliness of presentation.

On the contrary, attempts to shine as an incisive purist may all too often end in disaster. Mounted on a literary Rozinante, out of a Cliche by Verbiage, the would-be stylist prances around with all the appearance of one about to hit the point fairly and squarely and then comically misses it. For instance:

"Now the current resulting from the application of an E.M.F. if *allowed to flow long enough* will cause a change (of) resistance and therefore voltage/current ratio. By a process of confused thought this change is sometimes adduced as a reason for saying that the device is a non-linear resistor, but, in the interests of clarity, this error should be avoided."

Good writing makes many demands upon patience and skill. Someone, recalling the pain of effort, has spoken of it as "chipping words out of the breastbone." The pain may nevertheless be supported for the sake of ultimate pleasure. To clear sentences of ambiguity, to set exactly the right word in exactly the right framework, to feel words flow in an effortless sequence, is to experience just as much delight as that which rewards discovery or invention.

"Where every word is at home,
Taking its place to support the others,
The word neither diffident nor ostentatious,
An easy commerce of the old and new,
The common word exact without vulgarity,
The formal word precise but not pedantic,
The complete consort dancing together."[1]

Before ever good writing is possible, one must have that strong desire, felt by all real people, to share delights with others. This desire to share is seen to dominate the child who, with two bits of stick and a twist of string, has invented perpetual motion and runs to a parent to tell how. The shades of the prison house do not encourage the fragile growth of sensibilities; as the man supplants the child he no longer indulges the natural gaiety that comes with "making things," for fear he might not be taken seriously. And so gradually he becomes the slave of specialization. This is a slavery that may well condemn a man to mental and physical destitution, should he never escape it.

The quality of writing is style: style is the performance of an act without waste; it is the contribution of expertism to culture. Culture is the refinement of mind, taste, and manners. To attain style is to become a whole man; to become a whole man is to live in the whole world, not in the slit trench of specialization. *Le style c'est l'homme.*

If between the lines of this writing the reader may detect an old man's tantrums, let these be excused because, in practicing in the past what he preaches in the present, he has failed to find any high place in the ranks of a rewarding bureaucracy.

* Original manuscript received by the Institute, November 12, 1948.

† Telephone Manufacturing Company, St. Mary Gray, Kent, England.

[1] By T. S. Eliot.

Atomic Energy—Its Release, Utilization, and Control*

R. A. MILLIKAN†

The PROCEEDINGS OF THE I.R.E. is publishing, in accordance with the policy of the Institute, a series of papers dealing with instrumentation and controls in the field of the production and utilization of fissionable materials. The members of the Institute present at the IRE West Coast Convention at Los Angeles in 1948 were fortunate in hearing a paper dealing with the broad aspects of atomic energy, delivered by a world-renowned physicist. The speaker was a Nobel prize winner, a pioneer in the field of studies of atomic structure and of cosmic rays, and a leader in physical research.

The utilization to best advantage of fissionable materials available on earth is manifestly of major human importance. The following paper deals primarily with this subject, and is commended to the attention of the readers of the PROCEEDINGS OF THE I.R.E.—*The Editor.*

I AM NOT presumptuous enough to assume that I can add much to the knowledge of physics—classical or modern—possessed by a group of well-trained, active, and competent radio engineers. I have only one advantage over you. Many of the developments which you have learned from books I have had first-hand contact with, for they have all come within my active lifetime—in fact, since 1895—so that I may be able to fill in details and experiences which might throw some new light into the darker corners of your pictures of events, and therefore be at least of some interest to you.

We physicists are wont to take Roentgen's discovery of X rays as marking the beginning of what we call modern physics and closing the period of classical physics. The practical dollar-and-cents value of that discovery of Roentgen's is brought home to every family on earth when it is presented with the doctor's bills for the X rays he now insists upon seeing before telling any one of the family what is the matter with him.

But now, going beyond mere dollar-and-cents values, why was this discovery so important scientifically? This following incident in my life furnishes the answer. I had attended the convocation at the University of Chicago in the summer of 1894 at which Professor Michelson had said in his commencement address that it was probable that all the great discoveries in physics had already been made, and that further progress might have to be found in measurements in the sixth place of decimals. Again, in my second year of attendance at Columbia I had lived in a fifth-floor flat on Sixty-fourth Street, a block west of Broadway, with four other Columbia graduate students, one a medic and the other three working in sociology and political science, and I was ragged continuously by all of them for sticking to a "finished," yes, a "dead subject" like physics, when the new "live" field of the social sciences was just being opened up. But here in Roentgen's discovery, only one year later, was a door opening into a new, theretofore-undreamed-of field of physics, a big, qualiattive field which had nothing to do with great refinement of measurements, and there I was in Berlin in the year 1895–1896, participating in the discussions just where and when it was being born. It was clearly just to be there in 1895 that I had been born in 1868!

The reason we physicists usually date the rise of modern physics from the discovery of X rays in December, 1895—I shall here designate it as discovery number 1—is that in the immediate train of that discovery, and stimulated by it, there followed quickly the discovery in Paris in 1896 essentially of the transmutation of the elements through Becquerel's proof of the existence of radioactivity—discovery number 2. Again, there followed a year later, with some aid from X rays, J. J. Thomson's demonstration in England, to practically everyone's satisfaction, of the concept of the negative electrons, called by Thomson "corpuscles," as fundamental constituents of all atoms in the universe—discovery number 3. Further, these two discoveries were soon followed by Planck's even more fundamental one—discovery number 4—conceived in the course of lectures I took with him in 1896; namely, the discovery of *discontinuous*, jump-like, or "quantum" energy and momentum changes. The reason this discovery had not been made earlier was that man here was entering a thus-far almost completely unexplored domain; namely, the domain of subatomic or microscopic, as distinguished from ordinary or macroscopic, energy and momentum exchanges. This field is most simply entered through photoelectric research, already found industrially useful. These three last discoveries—radioactivity, electronics, and quanta—actually determined the direction of my own study and research for the next fifty years.

Of these three great discoveries which were made in "the dead field of physics" during the last five years of the nineteenth century—namely, that of radioactivity, that of the negative electron as a fundamental constituent and cementing agent of all atoms in the universe, and that of the quantum as a fundamental unit involved in all atomic-energy transformations—the second, the electron, with its myriad of extensions and applications to radar, to communications of all kinds, to pictures, and to a score of other industries, has been *the most useful to mankind;* the first, radioactivity, and the third, quanta, *the most revolutionary to human thought*. Among the three, radioactivity claimed my own attention at Chicago first.

I have just said that radioactivity was revolutionary to human thought. That was because some, even of the "eternal atoms," namely, those of uranium and thorium, are unstable and are spontaneously throwing off with great energy pieces of themselves, thus transforming themselves into other atoms, following the universal tendency of matter to pass over into a state of maximum stability—a state that has already been practically reached by the atoms of all save a few very rare and very heavy elements; hydrogen, for the reasons soon to be given, being one exception.

But in 1905—a great year!—came the greatest advance of all—discovery number 5—when Einstein laid the foundations of a better understanding of the fundamental sources of energy in the universe than we had thus far had. He did it through the setting up, as a consequence of his special relativity theory, of the equation $E=mc^2$, in which m is mass in grams, c is the speed of light in centimeters (30,000,000,000 centimeters per second), and E is energy in absolute energy units, namely, ergs.

The conception here is the exceedingly important one, namely, *that matter is itself convertible into radiant energy*, or, to take a concrete case, that the sun has been able to continue for the past three billion years pouring out heat and light at its present prodigious rate only because it is continuously feeding its own mass into its furnaces and consequently shrinking its waistband all the time, and sending the lost weight out in the form of radiant heat in accordance with the requirements of the foregoing equation. This equation has now been repeatedly checked in nuclear experiments in our own and in many other laboratories, and has never yet been found to fail, when the reaction could be made to take place.

But note that Einstein's equation does not indicate under what conditions, if any, the transformation of matter into radiant energy *will actually* take place, if at all. It only indicates the quantitative relations which must be fulfilled if and when the reaction does take place. The great practical significance of this equation will only be seen by looking more closely at the factors involved. It is to be noted that it says that if one gram of matter is transformed completely into ergs of energy, to get it into ergs that one gram must be multiplied by the stupendous quantity C^2 (C being 3×10^{10} cm), and the result is 9×10^{20} ergs. Let us *assume*

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† California Institute of Technology, Pasadena, Calif.

that such a transformation does take place in one second, remembering that to get the result into practical power units, namely, kilowatts, we must divide 9×10^{20} by the number of ergs in a kilowatt, which is 10^{10}. But the kilowatts that you then have created are 9×10^{10}, or *90 billion kilowatts*. The transformation, then, of one single gram of matter per second should yield 90 billion kilowatts of power—almost too large an amount even for the imagination to grasp. In view of the immense number of grams of matter in the sun, it can be easily computed that the sun could keep pouring out heat at its present rate and last for hundreds of billions of years, and the *assumption* that such transformation can and does somehow take place would remove entirely the theretofore great mystery as to how the sun gets its heat. For, in view of Bethe's recent work at Cornell, practically all physicists now accept the evidence that the sun's heat is maintained by the transformation of mass into radiant energy; i.e., we regard this assumption as furnishing the correct explanation of the great mystery of the immensity of the heat output of the sun and the stars.

Here is, then, in fact, in the sun and stars the great *inexhaustible source* of atomic energy, though it is nothing but the actual source from which we have always known that we get all our power, namely, from present or past radiant energy shot down from the sun to earth and here bottled in oil or coal or wood or waterfalls. The total amount of available energy thus stored in coal and oil is unquestionably very many times larger than that which is stored in the unstable atoms of uranium and thorium, and which is also now practically releasable by man through the disintegration of uranium and thorium into lighter, more stable atoms. For reasons given in the next paragraph, this is called "packing-fraction energy."

For, according to the estimates of the astrophysicists, the universe is still 90 per cent hydrogen. Our sun is constituted largely thus. Bethe estimates that 80 per cent of the sun's atoms are hydrogen. At the enormous pressures and temperatures existing within the sun (temperature = 35,000,000°), four atoms of hydrogen can sometimes get together directly or indirectly and form one atom of helium, but the mass of a free hydrogen atom is 1.0082 atomic-mass units (AMU) and the mass of the four of them before their union was four times the foregoing number, or 4.033 units, while the *directly measured* mass of the helium atom formed by the combination of the four hydrogens is only 4.003 units, so that the difference, 0.03 AMU units, or roughly 1 per cent of the mass of the four hydrogen atoms involved, has been *lost* in the process of combining the four hydrogen atoms into one helium atom. The mass that has so disappeared is called the *packing-fraction energy* released in this particular reaction as radiant energy. The packing-fraction energy released in the building up of most of the common abundant atoms out of hydrogen, as they have in fact been built up, is not far per hydrogen atom from the above value involved in the building of helium.

When these packing-fraction energies were first worked out thirty-five years ago, some enthusiastic scientists followed their Jules Verne urge and got into print with the statement that there was enough atomic energy in a cupful of sea water to drive the biggest ship across the Atlantic, since two out of three of the atoms in every molecule of H_2O are hydrogen. They might have told a much bigger story than that, namely, that *if it were possible to make the hydrogen in all the seas combine at once into helium*, or indeed into any of the common elements, than we could probably explode the whole earth and transform it into a nebula with the packing-fraction energy thus set free.

The difficulty is that *so far as we can now see* this particular packing-fraction job is one which the Great Architect assigned only to the titanic pressures and temperatures existing in the interiors of the stars, not to puny man or even to cold-blooded Mother Earth, for it takes a warmer lady than she is to do that job.

This situation has been so well known that for the past thirty-five years I have not heard of a scientist or even a newspaper columnist who grew jittery over the prospect of man's setting off that explosion. *For the present, then, we have given up as unattainable on earth* (save as the sun sends it down to us) *this source of packing fraction energy*, incomparably greater than that obtainable from uranium, even if the reaction of its complete disintegration into smaller and stabler atoms, such as barium and krypton, for example, could be brought about on earth.

The only other kind of process which might release "packing-fraction energy" rapidly *on earth* and under man's control would be somehow to cause the heavy, unstable atoms of uranium and thorium to disintegrate into some of the common stable elements faster than they are doing through normal radioactive processes. This was the discovery made in 1939 of "uranium fission." It was found that when one of the uranium isotopes called U235 is sprayed with neutrons—which are essentially the nuclei of hydrogen atoms minus their positive charges—it instantly disintegrates into two nearly equal fragments with an emission of more neutrons which enter the surrounding U235 nuclei and split them up, thus starting a chain reaction which instantly spreads to all the U235 nuclei within reach. This constitutes an explosion some hundreds of thousands times more powerful, weight for weight, than TNT. This problem of thus instantly releasing the packing-fraction energy of uranium has been an elegant and very difficult scientific job, and it has been accomplished and spectacularly demonstrated to the whole world in the atomic bomb. The foregoing method of releasing atomic energy is limited, so far as we now know, to the two elements, uranium and thorium. These elements are already very rare and therefore very expensive, and of course if we should begin to use them up for ordinary power or heating purposes they would very quickly become rarer and their cost would skyrocket. Whereas, when the bottled sunlight energy, which we have in oil and coal, is all gone, say four thousand years hence, we can use water, wind, and tide power more fully than now, and grow fuel crops in the tropics, and also develop heat engines for using the sun's rays directly, as we do in a small way in solar heaters now, and thus get our power somewhat more expensively than now, it is true, but not prohibitively so.

The results of the fission of uranium can be applied in many useful ways, mostly biological, but in my judgment *only where high costs are not an important factor*. As an *economical, long-range* source of power for the power industry, this method is impractical because of the rarity of the uranium and thorium. I venture this opinion in spite of George Eliot's warning that prophecy is the most gratuitous form of mistake. Whose mistake it is, the next generation will know.

So far I have said nothing about the precise process by which packing-fraction energy is released. Please recall, then, that all atoms are built up out of the primordial atom of hydrogen—more accurately, out of protons and neutrons. Then, with the aid of Aston's packing-fraction curve (Fig. 1), we can see the whole story. In this curve, abscissas are simply the increasing atomic weight of the elements from 1 to 240, and each ordinate is the accurately measured atomic mass of the element in question divided by the number of hydrogen atoms that have entered into its composition; i.e., it is the mass spectrograph's direct finding of the mass of the hydrogen atom *after its incorporation into the structure of the atom under consideration*. But, in becoming so closely packed together, these constituent hydrogen atoms have lost potential energy, and hence, according to Einstein's equation, have also lost mass. This is why the atom of hydrogen is heavier when it is free than when it is in any of its combinations.

With the aid of this packing-fraction curve, the release of atomic energy in any possible nuclear act of transmutation from a less stable atom to a more stable one can be seen at once.

Thus, the curve shows clearly the two sorts of atomic transformations which, *if they can be made to take place*, will release energy; namely, first the building up, or synthesis, of any of the elements out of the primordial element hydrogen.

But, secondly, the curve also shows that the mass of the hydrogen constituents of the very heavy elements at its extreme right end is greater than the mass of the hydrogen atoms that are constituents of the common elements lying along or near the bottom of the curve. Hence, any *disintegration* of these heaviest atoms, like uranium and thorium, into the stabler atoms that lie on the lower part of the curve must also be accompanied by a release of atomic energy, though of much lesser amount per hydrogen atom involved than in the case of the synthesis discussed above. *It is thus synthesis, not disintegration, that is the great source of atomic energy.*

The bottom of this curve is a position from which atoms cannot be transformed into other atoms *with any release of packing-fraction energy at all*. In other words, there is no atomic energy available on earth or, according to the generally accepted view of Bethe, even in the sun and stars, save packing-fraction energy, and packing-fraction energy never permits more than about 1 per cent of the mass energy of the participating hydrogen atoms to be transformed into heat.

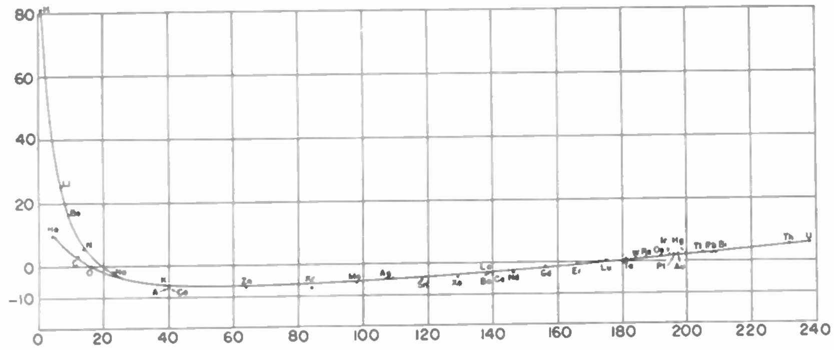

Fig. 1—The packing-fraction curve. Abscissas are the series of elements of atomic weight from 1 to 240. Ordinates represent the difference between the weight of the uncombined hydrogen atom 1.0082 and the same atom in combination in each of the series of elements. This difference is the mass available for the transformation into radiant energy in accordance with the Einstein equation $E=mc^2$.

This puts a limitation on the present and future supply of energy from atomic sources that even some of the workers in nuclear physics may not yet have realized.

The following figures have a significant bearing on the available uranium packing-fraction energy.

The total world supply of coal alone is estimated to be about 8,000,000,000,000 tons. The present yearly consumption is about 2,000,000,000. At this rate, the coal will last 4,000 years. The world's "easily available" supply of uranium has been estimated[1] at 30,000 tons having a fissionable content of 0.7 per cent, or 210 tons. The Smythe report gives, energy-wise, 1 ton of U235 = 3,000,000 tons of coal. The annual energy drain on coal beds would thus exhaust the world's supply of U235 in four months. If all uranium atoms were made fissionable (possible, but *very* costly), the foregoing uranium supply would last forty-five years.

Again, how rare, from a *practical* point of view, uranium is can be seen from that portion of the history that I have myself followed closely since 1900, when I began searching the world over for uranium ores. For the first ten or twelve years after the discovery of radium in 1898, it was the pitchblende ore from the Joachimsthal mines in Austria (now Czechoslovakia) that supplied the world demand for radium for scientific and therapeutic uses. Then, because of its extreme scarcity, the Austrian government put an embargo on the export of this ore. At this time Joseph Flannery, who had been president of the Vanadium Company of Pittsburgh and had promoted the use of vanadium as an alloy of steel, turned his attention to a careful study of the uranium situation. He found that "the ores of Europe were out of the question" because of the Austrian monopoly, and that "the few deposits that were found in other parts of the world were not of sufficient extent to justify serious consideration." However, in a desolate section of southwestern Colorado and northeastern Utah there were 800 square miles of carnotite deposits—sands on the grains of which the yellow oxide of uranium has crystallized out. These low-grade Colorado carnotites averaged less than 1 per cent uranium, while the Joachimsthal pitchblende had run up to 10 per cent uranium.

Knowing that both Professor Herbert McCoy and I had been working on uranium ores, Mr. Flannery approached us to find a young graduate student whom we had trained. We recommended to him Charles D. Viol, who went to Pittsburgh as director of Mr. Flannery's new Radium Laboratories of the Standard Chemical Company of Pittsburgh. Under Dr. Viol's directorship this company succeeded in developing a process for the extraction of radium from carnotite, turned out its first commercial radium (2.1 grams) in 1913, and between 1913 and 1918 produced 39 grams of radium element, which was then marketed at $125,000 a gram. Dr. Viol also estimated that the total radium production in the United States up to 1919 was 55 grams of radium element, "probably more than half of all the radium produced in the world up to that date." By 1921 this company had raised its total production from the Colorado carnotites to 72 grams. In the single year of 1920 it produced 18.5 grams. Further, in view of the fact that the United States Bureau of Mines had estimated the total radium content of the Colorado carnotites to be but 135 grams all told, and that the Standard Chemical Company's carnotite holdings were much larger than those of any other single company, the latter organization made a careful survey (in part by systematic diamond drilling), and estimated the total Colorado carnotite's content of radium metal as at least 500 grams, in place of the Bureau of Mines' estimate of 135 grams—excellent agreement as to order of magnitude.

In 1922 I was C. R. B. Exchange Professor to Belgium and was a guest of Mr. Franqui, Belgian industrialist and financier, who, as a young man, had been instrumental in acquiring for his country the Belgian Congo. He told me he was in a position to destroy the United States' monopoly in the production of radium, for, said he, "Your poor carnotite ore—and there is no other available to you—cannot compete with the new rich Congo pitchblende ore, which occasionally runs up to 15 per cent uranium metal."

Mr. Franqui made good on his boast, and within two years the two chief radium-producing companies in the United States had gone out of business, while the Belgian Congo took over the world's monopoly in the production of uranium, holding it for some ten years, until competition from the Big Bear Lake uranium mines in Canada began to bring down the price somewhat. My memory is that the second gram of radium presented to Madame Curie cost but $70,000. But even in 1940–1945, 90 per cent of all the 1,100 tons of pitchblende ore which was purchased by the United States and used in the development of the uranium bomb came from the Belgian Congo.

Thus, after fifty years of search by government bureaus of mines and private individuals and companies, there are now but two places in the world—Canada and the Belgian Congo—from which uranium ores can be obtained in considerable commercial quantities, and two more places—Colorado and Joachimsthal—where smaller deposits exist. In view of that record and the above-quoted abundance figures, I do not anticipate that either intelligent governments or intelligent individuals will use uranium for any major fuel or power purposes. It is too valuable a material for scientific purposes, for public health purposes, and for such industrial purposes as require small quantities (luminous paints, alloys, etc.) to be wasted on major power projects so fully provided for by the inexhaustible supplies of solar energy, past or present. According to the Atomic Energy Commission, uranium and thorium *are the only elements which have a chance in disintegrating of releasing atomic packing-fraction energy*, and there is no available energy save packing-fraction energy. Because of the extreme rarity of these elements, wisdom calls for their conservation rather than for their use for fuel or power.

Alfred Nobel, a great pacifist, was convinced that his invention of nitroglycerine had done more to bring about the elimination of war than all the sermons and peace conferences had ever done or could ever do; and with the experiences of the two world wars of the past thirty years before them, the great majority of intelligent men had determined, even before the advent of the atomic bomb, to join the ranks of those determined to put a stop, if possible, to international wars, since otherwise the race is headed for self-annihilation. *The great service to mankind of the advent of the atomic bomb has been to make as clear as crystal, to all classes and conditions of men the world over, the necessity for such action.* In my judgment, no other service that the study of atomic energy can render to mankind is likely to be in any way comparable to this service; and the other services could all be profitably discarded if this one succeeds.

The necessity that it has placed upon all mankind to find a substitute for war in the handling of its international relations is that which, in itself, without reference to further inventions of any kind, will make *a new world;* and its influence in making a new world will probably be greater than all other influences combined. Without it, there certainly will be no worth-while new world. I expect the influence of industrial applications to be wholly negligible in comparison with this supreme service.

[1] Clark Goodman, "Petroleum versus plutonium," *Lamp*, vol. 25, p. 10; February, 1946.

Quality Control in Radio-Tube Manufacture*

J. ALFRED DAVIES†

Summary—**Methods of quality control in the radio-tube manufacturing industry are surveyed. Typical mount-inspection service, use of statistical control charts, and sampling procedures are discussed.**

Introduction

In the past five years, quality control has firmly established itself in the radio-tube industry. Although the methods are relatively new, the rapid acceptance of quality control shows that it has proved a considerable aid to engineering, production, and inspection. However, statistical quality control in this industry is still only partially developed. With continued developments of statistical methods which are easily applied in the factory, the future of quality control is indeed promising.

Survey of Quality Control in the Radio-Tube Industry

The over-all view of quality control in this industry is a broad and progressive one, indicating that the proven tools of statistical quality control have been adapted to each manufacturer's requirements for inspection of finished tubes, and are now being extended to cover most phases of production.

The previous statements are based on the current status of quality control as revealed by an industry-wide survey which the author made recently. A quality-control questionnaire was sent to each radio-tube manufacturer listed in the RMA index. The response was encouraging and quite appreciated. Replies were received from about two-thirds of these companies, including the three top producers. A summary of the replies is given in the following statements.

1. All radio-tube manufacturers responding use statistical quality control.
2. These companies all have a quality-control department or section.
3. In general, the quality-control department is a separate group which reports directly to top management.
4. The emphasis is placed much more on manufacturing than on engineering operations. In manufacturing the emphasis is very strong on finished tubes, strong on incoming materials and parts processing, and low on assembly and exhaust. In engineering operations a fair amount of emphasis is placed on specifications and on design.
5. The types of quality-control techniques applied in order of usage are sampling-acceptance plans, control charts, and miscellaneous methods.
6. The extent of these techniques, according to their application, is as follows:

A. Among sampling plans, double sampling is heavily favored. Single sampling and sequential sampling follow in order.

B. Under control charts, the percentage defective is used considerably, while charts for defects per unit rank next, and charts for average and range, and for average and deviation, come last.

C. In the miscellaneous group, the leader is frequency distributions. The assorted techniques include some unique schemes among which are a quality control on testing equipment, a percentage rating for tube-life tests and a grand-lot continuous sampling for finished tubes.

7. The most successful applications have been in control of quality of finished tubes. This conclusion is right in line with the fact that the most emphasis has been placed on finished tubes in the manufacturing operations.

Manufacturing Variation

There are many applications of quality-control methods in radio-tube manufacture, but all have the same aim—the control of variation. Variation is inherent to the nature of all manufactured products. It is the main function of quality-control engineers to devise and apply methods to stabilize this variation within expected limits.

Immediately the following questions arise: How much variation is normally expected? When does it become excessive? How did it develop?

Statistical quality control supplies answers to the first two questions by indicating probability limits of variation and the trend of quality. Engineering investigation is required to uncover the cause, but statistical quality control can assist even in this direction. Analysis of the production at the point at which the trend started saves the production engineer a great deal of time in identifying the cause of abnormal variation. Also, the production man is spared the waste motion of trying to find a nonexistent cause of product variation when it is within expected limits. *It is evident that these so-called "control limits" are also economic limits.* Their application is made through the quality-control techniques, such as control charts, frequency distributions, or sampling inspections.

Quality Control in Owensboro Tube Works

It is felt that a description of the methods used at the Receiving Tube Works of the General Electric Com-

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† General Electric Company, Owensboro, Ky.

pany will portray an adequate picture of how quality of radio tubes is controlled for the joint interest of manufacturer and consumer.

We have centered the body of controls on finished tubes, but also are now attempting to establish sound quality-control procedures throughout the manufacturing operations from purchased materials to shipped tubes—and finally to satisfied customers.

Fig. 1, showing an operations flow chart, will be helpful at this point. This diagram shows the general order of manufacturing operations, together with the main materials and processes from incoming materials to outgoing tubes. Our quality-control functions either have been or are now being developed in the following main divisions of manufacturing; incoming materials; parts and processing; grid making; mount assembly; exhaust; finished tubes; life testing; preshipment inspection; and rework. In addition to these connections with manufacturing our quality-control section has tie-ins with Specifications for setting or re-setting limits; with Design Engineering for analysis of pilot runs, tube development, and evaluation of tube quality in field application; and with Production Engineering for design and analysis of experiments.

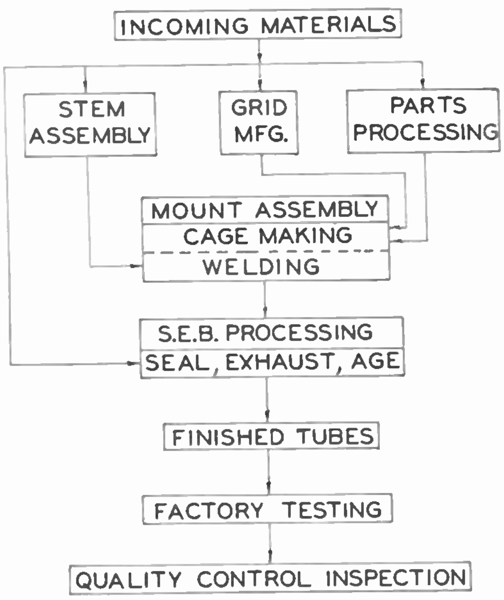

Fig. 1—Operations flow chart.

Our quality-control functions are administered by two groups: Process Quality Control, and Quality-Control Section. As the name implies, the function of the former group embraces all phases of production from incoming materials up to finished tubes. Operations of the latter group begin with finished tubes and extend through the shipped product.

It has been our experience that there is no best place to initiate a new control, since there is no such thing as an isolated quality-control operation. A control method started in one section of manufacturing always shows effects in several others. Although the first point considered for starting control methods is usually at inspection of finished tubes, it is not necessarily the most effective, and certainly not the most economic. The swing is toward maintaining a high quality level throughout production by control methods along the way at key spots. This concept is in direct opposition to the former idea of sorting the good from the bad tubes after manufacture by 100 per cent inspection. Quality cannot be inspected into a tube—quality must be built in.

Quality-Control Methods for Description

Generally, the usual statistical quality-control techniques of sampling-acceptance plans or control charts have been adapted to our local plant personnel. In addition, we have developed some modifications of the usual control methods which have proved quite successful. Two of these methods which will be explained are:

1. Assembly or mount inspection which uses a combination of sampling for attributes and a control chart for percentage defective.

2. Finished-tubes quality control effected by a median control chart and double sampling.

Assembly or Mount Inspection

Utility

Mount inspection is an extremely important operation which is used as a barometer for forecasting the amount of factory shrinkage or tube losses at final testing, and the quality level of finished tubes. Although no absolute relation exists between the results of mount inspection and over-all quality at final test, a sufficiently close correlation for all practical purposes has been established using a time lag of two to three days. A separate correlation is made by tube type, so that final inspection results can be estimated from mount-inspection records. To focus attention on the importance of completely rechecking trays of mounts that showed questionable or inferior quality at inspection, the device of Inspection Service was developed.

Fundamentals of Mount-Inspection Service

Briefly, Inspection Service means sampling inspection for attributes (either OK or defective mounts) without a decision on acceptance or rejection, but merely a visual record of results at inspection. The decision concerning the disposition of the lot is made the responsibility of the Mounting Section. This is probably better psychology than quality control, but in practice it has directed action at the right place and time, thereby disclosing the cause of excessive rejects. This method has also been instrumental in selling the idea that quality control is a guide, not a police force. We have found that the mounting operators are more critical of their own work and quality than inspectors ever have been. A control chart which records the inspection results is kept by the assistant foreman of Mounting for each mounting unit.

Procedure of Mount Inspection Service

All trays of mounts (50 per tray) are subject to mount inspection. A sample of ten mounts or 20 per cent of the number of mounts per tray is carefully inspected visually under a magnifying glass by Mount Inspection personnel. Inspected mounts are judged OK, questionable, or defective. Since the inspection is visual, distinct definitions of questionable and defective mounts have been made to minimize incorrect decisions by inspectors. Defective mounts are defined by tube type as those mounts which, in the opinion of Engineering and Mounting Supervision, are made inoperative by improper assembly. Each defective mount is tagged with a yellow ticket fastened to the exhaust tube, and is counted as a definite cause of mounting shrinkage. A questionable mount is tagged with a white ticket and is not recorded, being considered only a possible cause of mounting shrinkage.

Each tray of mounts, together with its Inspection Record, is turned over to the Mounting Section foremen who make the decision concerning the disposition of that tray of mounts. Disposition may be made in any one of the following ways:

1. Release tray to Exhaust after reworking the defective mounts tagged.
2. Release tray to Exhaust after defective mounts have been repaired and the remainder inspected by the mount unit responsible.
3. Request additional inspection by Mount Inspection after carefully checking over all mounts on the tray.

A daily summary for each mount unit is prepared by inspection personnel showing the number of mounts inspected, number defective, percentage defective, and type of defects. The percentage defective is plotted on a control chart for each mount unit. Analyses of out-of-control points are traced back to the operator responsible by means of the type of defect which shows whether cage-making or welding is the cause.

Use of Mount Inspection Records

Attention is first directed toward a tube type whose over-all control chart shows a point out of control. Next, the charts for all mount units making that type are reviewed to locate the unit or units responsible for excessive rejects. Finally, the summary sheet for type of defects is used to disclose the nature of main rejects. Armed with this specific information daily, the group leader can readily spot the operators whose work is inferior, and thereby correct the mounting problem at its source. However, the operator may not be at fault. The reason for defective mounts may be poor materials or a faulty welder. In any case, the welding machines and the parts are checked.

The inspection records provide a current check on quality. Engineering and Mounting use the records to dispose of the portion of production which is sampled and to reduce defects of a similar kind in future production.

The control charts are the usual so-called "p-charts" using the percentage defective rather than the fraction defective. Fig. 2, showing a mount-assembly control chart of percentage defective, illustrates the nature of these charts. This is a typical control chart, although all figures used in it are hypothetical. A solid line labelled

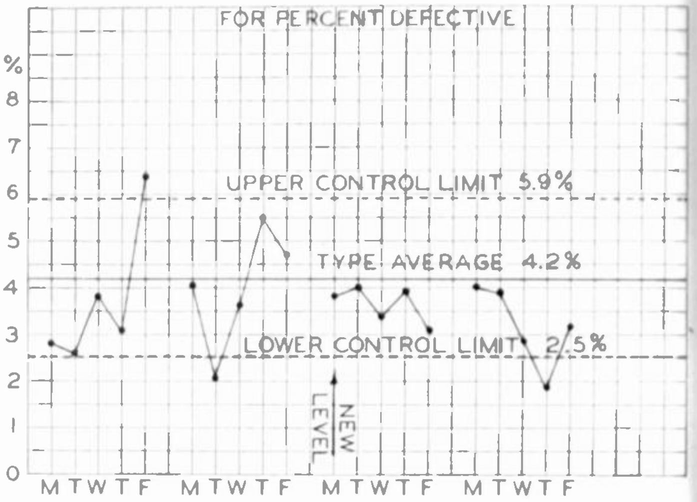

Fig. 2—Mount assembly control chart.

Type Average represents the average percentage of defective mounts for this type during the previous period of two to five weeks. This period depends on the production schedule as well as change in quality level, since the limits for expected variation are a function of the number inspected and a prescribed stability of the percentage defective. The control limits, upper and lower, are shown by dashed lines placed at 3 sigma units distance from the average. The sigma unit for the percentage chart is calculated from the formula $\sigma p=\sqrt{\bar{p}(1-\bar{p})/\bar{n}}$, where σp (the sigma unit) is the standard deviation of the average percentage of defects $\bar{p}$, and $\bar{n}$ is the average number of mounts inspected. For example, the limits for Fig. 2 are determined as follows: The type average $\bar{p}=4.2$ per cent and the average number of mounts inspected daily $\bar{n}=1,200$ are determined from previous records. Then $\sigma p=\sqrt{0.042\times0.958/1,200}=0.0057$ or 0.57 per cent; $3\ \sigma p=1.7$ per cent. Hence, the control limits are 4.2 per cent $\pm$ 1.7 per cent, or 2.5 per cent for the lower control limit (LCL) and 5.9 per cent for the upper control limit (UCL). The spread between limits represents the maximum variation expected from units producing mounts at the average level. Any daily percentage point above the UCL is circled in red to call attention to too many rejects. Likewise, any point below the LCL is circled in green to show better than average work. The 0 per cent points are starred in green as indicative of superior work.

Any noteworthy items that might aid or hinder mounting, thereby causing a decrease or an increase in the percentage of rejects, are marked on the chart at the proper date. Every out-of-control point is definitely explained and a continuous record is built up showing the difficulties encountered and the improvements made. Hence, a close check is kept on each operator, welding machine, and parts. As an illustration, it was noted that on one tube type open heater welds suddenly developed. The group leader traced the defects to a particular operator after it had been determined that the welder and heaters were all right. As a follow-up, a 100 per cent continuity check for heaters was made for a while. As a result of the careful instruction and follow-up by the group leader, this run of defects was practically eliminated as evidenced by subsequent points on the chart "in control."

Out-of-control points that lie near the control limits must be interpreted carefully because constant limits are used for convenience, although these limits are not exactly constant due to two major factors. First, a marked change in the number of mounts inspected will modify the control limits inversely with the square root of the number, as the formula shows. Second, we have noted a weekly cycle established over a year's use of the charts. This weekly cycle is roughly U-shaped, with the lowest percentage of mount rejects on Wednesday and the highest on Monday. The daily index, which represents the percentage ratio of rejects per average Monday, Tuesday, etc., to the weekly average, is: Monday, 106; Tuesday, 102; Wednesday, 93; Thursday, 98; Friday, 101. Since this cycle of rejects correlates inversely with the work-efficiency curves determined by time-study engineers, we feel that the day of the work week is also an important factor, and we modify our interpretation of the quality-control charts accordingly.

The chart also illustrates how a new level of quality is detected by the technique of runs. Beginning at the point indicated by the arrow as "new level" there are ten successive points under the Type Average, although only one falls below the LCL. The hypothesis that these figures are representative of the average level of 4.2 per cent rejects is untenable, since the chance of ten successive sample averages falling on the same side of center is $1/2^{10}$, or roughly 1 in 1,000. Thus, the assumption is made that a new quality level has been established about 3.4 per cent. The newly determined Type Average is 3.4 per cent and its 3-sigma deviation is approximately 1.6 per cent. Hence, the following control chart will have control limts of 3.4 per cent $\pm$ 1.6 per cent or LCL 1.8 per cent and UCL 5.0 per cent.

Efficiency of Mount Inspection

An excellent check on the efficiency of Mount Inspection is provided by the 100 per cent factory test made on finished tubes to eliminate inoperatives. Inoperatives consist mostly of two types of defects, shorted elements or open elements. It is recognized that a 20 per cent mount inspection will not disclose all of these defects, but neither will 100 per cent mount inspection eliminate 100 per cent of the defects. During a month's trial of 100 per cent mount inspection on a specific type, analysis of the shorts and opens found at both Mount Inspection and Factory Test showed that 100 per cent inspection of mounts was effective in removing 70 per cent of the shorts, 91 per cent of the opens, and 77 per cent of the inoperatives over-all.

These efficiency figures are very good for a visual inspection of the mounts in question. Further, they are in keeping with actual experience at mount inspection. Obviously, an open element can be observed more easily than a short, due to the very nature of mount structure which often prevents a complete visual examination of a cathode or grid enclosed within a plate.

Results From Mount Inspection

Over an eight months' period the relative percentage of defective mounts has dropped 31 per cent, almost one-third. *This is a real benefit, since the initial figure was not high by any means.* Further, a corresponding improvement of 29 per cent was noted at the exhaust-machine port check. These figures are well reflected in a 28.5 per cent reduction in mounting shrinkage for finished tubes over this period. Aside from the considerable tube savings, the intangible results of improved quality are extremely important. Reduced variation in production from mounting requires fewer changes at the exhaust machine and aging rack. Hence, better uniformity of tubes is obtained, assuring the factory of more production with less spoilage and the customer of high-quality tubes.

Quality Control of Finished Tubes

Test Classifications

Finished tubes naturally occupy the center of attention as the end-point of manufacturing operations. Packed production of excellent quality is desirable both from a manufacturing viewpoint and in keeping with the company policy to produce more goods for more people for less.

Three objectives of the quality-control program for finished tubes are dictated by practical requirements. The tubes must be unquestionably operative, must be interchangeable by type, and must evidence good workmanship. Translation of these requirements into characteristics for inspection gives a natural classification of test items. Generally, items for tube operation comprise the inoperatives and the major functional characteristic of the tube such as power output, or transconductance, etc. Tube interchangeability is effected by controlling the electrical test items for center and spread. Good workmanship is maintained by inspection for mechanical test items such as appearance of etch, solder, etc.

We use several quality-control techniques to control these different groups of test items. The "go—no-go" items of operation and the mechanical test items are controlled by sampling inspection by attributes; i.e., either it is or it is not conforming. From this type of test we obtain information about the per cent defective on operation and workmanship. The electrical characteristics are controlled by sampling inspection by variables which shows the central tendency and the spread of each electrical test item, as well as the per cent defective.

Attributes-Acceptance Test

Our sampling inspection by attributes follows standard quality-control practice for double sampling. In double-sampling procedure the results of the first sample indicate whether to accept or to reject the lot at this point, or to test a second sample. If a second sample is required, a decision for acceptance or rejection of the lot is made on the results of the combined samples. Example: An inspection of a production lot per tube type for a certain lot size is performed on a first sample of 150 tubes with acceptance number 3 (maximum of three allowable defectives) and a second sample, if necessary, of 300 tubes with acceptance number 9 for the combined samples. This is the so-called "normal" acceptance which will usually pass lots with less than 1 per cent defectives. This plan has an "average outgoing quality limit" (AOQL) of about 1.6 per cent which means that, with 100 per cent inspection of rejected lots, the average per cent of defectives in the outgoing product will not exceed 1.6 per cent approximately. If the average per cent defective exceeds 1.6 per cent for the last 1,000 tubes in cumulated first samples, the tube type represented is immediately put on "stricter" acceptance which has an AOQL of slightly less than 1 per cent. The stricter-acceptance plan uses the same sample sizes, but smaller acceptance numbers. Thus, for the same lot size as indicated in the previous example of normal acceptance, the stricter acceptance plan is: first sample, 150 tubes with acceptance number 2, and second sample, if necessary, of 300 tubes with acceptance number 4 for the combined samples. This type must remain on stricter acceptance until accumulated first samples totaling 2,000 tubes indicate a process average of 1 per cent or less.

By this method of normal-or-stricter acceptance, we get a twofold protection:

1. The manufacturing section is guarded against rejection of lots of better-than-average quality by a freak sample as long as the production level of quality is high.

2. The customer is protected against the acceptance of lots of worse-than-average quality by the operation of stricter acceptance as soon as production shows questionable quality. Continuation of the stricter plan further assures the customer of excellent quality while the product variation is being corrected.

AOQL Versus Process Average

Let us give some consideration to what the previous 1 per cent and 1.6 per cent AOQL values mean practically. The AOQL is the maximum value which the average quality after inspection can reach regardless of the percentage defective incoming, under a procedure which calls for 100 per cent inspection of rejected lots and elimination of all defects found. This maximum value occurs because more frequent rejections and consequent screening of low-quality lots force the over-all outgoing percentage defective down. Evidently, the production average of rejects must be maintained at a level much lower than the AOQL, in order to minimize rejections which disrupt manufacturing schedules. The process average is usually not more than one-third to one-half of the AOQL for efficient operation in the factory. This means that our tubes must run under $\frac{1}{2}$ per cent to $\frac{3}{4}$ per cent rejects under normal acceptance, and about $\frac{1}{3}$ per cent to $\frac{1}{2}$ per cent rejects under stricter acceptance. The $\frac{1}{2}$ per cent average reject figure is very low and practically can be considered as perfect quality, since even with the sensitive test sets and meters the trained technicians cannot test radio tubes continually with better than $99\frac{1}{2}$ per cent accuracy. Since this acceptance test by attributes holds the production average of defects to approximately the limit of test accuracy, we feel that our acceptance-test control method assures our customers of receiving tubes of very high quality.

Inspection by Variables

However, controlling the percentage of rejects, or tubes out of limits, is only part of the objective of our quality-control program. Tube interchangeability is equally important from the customer's viewpoint. This requires a control of the center and spread of every tube characteristic. Obviously, this is a big job, since some tube types—such as converters and duplex tubes—have over twenty electrical test items. For this purpose we have developed a method of control by variables applied to each characteristic. Much credit for the impetus and direction of this method is due to Walter Kirk, designing engineer, and to James Campbell, quality-control supervisor.

Quality tests for characteristics that are read as continuous variables give a much better picture of how production is going than corresponding tests on a "go—no-go" basis do. At the low level of percentage rejects found in quality-acceptance tests of our tubes, under 1 per cent, a single sample of about 100 tubes is required to test by attributes. However, at the same quality level, testing by variables will give the same required accuracy with less than one-fourth the number of tubes in the sample. The variables method makes use of a twenty-tube sample.

The variables control is a composite of the standard quality-control techniques of control chart and frequency distribution with the extra feature of graphic presentation for comparison of test items. Admittedly, this is a definite hybrid which escapes classification. For the purpose of reference, it will be called the "median-control method."

Stripped to its absolute essentials, the median-control method is a daily chart on which the distributions of readings on the twenty tubes for all test characteristics are plotted. The center value, or median, of each test in the sample is marked on the distribution. Control limits for the median of each test are predetermined and marked on the chart. Each median will fall within its control limits, providing the production is centered near bogie, the published value of the characteristic. This chart gives a clear picture which is readily interpreted in the light of day-to-day variation, and also shows the tendency for any distribution to wander off bogie seriously or to have excessive spread.

What It Is

An example of the median-control chart is given by Fig. 3. A special form is prepared for each tube type so that all test items are included and their scales along the abscissa are designated. The desired center values (bogies) for all tests are located along the long dashed line, one under another. The short dashed lines mark the control limits for the median of twenty measurements. Solid lines show the tolerance limits for individual tube measurements. Scale values are chosen as a compromise between meaningful intervals and sufficient spread to include tolerance limits. Unequal class intervals are used frequently to include the tolerance limits, and yet make the control interval sensitive. The precision of measurement is considered in selecting the class intervals.

The distribution is obtained by marking an x for each tube reading in the appropriate column. The resulting tabulation of twenty crosses for each test takes the form of a discontinuous frequency distribution for which the number of crosses in each column show the frequency of occurrence of the readings. Thus, the central tendency and spread of each test item are easy to read and to compare with bogie and with other test items.

Since this graphic tally automatically arranges the readings in order of magnitude. it becomes a simple matter to count to the middle x in the array. This middle value, between the tenth and eleventh marks, is the median of the twenty readings. The position of the median is shown by a large red dot placed between the tenth and eleventh crosses. Its approximate value can

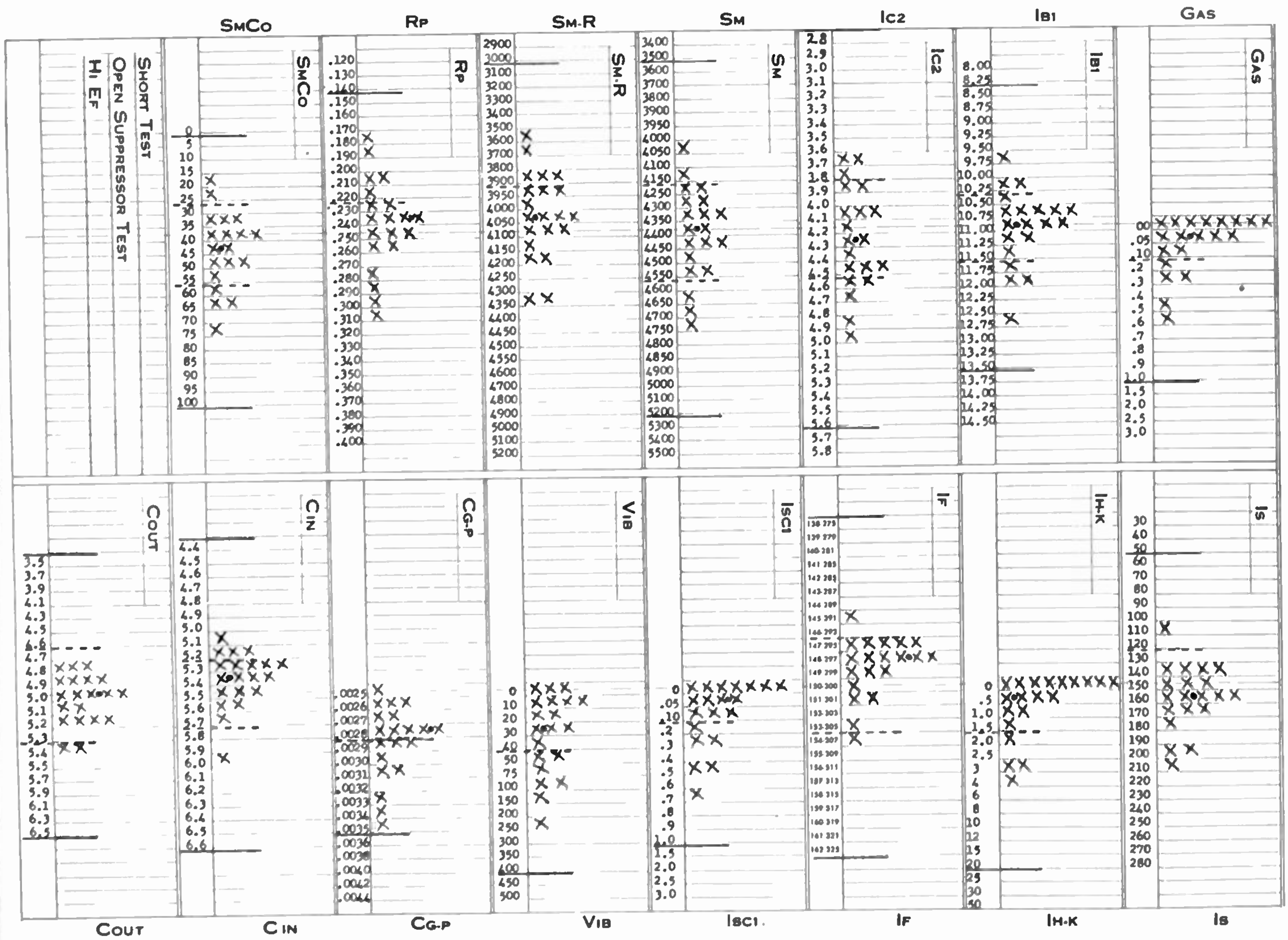

Fig. 3—Median control form.

be observed from the figure at the foot of the column in which it occurs. A more precise value of the median may be obtained by interpolation. The counting procedure locates the median very quickly and without any calculations. A median which falls outside the control limits is an excellent indicator that the production is not well centered. In this case, one of two courses of action is possible:

1. The distribution should be recentered for that characteristic.
2. The limits are not real, meaningful figures, and should be investigated for the purpose of changing.

Use of the median of the distribution is merely a test for stability of the process center, for efficient tube interchangeability.

Why Use The Median?

Analysis of considerable test data indicated that the median represents the center of production effectively for all electrical test items. The reasons for selecting the median rather than the average or the modal value as the indicator are:

The median is practically as good an estimate of central tendency as the average when the distribution is near-normal or approximately symmetric. It is easier for operators to determine, being located by counting rather than by computation. Since it is an "average of position" rather than a mathematical average, the median is not affected by extremely high or low readings which occur as mavericks. Finally, the median is more representative of the production center when the distribution is quite skewed or J-shaped.

Wherever the median of the sample occurs, we know that about 50 per cent of the production readings lie on either side of it, regardless of the shape of the distribution. The only disadvantage of the median is that theoretically it is not so sensitive an indicator as the average, since the median has a sampling variation about 21 per cent greater than that of the average, for samples of twenty.

Control Limits

Control limits which we use are boundaries for permissible variation of the median due jointly to sampling and slight product variations. Theoretically, the entire interval between control limits is due solely to sampling variation, which depends on the product variation, the number of tubes included in the sample, and the degree of assurance desired for control. The control limits we use include an interval which is 5 to 10 per cent wider than the theoretical one. This additional spread permits a small shift in the production center. These modified control limits are obtained by rounding-off upwards the actual standard deviation of each characteristic as much as 10 per cent, or by using an approximation equation for computing control limits.

The theoretical control limits are determined from the relation:

$$\text{Control Limits for Median (CLM)} = \text{Bogie} \pm 3(1.214)\sigma_x/\sqrt{20}, \tag{1}$$

where

Bogie is the desired (published) process center

3 is the number of sigma units

1.214 is the factor which accounts for the extra variation introduced by the median

20 is the sample size

σ_x is the standard deviation of the characteristic.

This relation reduces to

$$\text{CLM} = \text{Bogie} \pm 0.814\sigma_x. \tag{1a}$$

Modified control limits are computed from the approximation equation:

$$\text{CLM (adj.)} = \text{Bogie} \pm 0.1(\text{Maximum} - \text{Minimum}),\ \text{where Maximum and Minimum refer to specified tube limits.} \tag{2}$$

This approximation (2) is possible in our control program simply because a definite relation exists between the product standard deviation and the maximum and minimum test limits for individual tubes for most tests, excluding those giving J-shaped distributions. In general, the standard deviation is slightly less than $\frac{1}{8}$ of the spread between these limits. Actually, $8.6\sigma_x = \text{Max-Min}$ is the average relation found for various characteristics investigated. By using this relation in (1a), we obtain the exact equation, $\text{CLM} = \text{Bogie} \pm 0.0947\ (\text{Max-Min})$, which is given an adjustment of 5.6 per cent in spread to yield the approximation (2). This approximation equation is quite handy to use since it permits setting modified control limits directly from the published center and the tube limits. If a single-ended tube limit is specified, twice the half-spread is used in the approximation formula, giving control limits as follows:

$$\begin{aligned} \text{UCLM} &= \text{Bogie} + 0.2\ (\text{Max-Bogie}), \text{ and} \\ \text{LCLM} &= \text{Bogie} - 0.2\ (\text{Bogie-Min}) \end{aligned} \tag{2a}$$

As an example of the calculation of modified Median Control Limits for a normally distributed item, the plate current (Ib_1) test for type 6BA6 is illustrated by Fig. 4, showing distributions and control limits. Our published bogie is 11.0 ma and the JAN limits are 13.5 ma maximum and 8.5 ma minimum. The spread of tube limits is $13.5 - 8.5 = 5.0$. Hence, the variation allowed the median is $\pm 0.1 \times 5.0 = \pm 0.50$. Then the control limits are 11.0 ± 0.5, or 11.5 upper (UCLM) and 10.5 lower (LCLM). These control limits allow the median a variation of ± 4.5 per cent from bogie.

For comparison, the exact control limits are given here. The actual standard deviation σ_x was determined from actual readings to be 0.56 ma. Thus, from (1a), the control limits are 11.0 ± 0.46 ma. It is evident that the modified control limits give about 9 per cent more spread than exact control limits in this case.

The test for grid emission (Isc_1) results in a J-shaped distribution as shown on Fig. 4. The upper control limit was determined from readings taken on 400 tubes. These readings gave a median of 0.005 and a deviation

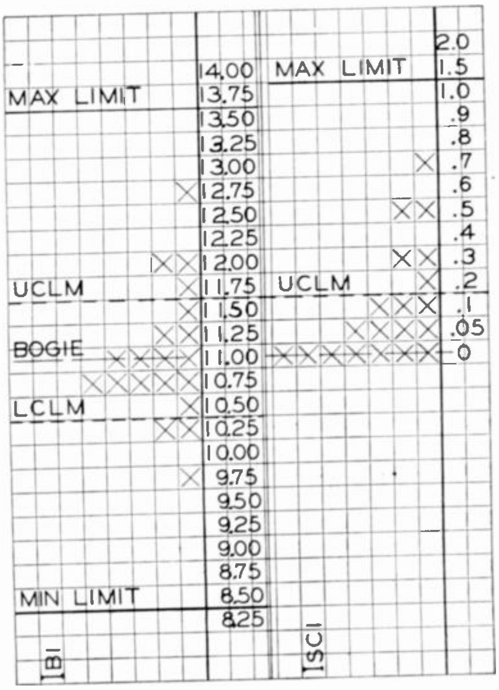

Fig. 4—Distributions and control limits.

of 0.109. Since bogie is 0, (1a) gives the control limit, ULCM $=0+0.814\ (0.109)=0.089$, which is used as 0.1 on the chart. The actual limits for the median are $0.005+0.089$, or 0 to 0.094, so the rounding off to 0.1 increases the actual interval about 6 per cent.

Distributions and Tolerances Related to Control Limits

All calculated control limits must be adjusted to fit the design of the chart. Adjustments are a practical necessity, since the continuous variables are plotted as discrete units. Note the unequal intervals used for plotting Isc_1 readings on Fig. 4.

The use of (1a), (2), or (2a) for calculating control limits is determined by the type of distribution for each test characteristic and the type of tolerance limits. Distributions usually are nearly normal, quite skewed, on J-shaped. Tolerance limits may be minimum only, maximum only, or both. Lower, upper, or both control limits are used, in keeping with the tolerance limits.

All test items having J-shaped distributions, such as the characteristics gas, $Ih-k$, Isc_1, and *Vib* on Fig. 3, have a maximum tolerance limit only since the desired center is zero. The "upper control limits for these items are determined from (1a).

All characteristics that have approximately normal distributions, such as *Ib*, *Ic2*, *Sm*, and *If* on Fig. 3, have both tolerance limits and bogie specified. For these tests the control limits are obtained by using (2).

The other characteristics on Fig. 3 have distributions that are skewed variously in combination with all three types of tolerance limits. Hence, control limits are computed for these tests by using either (1a), (2), or (2a) as experience indicates.

It should be remarked that (1a) can be used to compute control limits for test items having any type of distribution and tolerance limits. The approximation formulas are mentioned only to illustrate a short-cut in setting control limits based on a previous knowledge of the distributions in relation to tolerance limits. Although the exact formula for computation (1a) is completely general, the approximation method may not apply to a similar procedure by another manufacturer, unless his tolerance limits are known to be some fixed multiple of sigma units. We have found that it is better to use the exact expression for control limits on new installations of the method, and developed the shortcuts from subsequent experience with the procedure.

Control of Out-of-Limit Tubes

Centering all electrical characteristics by keeping the median within control limits holds the amount of scrapped production to a minimum, regardless of the type of distribution for any characteristic.

As previously stated, the specification tolerance is about ± 4.3 sigma units except for the zero-center test items. For test items which follow the normal distribution, the spread of ± 4.3 sigma units includes over 99.9 per cent of the product. For those characteristics which give asymmetric distributions the conditions are met for applying the Camp-Meidell inequality,[1] which shows that at least 97.6 per cent of the product will be within the ± 4.3 sigma units.

Those test items which have zero bogies give rise to the J-shaped distributions characterized by "die-away" curves. This type of distribution can be approximated closely by the probability curve $f(x)=1/\sigma e^{-x/\sigma}$. For this distribution, 98.6 per cent of the product is included between 0 and 4.3 sigma units.

It is evident that conformance of the median to the control limits in our procedure controls the out-of-limit tubes to not more than 2 per cent approximately, no matter what the nature of the distribution is.

Use

The information obtained from the median control chart is used correctively and preventatively to recenter an out-of-line test item, rather than to restrict the product on the basis of percentage of defects. Questionable lots are definitely indicated, and the next inspection station is advised in this case. Then decision concerning acceptance or rejection of the questionable lot is based on results of the attributes test made on a much larger sample. A questionable lot is used here as one whose sample shows the median out-of-control limits, but no tubes out of maximum or minimum limits.

As in regular control charts, trends of off-centeredness are indicated by runs of median points on successive charts above or below the center line. Interpretation of the charts by the technique of runs is most useful, since off-centeredness can be shown even when consecutive median points for any one test item fall within control

[1] $P(\pm t\sigma)>1-(1/2.25t^2)$.

limits. Two rule-of-thumb criteria used in connection with runs are:

1. Seven successive points on the same side of center give a strong warning that the characteristic is not centered.

2. Two successive points at the control limit on the same side of center give a very strong warning of an off-centered characteristic.

Since the chance of either one of these results happening if the product is well centered is small, the warning signs are quite forceful. Observance of their indications enables the engineers to anticipate and correct a shift in process centers before out-of-control points actually occur.

The main uses of the chart may be summed up as follows:

1. To provide an excellent visual record of how production is running for engineering and management information.

2. To point out uneconomic and unreal specification limits.

3. To indicate further action required on production represented by the sample.

4. To direct action to be taken for centering future production.

The feature of the median control chart in showing visually in advance *what* the coming change is, *which* way it is headed, and *when* it is expected to become serious makes this control method indispensable to management, engineering, quality, and inspection. Further, the results of finished-tubes inspection take on a double significance, being useful not only as a record of how quality has been previously, but also as a barometer to forecast future quality.

Keynote for Establishing Good Quality-Control Methods

In each one of the control methods established we have provided for assurance of quality by representative samples and clear-cut inspection methods, as well as for action through definite allocation of duties and responsibilities. Taking a cue from advertising men, who sum up their objective by the letters A, I, D, A, our objective for sound quality-control procedure might be designated by S, I, D, A. The letters have the following significance:

S for Sampling, which must be random and representative of the product;

I for Inspection, which must be so designed and defined that it will consistently indicate good, mediocre, and inferior quality, not merely sort good from very bad;

D for Disposition of the production represented by the sample, thereby providing an excellent spot for cost reduction in shrinkage, reprocessing, and reinspection; and

A for Action to be taken with regard to the quality of future production as indicated by the quality barometer.

By following this compelling directive, we expect continually to provide our customers with a high-quality product.

A Field Survey of Television Channel 5 Propagation of New York Metropolitan Area*

THOMAS T. GOLDSMITH, Jr.†, FELLOW, IRE, R. P. WAKEMAN†, MEMBER, IRE, AND J. D. O'NEILL†

Summary—**A comprehensive study of the performance characteristics of Du Mont television station WABD, New York, N. Y., embracing a new measuring technique, is discussed. A comparison of theoretical and experimental data is illustrated by photographs and charts indicating receiving conditions within the service area. Pertinent information concerning various interference problems is also considered.**

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† Allen B. DuMont Laboratories, Inc., Passaic, N. J.

Introduction

This paper covers field measurements made during the period of May through July, 1947, to determine the performance characteristics of television station WABD.

It was felt that a large quantity of spot measurements at an antenna height of 30 feet, when plotted on a true profile of the terrain between receiver and transmitter, would add much general information regarding television transmitting and receiving. It was also decided that pictures taken directly from a tube on a conventional television set would materially aid the interpretation of the results obtained.

On the other hand, the popular mobile method of observing field strength does not take full consideration of the three factors of receiving antenna height, terrain vagaries as contrasted with smooth-earth considerations, and the photographic records obtainable at fixed receiving installations. We feel that the cluster method of observation takes good cognizance of these variables.

This report also covers a series of airplane measurements made at radii of approximately 10 and 20 miles from television station WABD. Both sound and picture were measured, and a polar plot of the results obtained on the 20-mile-radius circles is shown in Fig. 5.

Equipment

1. *Transmitting Equipment*

(a) Television Transmitter: 515 Madison Ave., New York, N. Y., North Latitude, 40° 45′ 33″; West Longitude, 73° 58′ 25″.

(b) Transmitter Power Output: 2.5 kw (audio), 4 kw (video) peak.

(c) Antenna Power Gain: 3.96.

(d) Effective Radiated Power: 9.45 kw (audio); 14.25 kw (video) peak.

(e) Effective Antenna Height: 640 feet (average for 9 radials).

(f) Modulation: Video modulation test pattern and program.

2. *Receiving Equipment*

a. *Surface Measurements:*

(1) U. S. Army Signal Corps truck (6×6 feet).
(2) Type PE-197 generator and trailer.
(3) Model M-58 Measurements Corporation field-strength meter complete with standard dipole antenna.
(4) Hallicrafters S-36 receiver.
(5) Ferris Model 18B signal generator.
(6) Du Mont Type 72 dual-band antenna (height 30 feet).
(7) Du Mont Clifton receiver.
(8) 40 feet of 72-ohm dual coaxial transmission line.
(9) Exacta camera and tripod.
(10) 30-foot collapsible antenna mast.

Method of Measurements

At approximately twenty cluster points along one radial, voltage measurements were made on both a 12-foot antenna and a 30-foot antenna. The total number of measurements was in the neighborhood of sixty, and an examination of the data indicated that there was no reasonable conversion factor relating the two heights. The ratio of the voltage on the 30-foot antenna to that on the 12-foot antenna varied from less than 1 to as high as 8. It was therefore decided that all measurements should be made on the 30-foot antenna, as this is the approximate height of a home television receiving-antenna installation.

The selection of a 30-foot receiving height made mobile measurements impractical. Therefore, the cluster method was adopted and the measured points were selected to lie exactly on radials emanating from WABD. This selection of points of measurement also permits a correlation between any point and the entire profile from the transmitter to that point. This advantage is not realized when mobile measurements are made using existing roads, since it is impossible to make measurements along a true radial and each measured point lies on a different azimuth and consequently on a different profile.

Nine radials were laid out with WABD as a center, these radials so chosen as to pass through the largest centers of population and at the same time to be reasonably equally spaced. Along each radial approximately fifty cluster points were marked at convenient intersections of roads with the radial. The spacing of the cluster points varied from one-half mile in and around New York to two or three miles out on the ends of the radials. Measurements were carried out along each radial to a point where the received signal was insufficient to produce a picture. A minimum of three measurements was made at each cluster within an area having a radius of 100 feet. At each cluster, a photograph was made of the picture from the cathode-ray tube of the Du Mont Clifton model receiver, to provide a record of received picture quality and the presence of ghosts or other interference.

All measurements included in this report were made using the Measurements Corporation M-58 field-strength meter and a Du Mont Type 72 dual-band antenna at a height of 30 feet. The field-strength meter was calibrated by the Measurements Corporation prior to, and again during, this survey. In addition, as a further calibration check on the M-58, field-strength measurements were made at frequent intervals using the S-36 Hallicrafters receiver and the Ferris 18B microvolter. At points where the field strength exceeded 0.1 volt, an additional attenuator, calibrated at the frequency being measured, was placed in series with the antenna.

An analysis has been made of the operating log of station WABD during the period of these field measurements. It was found that the radiated power was substantially constant. The stability of the transmitter was further confirmed by means of a continuous tape recording of radiated signal as received at the Du Mont Cedar Grove field laboratory. An analysis of the recording tapes at this location also shows that the field strength remained constant within 10 per cent.

Description of Presentation of Data

Fig. 1 is a comprehensive map showing:

1. *Radials.* The radials are drawn emanating from WABD showing their true azimuths and the locations of identifying towns.

2. *Theoretical 5 mv/m and 0.5 mv/m contours.* These contours were obtained by calculating the 5 mv/m and 0.5 mv/m distances along each radial from K. A. Norton's curves and connecting these points with a suitable closed curve.

3. *Random sample of measured data.* Since it was impossible to put all measured values on this small map, the average value obtained at every fifth cluster along each radial was selected. As this sample is small, and the selection of every fifth cluster was purely arbitrary, these data do not necessarily indicate probable values in a large area about the selected points.

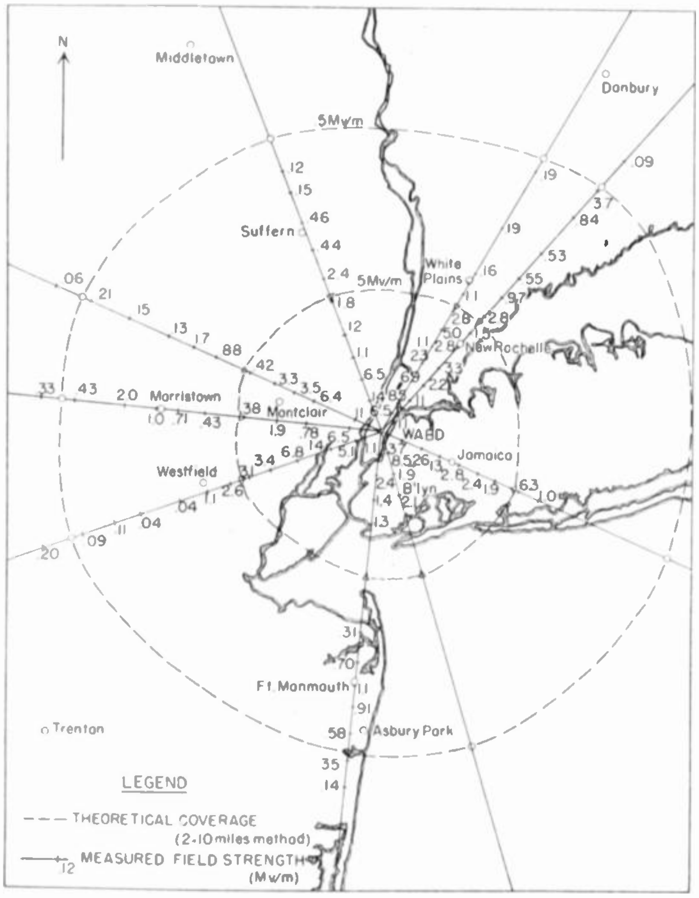

Fig. 1—WABD field test, August, 1947. Comprehensive picture of survey and results

4. *Statistical indication of measurements within each contour.* These data are not included in this report but are directly related to it. The entire data were analyzed and the percentage of the total points (within the theoretical 5 mv/m contour) was ascertained. The total number of points within this contour was found to be 725, of which 302 exceeded 5 mv/m. Thus our measurements indicate that the field strength is greater than 5 mv/m for only 42 per cent of this area. The same computations were made for the 0.5 mv/m contour. This area included 1,269 measured points, of which 1,014 exceed 0.5 mv/m. We therefore found that 80 per cent of this area could expect a field strength of greater than 0.5 mv/m. It is obvious from this that the more distant measurements come much closer to predicted theoretical values than do those in the immediate vicinity of New York City. It is probable that the primary cause of this is the large number of high buildings in the vicinity of the transmitter.

Figs. 2 and 3 show a detailed study of two of the radials. These two were chosen for illustration and discussion because they represent the extremes in smooth and irregular terrain. The abscissas of these graphs indicate the distance from WABD, and the two ordinates indicate the ground elevation and signal strength. The lower portion of the graph is, therefore, a profile of the radial on a true-earth's-radius basis but with a height versus distance ratio of 50 to 1. Cluster points are shown at their actual location (elevation and distance) along each profile. Directly above each cluster at the top of the graph, the several values of field strength measured at that cluster are indicated. A vertical line joins the extreme values, and, since the ordinate is logarithmic, the length of this line (regardless of the absolute value) is an indication of the percentage variation of measurements in the cluster. Since a field which varies widely over small distances is usually a result of multipath condition, the length of this line is also indicative of the probability of "ghosty" reception. A smooth curve was drawn through the arithmetic mean of the values measured at each cluster. Because all of the clusters lie on the same radial, this curve is a true indication of the field strength along the radial. In addition to this experimental curve, a theoretical curve based on K. A. Norton's curves utilizing the average elevation from two to ten miles from the transmitter is also shown.

As previously noted, photographs were taken at each of the cluster points. It was obviously impossible to include fifty photographs on each one of these sheets, and so a few representative photographs taken on each radial were selected for reproduction. The cluster point at which each photograph was taken is indicated above the picture.

Fig. 4 shows the results obtained from the three low-band New York television transmitters along the Montclair radial. Experimental curves for WNBT and WCBS-TV are incomplete, since these stations were not on the air at all times when measurements were being made. The effective antenna heights and effective radiated powers of the three stations are shown in Table I.

TABLE I

	Antenna Effective Height	Effective Radiated Power
WABD	640 feet	14.25 kw (video)
WNBT	1,241 feet	6.6 kw (video)
WCBS-TV	811 feet	*5.5 kw (video)

* The CBS radiating system was subsequently modified.

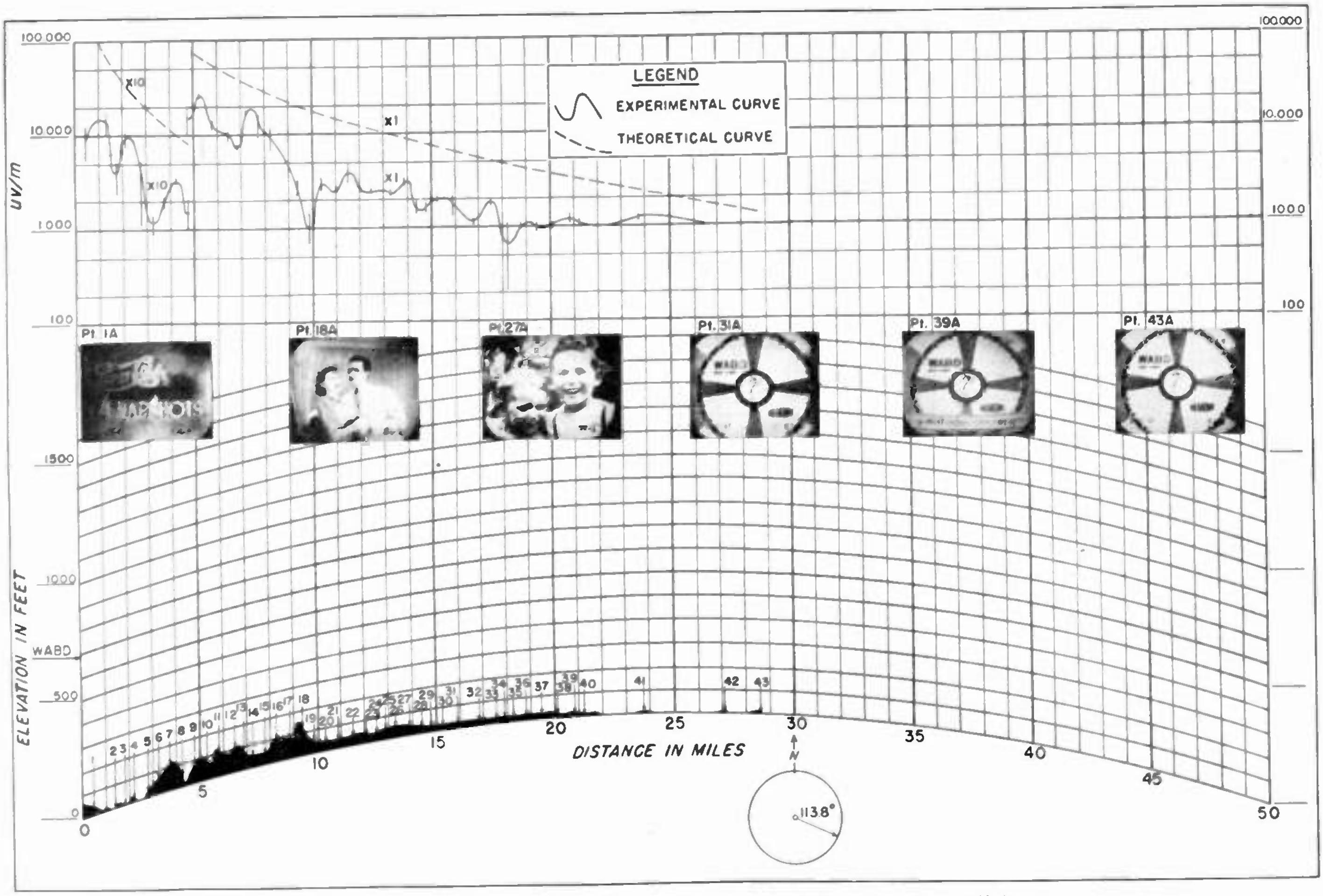

Fig. 2—Profile and field strength versus distance curves, Jamaica radial.

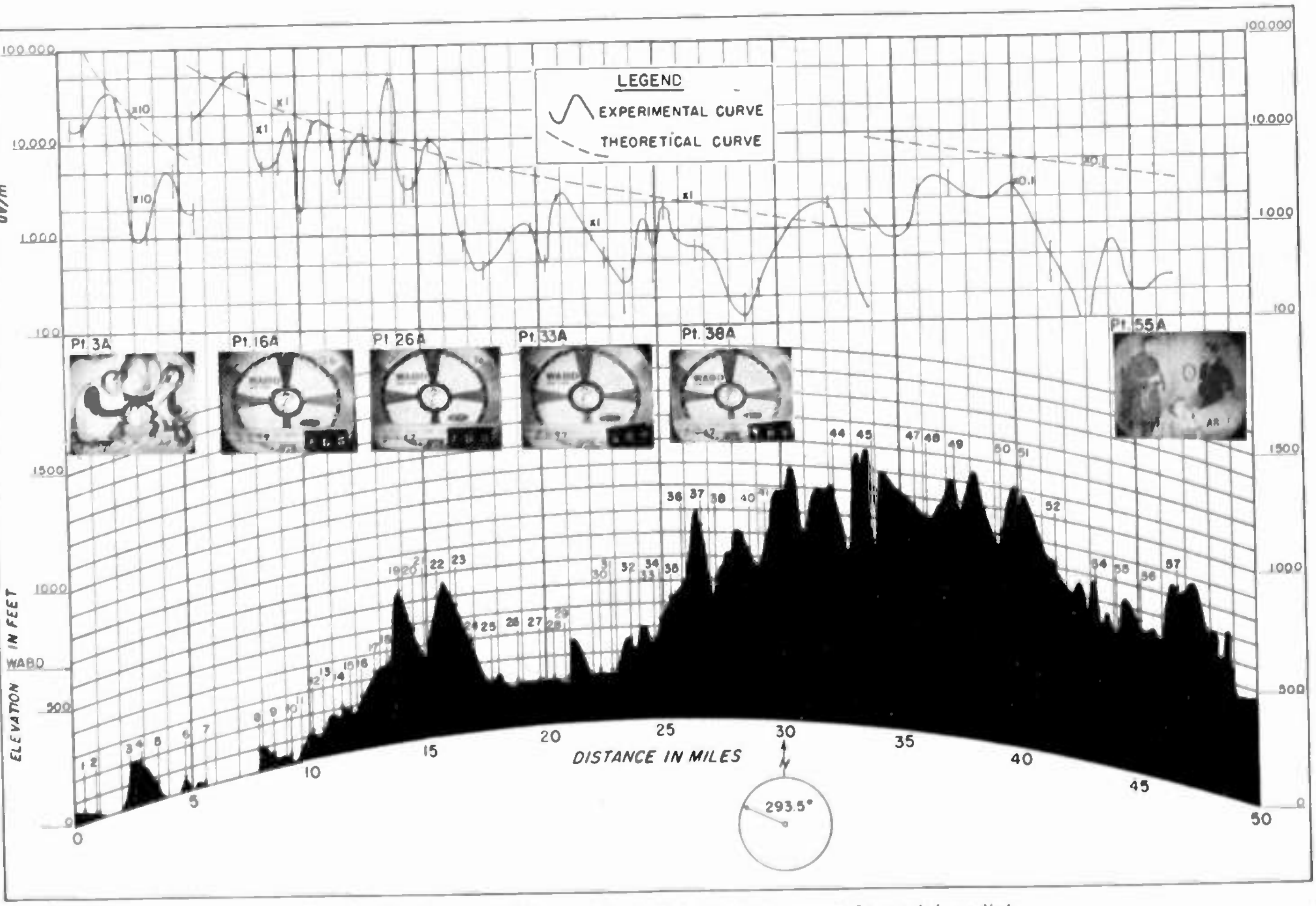

Fig. 3—Profile and field strength versus distance curves, Montclair radial.

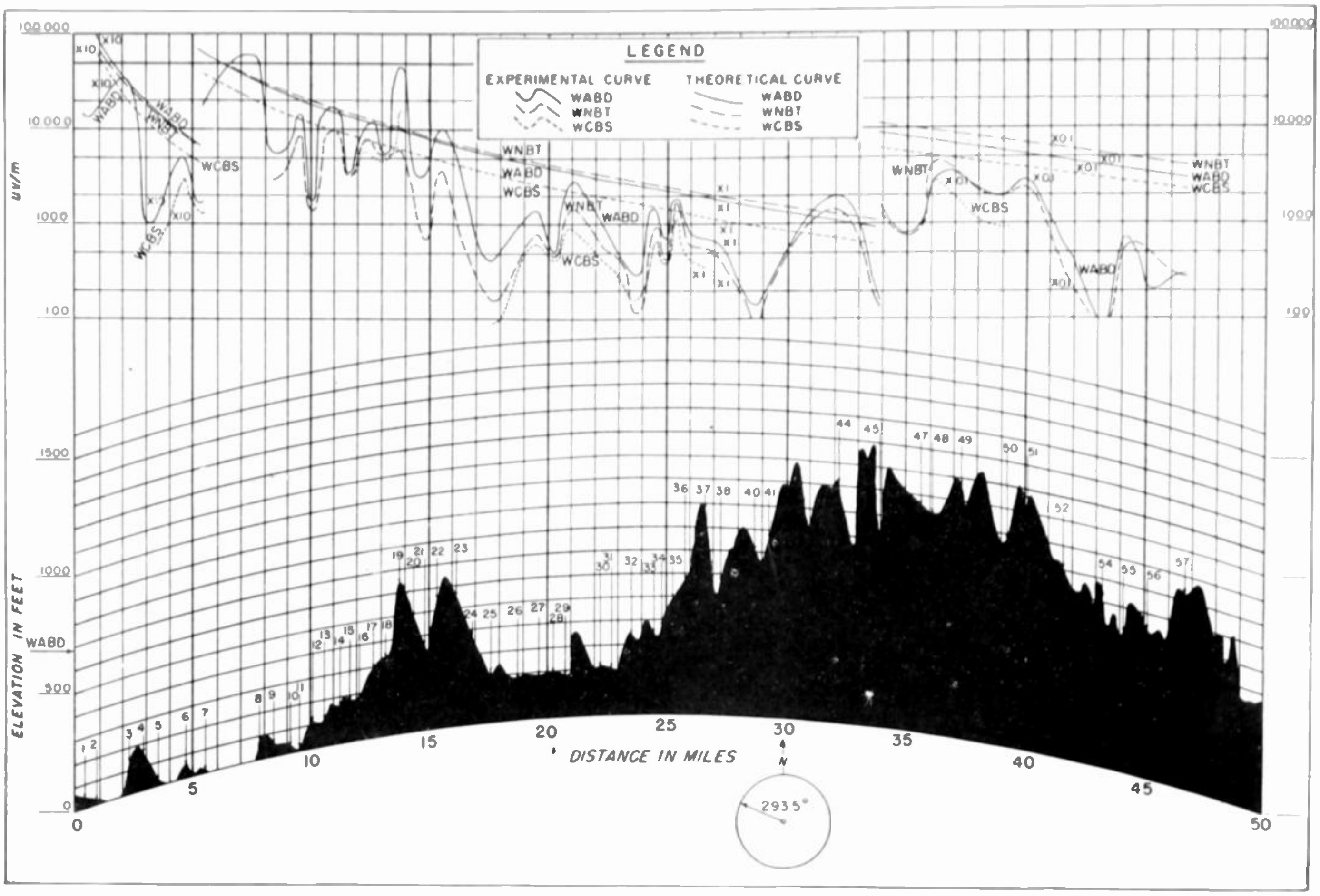

Fig. 4—Profile and field strength versus distance curves for television stations WABD, WNBT, and WCBS-TV, Montclair radial.

A brief discussion of some of the more pertinent points brought out by Figs. 2 and 3 follows.

In general, the variation of the field strength follows that of the profile quite well. The few exceptions were usually traceable to local obstructions or reflecting surfaces.

It will be observed that the experimental curve lies generally below the theoretical curve on all radials. The greatest deviation from the theoretical curve occurs within an area of a few miles from the transmitter, with the discrepancy decreasing with increasing distance from the transmitter. In particular, measurements on the streets in New York City are far below the theoretical values. Furthermore, in this same area, the measurements show a very great variation in signal strength, although the profile appears to be essentially flat. Both of these effects are due to the shadows and multipath reflections caused by the many high buildings.

An excellent example of multipath conditions appears on S-5, R-11, the Jamaica radial. It will be observed that, from about cluster 20 to cluster 43, the terrain is quite flat. Inspection of the plotted data shows the spread of measurement of each cluster to be very small, with the exception of cluster 35. At this point, the ratio of the maximum to minimum measured values is about five. When the received picture was photographed, it was noted that a very bad ghost condition prevailed. Upon looking around we discovered a large steel gas tank about 100 feet high and less than 500 feet from the truck. This was an extreme condition, but, in general, a large spread of measurements at a given cluster indicates bad multipath conditions.

In examining the pictures reproduced in Figs. 2 and 3, it should be realized that in most cases the picture appearing on the receiver screen was of considerably better quality than here illustrated. The original photographs themselves left much to be desired in the way of photography. At times, specular reflections were permitted to reach the camera from the cathode-ray-tube screen; the receiver was not always properly adjusted; photographs were sometimes made at an instant when noise conditions were very bad; and, in some cases, the developing and printing were at fault. In addition to these errors in photography, the reproduction and reduction in the printing process has resulted in further degradation of the pictures.

Aircraft Measurements

Fig. 5 is a polar plot showing signal strength versus azimuth as measured in an airplane. The plane used for these tests was a U. S. Army Air Corps converted B-25 medium bomber. Several antennas were permanently mounted on the plane, including a 75-Mc beacon antenna, under the nose of the ship. This antenna was used in conjunction with the M-58 field-intensity meter to obtain the data plotted in Fig. 5. A 6-volt storage battery was used to operate the field-intensity meter, since the airplane was not equipped with 110-volt, 60-cps power. A second antenna extending from the center of the fuselage to the tip of one fin was employed to feed a Du Mont Model 180 television receiver, operated on 115-volt, 400-cps power. Still photographs and motion pictures of the picture tube were taken on several flights.

The data for Fig. 5 were taken with the plane flying at an altitude of 2,000 feet in an approximate circle having a radius of 20 miles with WABD as its center.

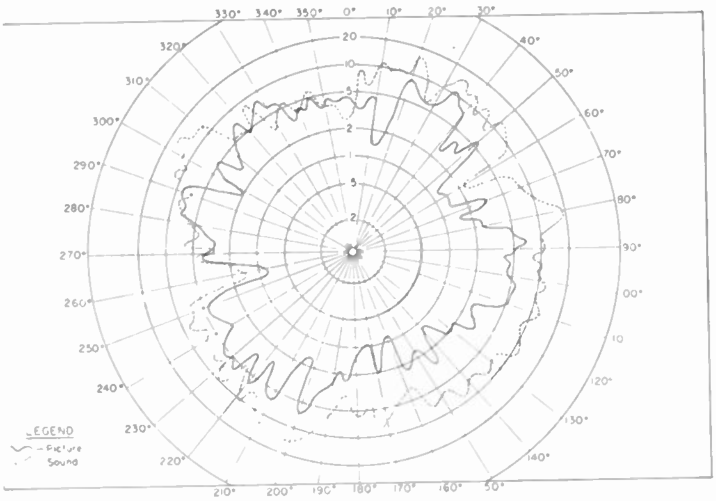

Fig. 5—Aircraft measurements, field strength (mv) versus azimuth, 20-mile radius, 2,000-foot elevation.

Two circular flights having this same radius were made with the M-58 field-strength meter tuned first to 81.75 Mc (sound) and then to 77.25 Mc (picture). The same types of measurements were made at a radius of approximately 10 miles, but the results are not shown here.

An examination of the data from about ten such circular flights showed very little correlation either among themselves or in relation to any known physical conditions of the New York area. A theoretical investigation was made to ascertain the effect on the field strength of deviations from a truly circular flight at constant altitude. It was found that a vertical variation of 250 feet at 10 miles was sufficient to change the path difference between the direct wave and the ground-reflected wave by one-half wavelength. At 20 miles, a vertical change of about 400 feet was required to produce this same change in path difference. Deviations in a 10-mile radius of $1\frac{1}{4}$ miles or in a 20-mile radius of 5 miles would also produce a change in the path difference of one-half wavelength. Flight variation of this order would permit the direct wave and the ground-reflected wave to run the gamut from complete phase addition to complete phase subtraction. It is known that the variations in the actual flights were of this order. Consequently, the data cannot be construed to indicate the actual radiation pattern of the transmitting antenna. In the event that more perfect instruments were used to insure a more nearly circular flight at constant altitude (as, for example, an IFF set with the interrogator in the plane and the transponder at the transmitting site), other factors would still make the interpretation of data obtained in this fashion highly questionable. Probably the most important of these is the variation in elevation of the reflecting point of the ground-reflected wave. As previously discussed, a flight elevation variation of 250 feet is sufficient to change the path difference by one-half, wavelength. A similar variation in the elevation of the reflecting point will have approximately the same effect. Another important factor is the variation of the reflection coefficient at the reflecting point. Any circular flight about New York City will result in the reflecting point occurring sometimes on water, sometimes on sandy soil, and sometimes in highly industrialized areas. These several conditions affect both the magnitude and the phase of the reflected signal, and consequently the magnitude of the resulting signal at the point of reception. Other factors which tend to make measurements of this type of questionable value are:

a. Conditions where several reflected signals arrive at the airplane.

b. Instantaneous orientation of the plane, with a resulting change in the receiving antenna characteristics.

c. Shadows from high buildings in the vicinity of the transmitter.

In conclusion, it should be pointed out that the only consistent effect found on all circular flights was an appreciable reduction of field strength at approximately 260°. This same effect occurs on ground measurements, and has been traced to the broad shadow caused by the RCA building.

Subsequent Analyses of Data Taken on Radial Measurement

Several statistical studies of these data have been made in an attempt to make possible the prediction of television coverage on a probability basis.

As previously noted, a study of sixty-three cases in which the field was measured at 12 feet and 30 feet yielded no constant conversion factor which would warrant the extrapolation to 30 feet of field strengths measured at 12 feet. However, the curves on Fig. 6 show that the data can be plotted in a manner which can be readily assimilated. On this graph the ordinate is the ratio of signal strength on the high antenna to that on the low one, and the abscissa is the percentage of points at which the indicated value of the ordinate was exceeded.

The curve labeled "channels 2, 4, 5" was plotted from data obtained in this survey, and that labeled "channels 7, 11" from data obtained from sixty-six measurements carried out by John Dreyer of "Dreyer Surveys, Inc." The height ratio for the low-frequency channels was 30 to 12 feet, or 2.5:1, whereas this ratio was 30 to 10 feet or 3:1 for the high-frequency channels.

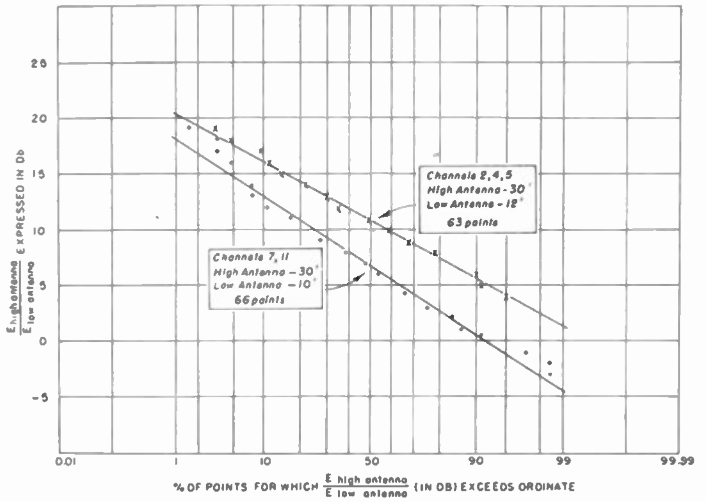

Fig. 6—Receiving antenna height-gain study.

The approximately straight-line nature of the two curves indicates a "normal" distribution of the ratios. The intersection of the curves with the 50 per cent line is, of course, the median value of the ratios, and the slope of the curve indicates the probability of a given measurement falling in the vicinity of this median. The median value for the low-frequency channels is 11 db or a ratio of 3.5. As the height ratio was only 2.5, this indicates an *average* improvement with height somewhat better than that predicted by a linear relationship. This is probably a result of getting the antenna above local obstacles. The slope of the line from 5 to 95 per cent is approximately 14 db, indicating that 90 per cent of all such measurements will lie within ±7 db of the median value.

The curve for the high-frequency channels shows a lower median value and a greater slope, thereby indicating less improvement with antenna height and a greater range of ratios.

Similar measurements made in the vicinity of 500 Mc show an increased continuance of this trend.

A second statistical analysis of these data was made with a view toward obtaining a factor to be used to modify coverage predictions based on a smooth-earth theory, for conditions where "smooth earth" does not properly describe the terrain.

The ratio of measured field strength to predicted smooth-earth ground-wave field strength was obtained for each point along a given radial. These ratios were then converted to decibels and plotted on "normal distribution" graph paper. This procedure was carried out for each radial, and the results for the Montclair and Jamaica radials are shown on Fig. 7. The horizontal line marked 0 db indicates the theoretical smooth-earth ground-wave field strength, and it is immediately obvious that the great majority of points measured lies below this line. It is also significant that the distribution is approximately a "normal" one, as indicated by the straight-line nature of the plot on this paper. The slope of the straight line indicates the variation of measured field strengths and its value expressed in decibels as measured between the 1 and 99 per cent points could be defined as the distribution ratio for the radial. Obviously, this figure is related to the roughness of the terrain, i.e., its deviation from smooth-earth conditions. Several schemes for defining the roughness of the terrain quantitatively were considered. It was decided that a significant description of the terrain might be given by the quotient obtained by dividing the length of the line representing the profile of the radial by the line representing the smooth-earth length of the radial. Our profiles were plotted on paper having a height-to-distance factor of 50, and this quotient varied from approximately 1.1 to 2.5 for the nine radials. Using the above definitions, it is to be expected that the distribution ratio would increase with an increasing terrain quotient. That this is the case may be readily verified by noting the slopes of the two curves on Fig. 7. As defined, the Jamaica radial has a terrain quotient of 1.126 and a distribution ratio of 21.9 db, whereas the Montclair radial has a terrain

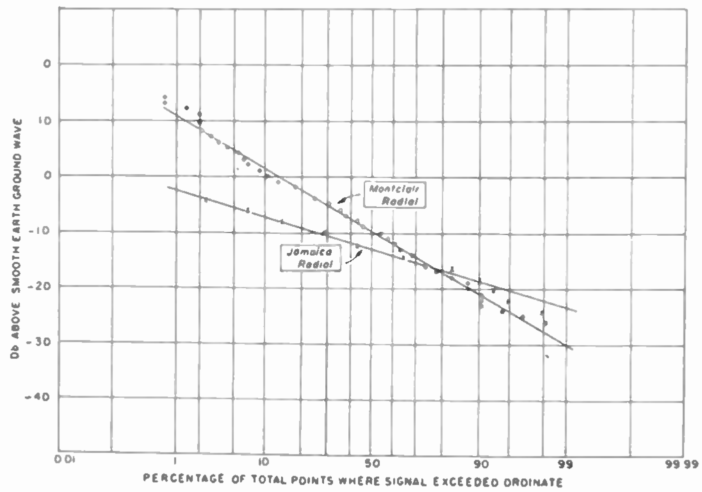

Fig. 7—Distribution of fields within short sectors along two radials.

quotient of 2.571 and a distribution ratio of 40.8 db. Most of the other radials fell into the appropriate positions indicated by their terrain quotients. A notable exception was the radial down through Brooklyn, which had a distribution ratio much too large for the measured terrain quotient. This was due, of course, to the fact that the profile, and consequently the measured terrain quotient, does not take into account the heavily built-up area existing throughout the length of this radial.

It was not possible to obtain any correlation between the terrain quotients and the median values of the distribution curves. This is not too surprising, since the illuminated slopes tend to increase the median value, while the shadowed slopes tend to decrease it, thereby increasing the slope of the distribution curve, but not necessarily changing its median value. This tends to indicate that it might be quite inaccurate to assign terrain factors to various types of terrain with the intent that the "smooth-earth ground-wave curves" be multiplied by these factors to obtain a closer approximation to actual conditions. However, the fact that the distribution ratio (the slope of the curve) does very definitely increase with an increasing roughness of terrain should be taken into account in predicting coverage. For example, the coverage area can be calculated based on the FCC smooth-earth ground-wave curves, and the percentage of the population within this area which will receive service then estimated from the roughness of the terrain. Over very flat earth this figure might approach 100 per cent, whereas over very hilly terrain it might be as low as 60 per cent.

The final analysis was made in an attempt to predict the probable signal strength existing in a deeply shadowed area.

From simple diffraction theory, without considering the ground-reflected wave on the receiver side of the "knife edge" hill, the curve shown in Fig. 8 may be derived. The ordinate is the predicted signal strength behind a hill expressed as a decimal fraction of the predicted signal strength in the absence of the hill. The abscissa is a parameter V which is a function of the distance from the hill, the distance below "line of sight," and the frequency, as shown in the sketch. An increasing value of V indicates deeper shadowing and consequently less signal.

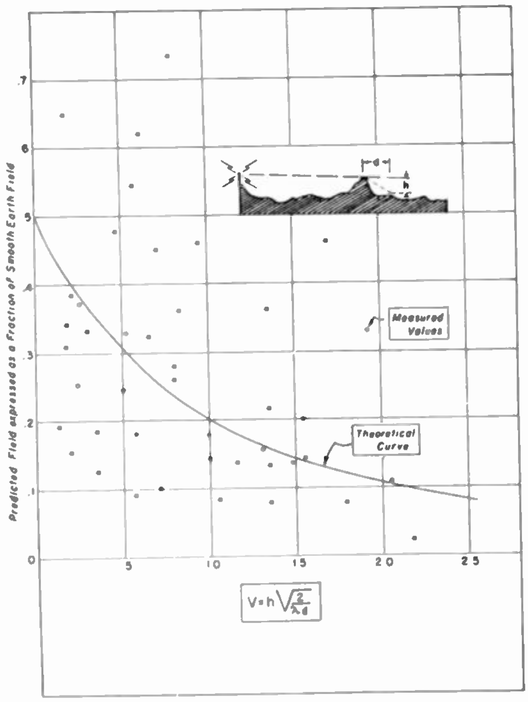

Fig. 8—Ratio of measured signal-to-ground wave predicted signal for 43 points lying in deeply shadowed areas.

Measurements of h and d were made for all deeply shadowed points on all nine radials, and the ratios of measured values to smooth-earth predicted values were computed. These points were then plotted on Fig. 8 at their appropriate locations. Next, the ratio in decibels of the measured value to the value predicted by the curve in Fig. 8 was obtained for each of the forty-three points, and these values were plotted on probability paper. The resulting curve is shown in Fig. 9 and is very

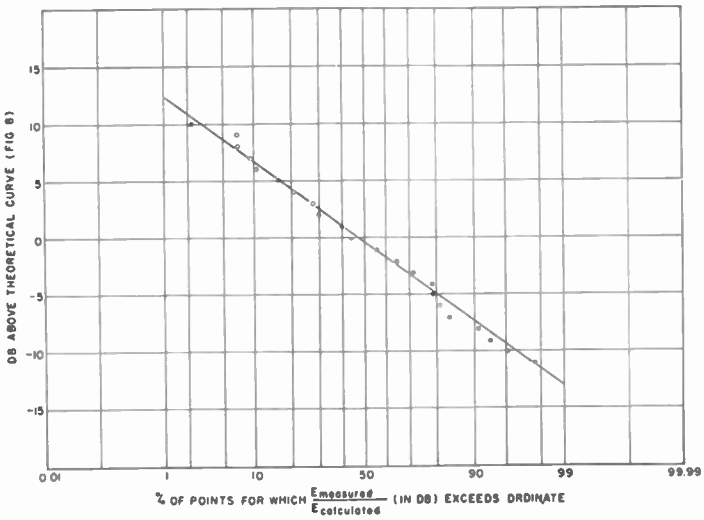

Fig. 9—Distribution of measured diffracted values about theoretical curve shown in Fig. 8.

interesting. The straight line indicates that the points on Fig. 8 are distributed about the theoretical curve in a "normal" manner, and the intersection of the curve with the 0 db and 50 per cent co-ordinates shows that, for this set of data, at least, the theoretical curve used is an excellent "average." The slope of the curve suggests that this prediction will be within ±10 db for 90 per cent of all points. In view of the approximations used in obtaining the curve in Fig. 8, and the neglect of such factors as reflection and refraction, the excellent correlation between theory and measurement illustrated in Fig. 9 is almost certainly due in part to a particularly fortuitous set of data.

Electronic Classifying, Cataloging, and Counting Systems*

J. HOWARD PARSONS†, MEMBER, IRE

Summary—**The determination of the distribution of a series of physical events according to magnitude is important in the study of the associated physical laws. Previous methods of determining this magnitude distribution were slow and cumbersome. Three new electronic systems which operate on events at a very high rate have been developed by the author while at Oak Ridge National Laboratory, and these are described.**

The analyzers can be used to determine the magnitude distribution of any series of physical events, if the characteristic under observation can be translated into a proportional voltage pulse. Two applications are discussed, and the advantages of the analyzers over other systems are shown.

Introduction

In the physical sciences it is often necessary to determine the magnitude distribution of a series of events. Each event in a series of physical events will occur with a magnitude that is controlled by the operating physical laws. A physical event can be defined as the occurrence of any phenomenon, and the magnitude of the event as a quantitative measure of a variable such as size, brightness, or duration. A series of events is characterized (in part) by the number of events in each increment of magnitude throughout the spectrum of magnitudes, and this distribution is the magnitude distribution. The physical laws that control a series of events can be studied under certain conditions if the magnitude distribution is known. For example, the laws governing grain size in a photographic film may be studied if the distribution of grain size produced under varying conditions can be obtained. The magnitude distribution of a series of events may be obtained by classifying each event as to size, and cataloging and counting the number of events in each increment of magnitude.

Such a determination, if performed manually, will be slow and have a low probability of being a true representation of the magnitude distribution of the series of events. These determinations are statistical in nature and involve large numbers of events each of which must be operated upon by the observer. The observer can only classify an event within a finite increment of magnitude; therefore, the determination is only approximate. Since manual determination is slow and inaccurate, and since there are series of events in which the events occur too rapidly for manual operation, automatic systems are required.

Electronic systems of classifying, cataloging, and counting (Fig. 1) have been developed that will operate on events at very high rates. Distribution studies may be made on any series of events by electronic means if the magnitude of each event can be translated into a proportional voltage pulse. The method of translation will depend on the type of event under observation, but a photoelectric scanning system might be a common one. In addition to this translator, the systems involve a discriminator that will distinguish between true observations and interfering effects, a classifier that will measure the representative pulse voltage, and a cataloger and counter that will record the data in tabular or graphical form.

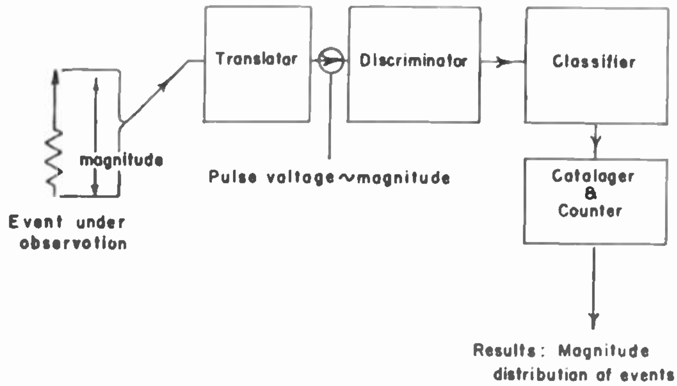

Fig. 1—Block diagram of the general method of obtaining magnitude distribution by electronic means.

The Determination of Energy Distribution of Alpha Particles

The study on which this paper is based was concerned with the measurement of energy releases during nuclear changes. In particular, we were interested in the distribution of kinetic energies of alpha particles as they were released from the nuclei of certain radioactive isotopes. The translation of the energies of the alpha particles into proportional voltage pulses was accomplished in an ionization chamber (Fig. 2). A sample containing a large number of atoms of a particular isotope was placed within this chamber. Alpha particles were continuously emitted from the sample in a fashion random with respect to time. Each alpha particle dissipated its kinetic energy in ionizing a part of the gas within the chamber, and the number of resulting ions was proportional to the energy of that alpha particle. A voltage change which was proportional to the energy of the alpha particle was produced on a collecting electrode within the chamber due to the collection of the ions produced by each alpha particle. Since each pulse represented a separate alpha particle, the pulses were distributed in time in a random fashion. Since these pulses were about 1-millivolt voltage transients, they were amplified to a conveniently measurable voltage. This

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† University of California, Los Alamos Scientific Laboratory, Los Alamos, N. M.

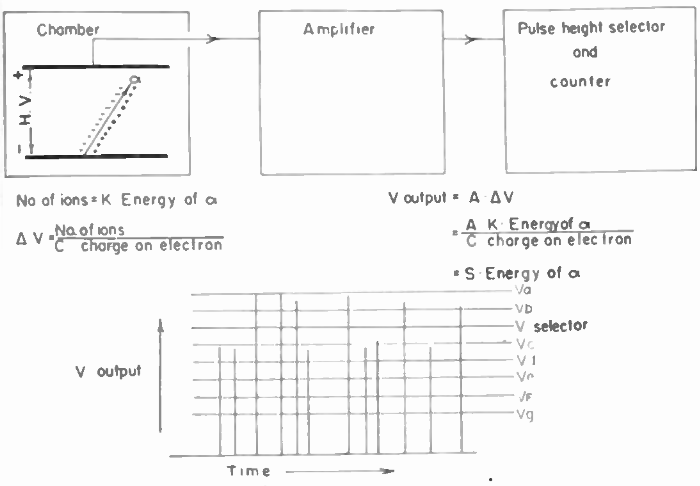

Fig. 2—The primitive electronic system for determining the energy distribution of alpha particles.

was accomplished by means of a low-distortion video-type amplifier of conventional design. A determination of the magnitude distribution of the representative pulses will be a determination of the magnitude (energy) distribution of the alpha particles.

Older methods for determining the voltage distribution of these pulses utilized a simple pulse-height selector and pulse counter. This combination would count all pulses whose peak voltage was above a certain preset value. A determination was made by setting the acceptance voltage of the pulse-height selector to some value, and counting by means of the pulse counter all of the voltage pulses whose peak value exceeded the acceptance voltage during an interval of time. Another count was made at a slightly different acceptance voltage for an equal interval of time. The difference between the values of the two counts would be the number of voltage pulses having peak values between the two acceptance voltages during the interval of time. By repeating the above operations for all voltage intervals, a complete distribution determination could be made. However, this determination was only approximate, since the energy intervals were finite in width.[1,2]

The particular case of alpha-energy determination has been discussed in detail to demonstrate the problem of pulse-height analyzing systems and to make obvious the requirements of the automatic electronic systems to be described. If electronic systems of classifying, cataloging, and counting are to be useful, they must operate on events at a high rate, and the results of these operations must reveal fine detail of magnitude distribution. If fine detail is to be retained, either the increment of classification must be very small, or the increment of classification must continuously vary its observational point; i.e., it must scan the various magnitudes of observation. Three systems of analyzers have been developed at Oak Ridge National Laboratory which meet these requirements.[3,4] Each of these analyzers was designed for a particular type of problem, and the application of each will be obvious from the following discussion.

Differential-Sweep-Type Analyzer

The differential-sweep-type analyzer (Fig. 3) is a pulse-classifying, cataloging, and counting system that operates by scanning the magnitude spectrum with a narrow "slit" or gate circuit, and counts the number of pulses that pass through the gate as it scans. Consider the operation on a series of representative voltage pulses. These pulses are impressed upon two pulse-height selectors whose acceptance voltages differ by a small fixed amount ΔV. Three conditions may exist between the peak voltage of any one pulse and the two acceptance voltages. If the peak voltage of the pulse is smaller than the two acceptance voltages, no signal will be received by the integrator due to the nonaction of the pulse-height selectors. If the peak voltage of the pulse is larger than the two acceptance voltages, no signal will be received by the integrator due to the action of the anticoincidence circuit. If, and only if, the peak voltage of the pulse is between the two acceptance voltages, a signal will be received by the integrator. The integrator is a counting-rate indicator whose output is a direct voltage which is proportional to rate of the pulses whose voltage is between the two acceptance voltages. This voltage is recorded as a position of the pen on a strip-chart recorder, and will be the ordinate in our graph of magnitude distribution.

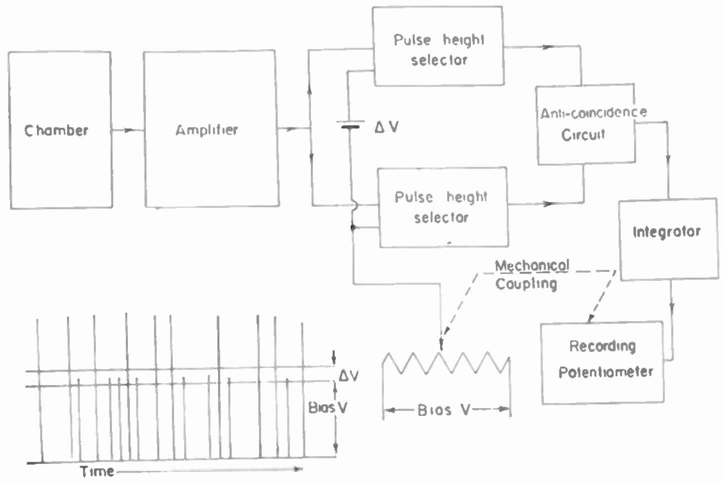

Fig. 3—Block diagram of the differential sweep-type analyzer.

The acceptance-voltage control is mechanically coupled to the strip-chart drive, and thus, as the chart moves with time, the magnitude of the events under observation is changed. The chart position is the abscissa of our desired graph. The resulting automatically plotted graph will be the number of events as a function of magnitude.

[1] A multichannel analyzer consisting of many pulse-height selectors has been used. This instrument gives only approximate distributions for the same reason.

[2] H. F. Freundlich, E. P. Hincks, and W. J. Oyeroff, "Pulse analyser for nuclear research, *Rev. Sci. Inst.* vol. 18, pp. 90–100; February, 1947.

[3] J. H. Parsons, "Differential Sweep Type Analyser," Oak Ridge National Laboratory, MonC-416, October, 1947.

[4] J. H. Parsons, "Displacement Analyser," Oak Ridge National Laboratory, MonN-432, p. 143.

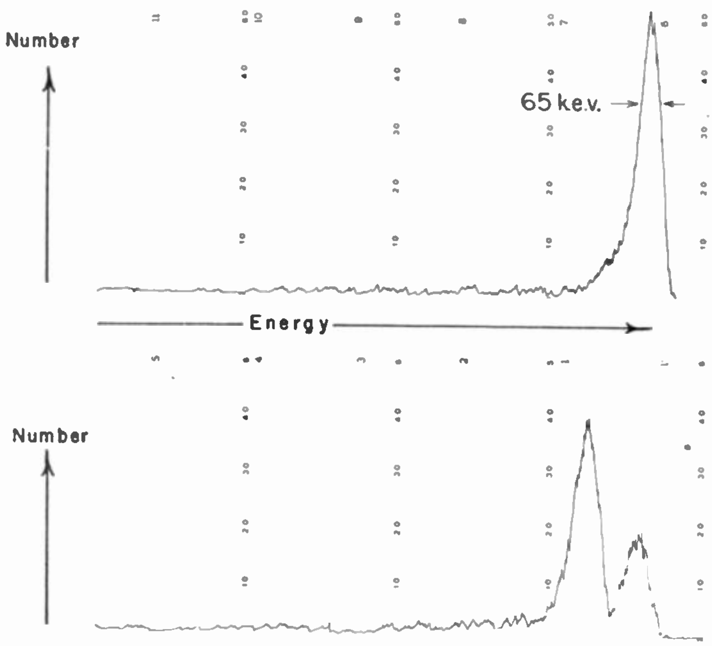

Fig. 4—Two typical charts of energy distribution made by the differential sweep analyzer.

The two charts shown (Fig. 4) were made with this system during a determination on energy distribution of alpha particles from a sample of radioactive material. The upper graph shows the energy distribution of monoenergetic alpha-emitting events. The Gaussian-type distribution caused by modulation of the pulse by amplifier noise can be seen. The lower graph was made under similar conditions, but with two monoenergetic samples under observation. If the amplifier were free of noise, these graphs would appear with very sharp peaks.

The only component circuit that may be of interest is the system of pulse-height selectors and the anticoincidence circuit (Fig. 5). Each pulse-height selector is a combination of biased diodes and a univibrator. The diodes are three IN38's in series and are negatively biased. If the impressed pulse voltage exceeds the bias voltage, the pulse will appear at the grid of the univibrator, causing it to act to supply one square wave. The pulse-height selectors are sensitive to a difference in pulse heights of about 25 millivolts, and this is the unsure region. The output square wave is the same for all pulses, if the time of action of the univibrator is long compared to the pulse duration. The biases of the two pulse-height selectors are separated by the small fixed voltage ΔV, which is supplied by a voltage cell and divider. The smallest ΔV commonly used is 250 millivolts. The anticoincidence action is accomplished by coupling the univibrators to opposite ends of a center-tapped transformer. If both univibrators act on a pulse, cancellation occurs in the transformer. The integrator will see a pulse only when one univibrator acts, indicating a pulse whose voltage is between the two acceptance voltages.

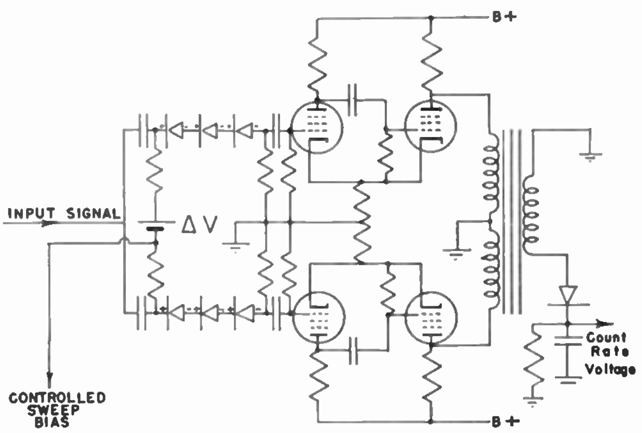

Fig. 5—The system of pulse-height selectors.

Displacement Analyzer

It is obvious that the differential-sweep analyzer does not classify, catalog, and count each and every event in a series, but only operates on a representative group. The sweep analyzer cannot be used on all magnitude-distribution studies since each and every event must be classified, cataloged, and counted if a few events are not representative of the entire series, if the events do not repeat in some fashion, or if the total number of events is small.

Two systems, the displacement analyzer and the synchronous analyzer (Fig. 6), have been developed that will act on each and every event in a series. The first function in these analyzers, classification, is the translation of the representative voltage pulse into a linear position of the beam of a cathode-ray tube. This is accomplished by electronic means. The second function, cataloging, is the recording of the series of resulting beam positions on a photographic plate. This photographic catalog of beam positions will have a displacement distribution that is an exact representation of the original magnitude distribution with all fine detail preserved. The third action, counting, is accomplished by means of a recording microdensitometer. This instrument will plot the distribution of beam positions, and hence the magnitude distribution of the series of physical events under observation. The two analyzers

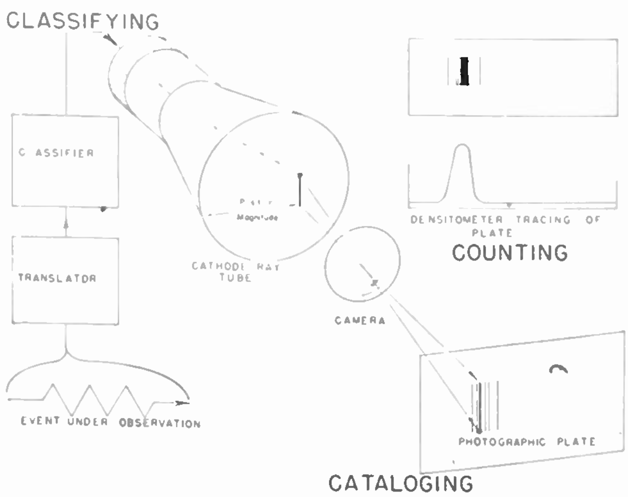

Fig. 6—The general system of the displacement and synchronous analyzers.

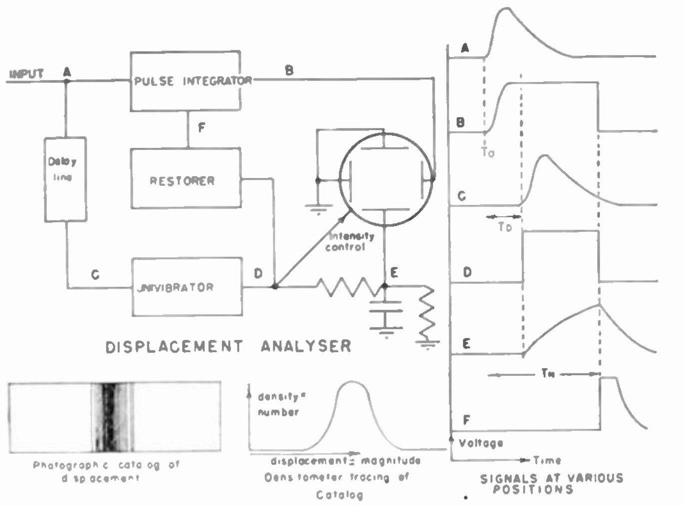

Fig. 7—Diagram of the operation of the displacement analyzer

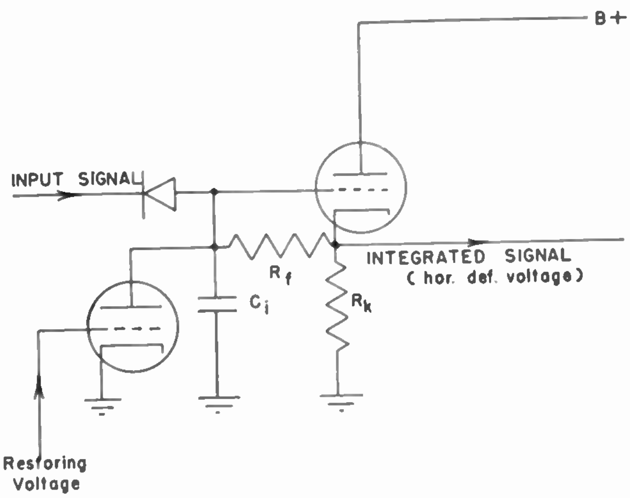

Fig. 8—Pulse integrator.

only differ in the method of translating the peak voltage of the representative pulses into linear beam positions.

The displacement analyzer (Fig. 7) directly utilizes the representative pulse to position the beam of a cathode-ray tube. Consider the operation for a single pulse in a series of representative pulses. The representative pulse is applied to the input of a pulse-integrator circuit. The output voltage of this integrator will follow the rise of the pulse until the pulse has reached its maximum voltage. This output voltage, which determines the horizontal beam position, will then remain constant until some restoring action takes place. The grid of the cathode-ray tube is negatively biased, and thus the screen will be dark until some intensifying action takes place. The original pulse is also applied to a time-delay line, and the delayed pulse triggers a univibrator. The univibrator supplies one square wave to intensify the beam of the cathode-ray tube. This square wave is also impressed through a network to the vertical deflection plates of the cathode-ray tube, causing a bright line to be drawn on the screen. When the univibrator returns to its stable position, the cathode-ray beam is turned off and restoring action takes place in the pulse integrator. At this time, the system is ready for the next voltage pulse. Thus, one bright line is drawn for each pulse, and the position of the line is determined by the pulse voltage.

All circuits except the pulse integrator (Fig. 8) are well known. The pulse integrator is a combination of a diode, a capacitor, an infinite-impedance cathode follower, and a restoring tube. The original voltage pulse is impressed on the capacitor through a diode. This capacitor is between the grid and the cathode-return point of an infinite-impedance cathode follower. The cathode voltage will follow the rising voltage of the original pulse, and then tend to remain constant after this rise. Any leakage from this capacitor will be balanced by the current through the high resistance from the cathode to the grid. The capacitor can be discharged only through the restoring tube when this tube is made conducting by an externally applied signal. This cathode voltage is the horizontal deflection voltage on the cathode-ray tube.

The action of this analyzer is limited by the rise time of the original pulse. If the pulse rise time is less than 1 microsecond, the system could perform about 10^5 operations per second on evenly time-spaced pulses.

The Synchronous Analyzer

The synchronous analyzer (Fig. 9) does not directly utilize the original pulse to determine the beam position, as in the displacement analyzer, but first translates the pulse voltage into a proportional time interval. This is advantageous, since it avoids the problem of linear driving amplifiers for the cathode-ray tube. Also, a smaller tube can be used because a circular sweep is utilized.

The general scheme of operation is shown in Fig. 9. The original pulse is applied to a pulse integrator; the peak voltage is retained as in the displacement analyzer. The output of the integrator is impressed on a voltage comparator system consisting of biased diodes and a univibrator. A delayed pulse actuates a univibrator and a sawtooth generator after the output of the integrator has reached its maximum value. The sawtooth generator is coupled to the bias system of the comparator. The

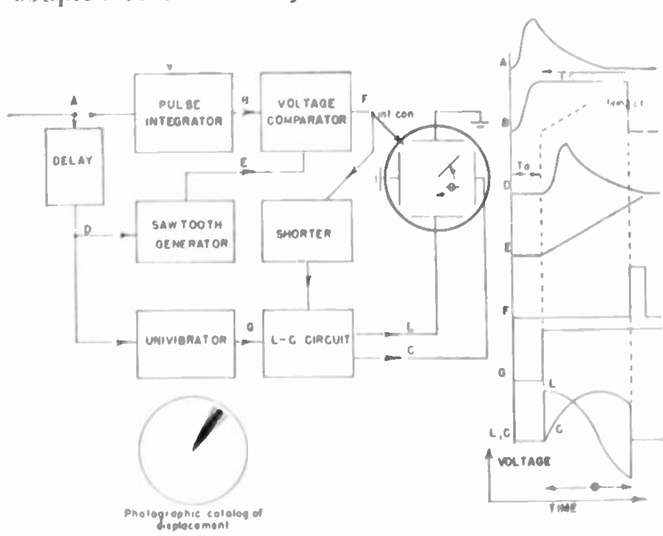

Fig. 9—Synchronous analyzer.

univibrator in the comparator will trip when the voltage of the sawtooth is equal to the voltage on the integrator. The time required for the sawtooth voltage to go from zero to the voltage on the integrator (i.e., peak pulse voltage) will be proportional to the voltage on the integrator. A measure of this time will be a measure of peak voltage of the original pulse. The univibrator started its action when the sawtooth generator started. The generated square wave is impressed on an *LC* circuit, and the resulting out-of-phase voltages cause the beam of a cathode-ray tube to sweep in a circular fashion. When the comparator is actuated, the beam is intensified, and the voltages on the components of the *LC* circuit are shorted and reduced to zero. This produces a bright radial line on the screen. The angular position of this line is the measure of the proportional time. The circuits are then restored to normal positions.

Thus, the angle of the radial line (its position) is a measure of the peak voltage of the pulse. A photograph of a series of these radial lines will be a catalog of the original series of pulses. The angular distribution of these lines will be the same as the original magnitude distribution of the events with all fine detail preserved. A variation of this circuit would involve using the sawtooth voltage to sweep the beam. This would result in a linear catalog, which would not be conservative of tube size. A spiral catalog is very conservative of tube size, and may be obtained by replacing the *LC* circuit with an *RLC* circuit.

Illustration of Application

A specific demonstration and comparison of the various methods of obtaining the magnitude distribution of a series of physical events will clarify the advantages and applications of the new analyzers. Let us use a very real problem in astronomy.

General star counts are important in determining the space distribution of stars. In star counts we are not interested in the exact location and magnitude (brightness) of a given star. We only wish to know the magnitude distribution for a particular area of the sky. The star counts can either be made directly by an astronomer peering through a low-power microscope at a photograph of the star field, or by one of the systems of classifying, cataloging, and counting.

If it is done manually, each photographic star image is examined through a low-power microscope, and a density, or magnitude, is assigned to that image. After the images have been assigned densities, the number of images (or stars) in each increment of magnitude is determined. These studies may involve hundreds of thousands of stars, and, therefore, are very laborious.

A form of analyzer developed by McCuskey and Scott in Cleveland[5] has been used on star counts in the past. In this machine the photographic plate of the star field is scanned by a photoelectric densitometer,

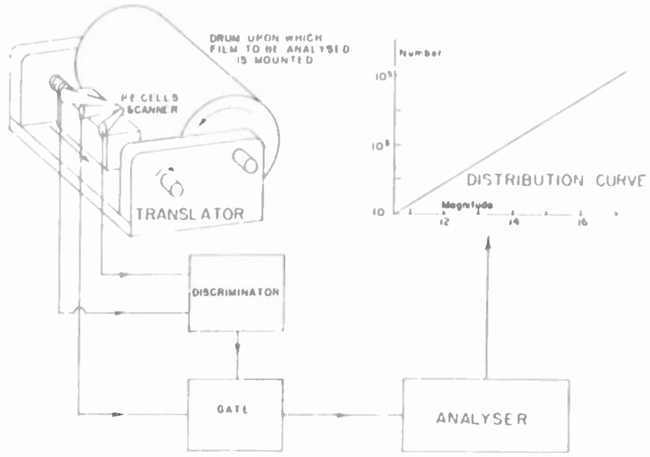

Fig. 10—Determining the distribution of magnitudes within a star field.

and the representative voltage pulses are measured by a voltage-sensitive gate circuit and associated counter. This system only determines the number of stars between set limits of magnitude, and thus furnishes no information on fine detail of distribution.

It is possible to use the displacement analyzer or the synchronous analyzer (Fig. 10) to determine the magnitude distribution of stars. Some sort of a photoelectric scanning system would be used, and the resulting series of representative pulses would be classified, cataloged, and counted by one of the new systems. The scanning system would involve two additional pickups and a discriminator to prevent spurious pulses that result when the scanner passes over part of a star image from entering into the determination. The resulting distribution determination would not only reveal the number of star images between certain limits of magnitude, but would also give the rate of change of the number versus magnitude. This determination would be much faster than the laborious manual counting, and it should be possible to reduce the time required for this type of determination from days to a matter of hours.

Conclusions

These systems of classifying, cataloging, and counting could have many applications in the fields of observational science. They may be used to determine the magnitude distribution of any series of physical events if the variable under observation can be translated into a proportional voltage pulse.

The type of analyzer used in each determination will depend on the type of magnitude distribution. The sweep analyzer is useful when a sample group is representative of the whole series of events. The displacement analyzer and the synchronous analyzer are useful when each and every event in a series must enter into the determination of distribution. The fine detail of magnitude distribution is retained in the determinations made by each of the analyzers. The analyzers could operate on events at the rate of 10^5 per second, but the speed of operation is usually limited by the rise time of the pulse from the translator.

[5] S. W. McCuskey and R. M. Scott, "A photoelectric star counter," *Rev. Sci. Instr.*, pp. 597–601; December, 1941.

Graphical Analysis of Linear Magnetic Recording Using High-Frequency Excitation*

MARVIN CAMRAS†, SENIOR MEMBER, IRE

Summary—**The addition of a high-frequency component to an audio signal which is to be recorded magnetically results in a low-distortion, linear recording characteristic under certain conditions. This paper gives a graphical method for constructing the recording characteristic from the B_R versus H curve of the record material. An analysis accounts for such magnetic-recording characteristics as variation in sensitivity with bias, linearity at low recording levels, adjustment for maximum sensitivity, and adjustment for minimum distortion.**

ONE OF THE most important factors contributing to the high quality of modern magnetic wire and tape recorders is the use of high-frequency excitation superimposed on the audio signal to be recorded. The process has been designated by such names as "supersonic bias," "high-frequency bias," "high-frequency shaking," "anhysteretic excitation," "ac bias," "high-frequency excitation," and the like, and a number of explanations of the nature of this process have been offered.[1-3] The wide diversity among theories indicates that the recording process is complex; attempts at simplification generally assume conditions which are different from actual operating conditions. Hence, the theoretical explanations must be modified considerably to agree with experimental results.

This paper offers an analysis consistent with observed results under the conditions specified. Results under usual operating conditions also seem to follow the theory. A graphical method is given for calculating the recording characteristic curve, which explains much of the experimental data on such things as variation in sensitivity with bias, linearity at low recording levels, adjustment for maximum sensitivity, minimum distortion, etc.

Anyone who has tried to record magnetically on a demagnetized wire or tape has found that this method gives extremely poor quality and low sensitivity. Inspection of the B_R versus H characteristic of a typical magnetic recording material (Fig. 1) shows the reason for this deficiency. A recording field below 50 oersteds leaves practically no magnetization in the wire. Only the peaks of signals above 50 oersteds will be recorded, producing a distorted wave form of retained flux (Fig. 2).

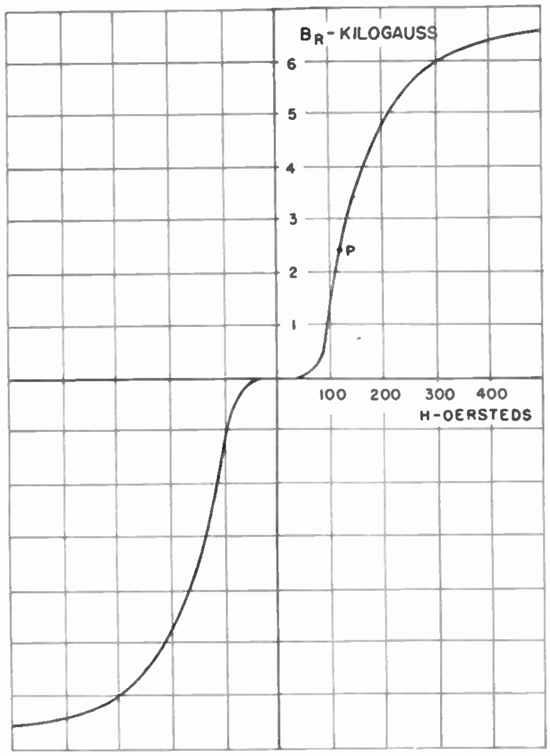

Fig. 1—Retained flux density versus field for a typical magnetic recording medium.

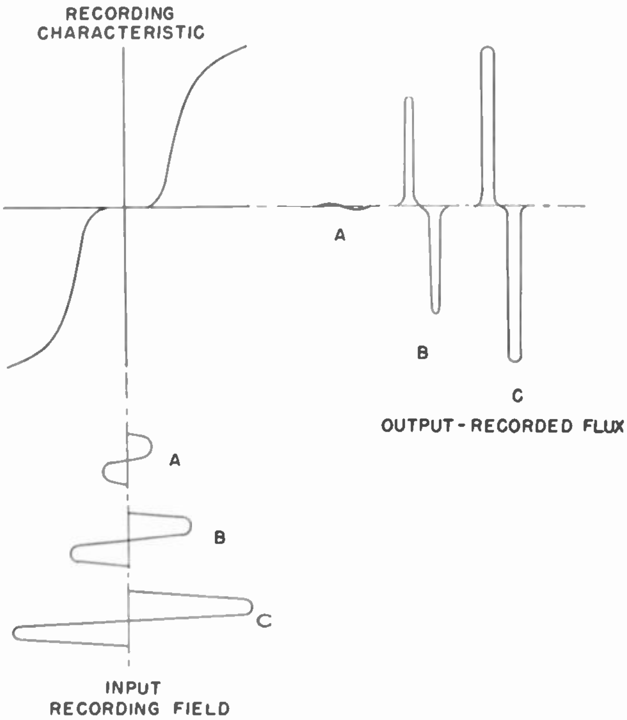

Fig. 2—Distortion caused by attempt to record directly on a demagnetized record medium.

* Decimal classification: R365.35. Original manuscript received by the Institute, May 17, 1948; revised manuscript received, November 3, 1948.

† Armour Research Foundation, Illinois Institute of Technology, Chicago, Ill.

[1] H. Toomin and D. Wildfeuer, "Mechanism of supersonic frequencies as applied to magnetic recording," PROC. I.R.E., vol. 32, pp. 664–668; November, 1944.

[2] L. C. Holmes and D. L. Clark, "Supersonic bias for magnetic recording," *Electronics*, vol. 18, p. 126; July, 1945.

[3] M. Camras, U. S. Patent number 2,351,004.

To give a more linear recording characteristic, a dc field of about 125 oersteds could be used together with a small audio signal, thus operating on the straighter portion of the curve in the vicinity of point P in Fig. 1. The classical method of achieving linearity was first used by Poulsen. Here the wire is prepared by saturating in a dc field. In the recording head it encounters an opposite dc field plus the audio to be recorded. Neither of these methods is ideal from a standpoint of distortion, operating range, or noise.

It was discovered that adding a steady high-frequency flux to the audio signal when recording on a demagnetized material increased the sensitivity. Moreover, if the high-frequency excitation was adjusted to the proper magnitude, a straight-line recording characteristic would result. The modified values of residual flux, B_{RM}, are plotted against H in Fig. 3. Some other interesting facts were noted:

1. The new recording characteristic is especially linear in the vicinity of the origin; i.e., for small values of H.
2. The recording characteristic is linear, even through corresponding sections of the B_R versus H plot are curved.
3. The range over which linearity could be obtained is greater than any fairly straight portion of the B_R versus H curve.
4. The frequency of the high-frequency component is not especially critical. In fact, frequencies in the order of 1 kc or less will straighten the recording characteristic (although the steady tone together with the output signal is objectionable).

The fact that the high frequency need not be beyond the recording range of the system leads to a rather interesting explanation of how the system operates under such conditions. Referring to Fig. 4, let a high-frequency signal of maximum value E, whose average value is displaced from the zero axis by an amount A, be applied to the B_R versus H curve. The total range covered will be between points 1 and 2 on the B_R versus H curve, and the recorded flux will also lie between these limits. Note that the average of the recorded flux (4) does not correspond to the average of the input signal (3).[4] Fig. 4 gives a convenient graphical solution for the average recorded flux corresponding to a displacement A of a high-frequency signal of amplitude E. Measure a distance E horizontally to the left of point 2 to get point 5. Measure a distance E horizontally to the right of point 1 to get point 6. The midpoint between 5 and 6 gives 7, which is the average recorded flux corresponding to A.

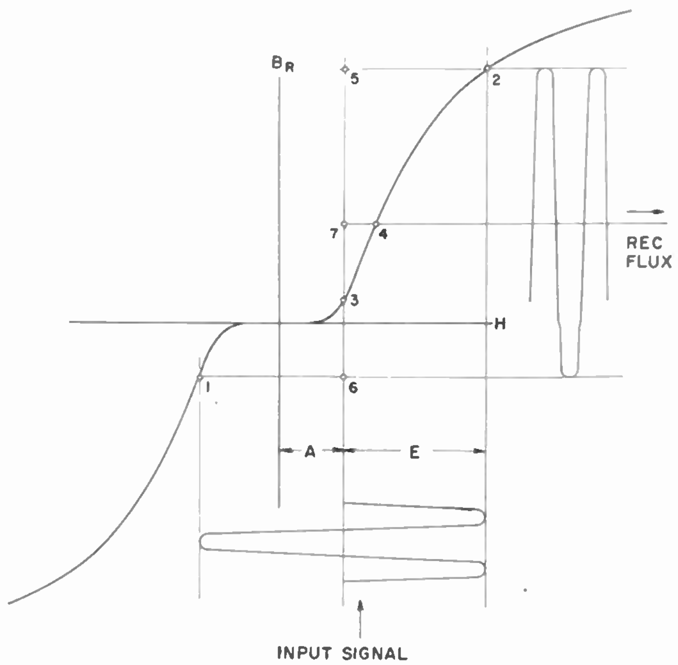

Fig. 4—Average retained flux for a displaced high-frequency signal.

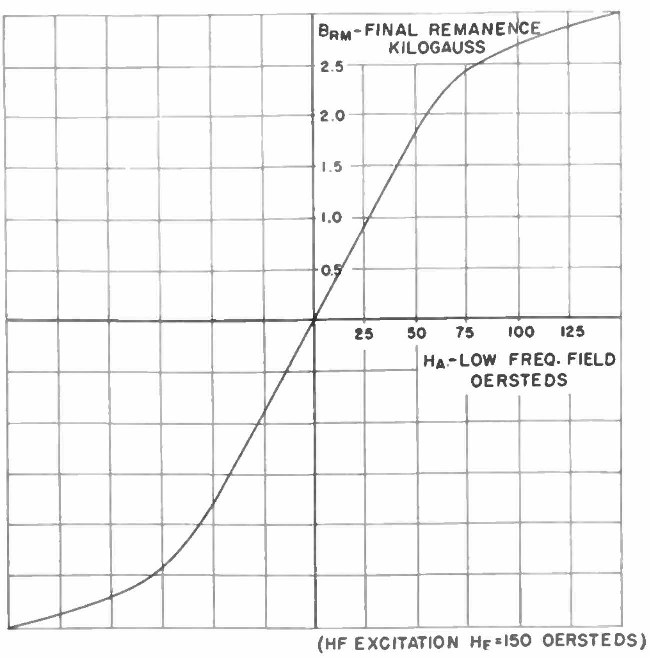

Fig. 3—Linear recording characteristic obtained by the addition of high-frequency excitation.

Suppose that the high-frequency excitation is held at a constant amplitude $H=E$, and the displacement A is varied according to the desired audio signal to be recorded. The locus of all points such as 5 (Fig. 4) is the B_R versus H characteristic moved horizontally to the left a distance E. Call this curve B_{R1} (Fig. 5). Similarly, curve B_{R2} is the locus of points such as 6. Finally, by taking the midpoints between B_{R1} and B_{R2} the locus of points as 7 is obtained. This curve, B_{RM}, gives the recorded audio flux corresponding to the input audio field

[4] The arithmetic mean of the range of recorded flux is assumed to be the average. Because of distortion, this is strictly true only when A is zero. For small values of A, the assumption gives a close working approximation. In playing back, the high-frequency component must, of course, be averaged out by an element that responds to average, rather than rms, values.

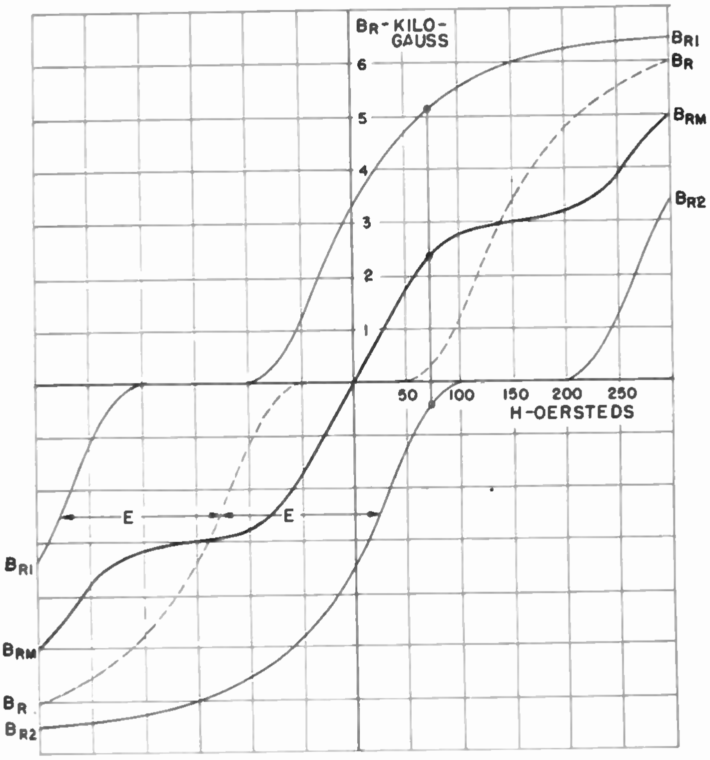

Fig. 5—Graphical construction of recording characteristic B_{RM}.

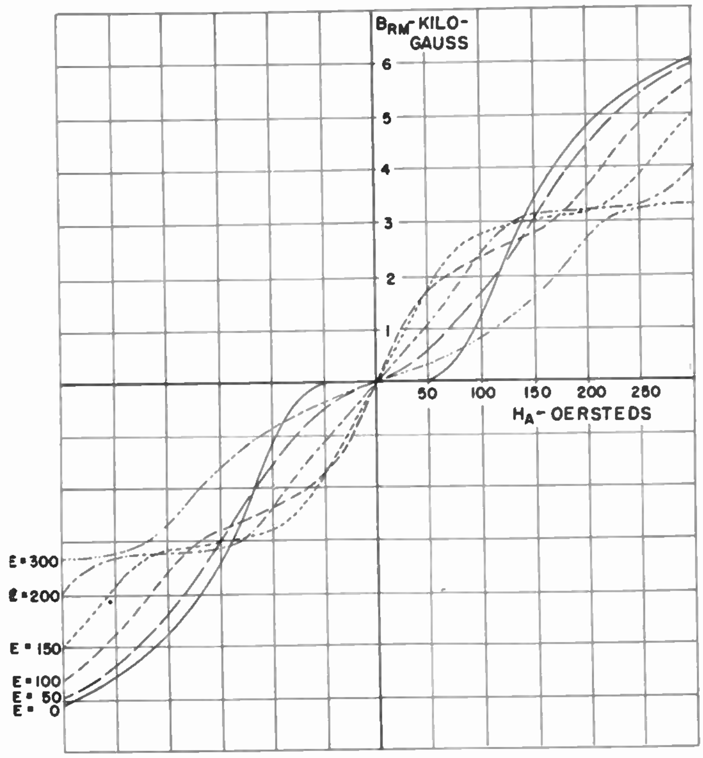

Fig. 6—Recording characteristics of a typical record medium for different values of high-frequency excitation.

(the high-frequency component being averaged out in each case). Using functional notation, B_R is a function of H, B_{R1} is the same function of $H+E$, B_{R2} is the same function of $H-E$, B_{RM} is the mean of B_{R1} and B_{R2}.

$$B_R = F(H) \tag{1}$$

$$B_{R1} = F(H + E) \tag{2}$$

$$B_{R2} = F(H - E) \tag{3}$$

$$B_{RM} = 1/2[F(H + E) + F(H - E)]. \qquad (4)^5$$

Fig. 6 is a plot of the derived B_{RM} versus H curves for a typical sample of recording wire, and Fig. 7 gives the calculated input versus output curves. The form of these curves checks very well with experimental data.

From the construction in Fig. 5 and the general form of a B_R versus H curve, a number of interesting deductions can be made:

1. *If, for purposes of comparing output versus input, we define the recording sensitivity by the equation $S=B_{RM}/H$, then the recording sensitivity for small audio signals,*

$$S = \left.\frac{dB_{RM}}{dH}\right|_{H=0}, \tag{5}$$

is equal to the slope of the B_R versus H curve reached by the peak[6] of the high-frequency excitation. (Slope of B_R versus H curve at $H=E$.)

[5] In the general case, $B_{RM}=G(H)[F(H+E)+F(H-E)]$ where $G(H)$ is an arbitrary function.

[6] The effective peak may be higher than the theoretical peak because of magnetic lag, etc.

To show this, refer to Fig. 5. For any value of H,

$$B_{RM} = 1/2(B_{R1} + B_{R2}). \tag{6}$$

The sensitivity at the origin is equal to the derivative of B_{RM} with respect to H at $H=0$.

$$\frac{dB_{RM}}{dH} = 1/2\left[\frac{dB_{R1}}{dH} + \frac{dB_{R2}}{dH}\right]. \tag{7}$$

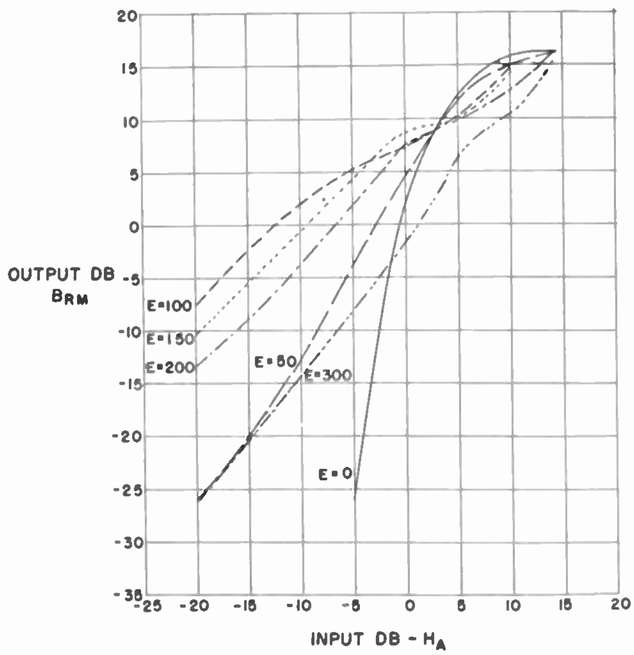

Fig. 7—Input versus output curves corresponding to the recording characteristics of Fig. 6.

From symmetry, it is noted that $dB_{R1}/dH = dB_{R2}/dH$ at $H=0$. Also,

$$\left.\frac{dB_{R1}}{dH}\right|_0 = \left.\frac{dB_R}{dH}\right|_E,$$

since curve B_{R1} is merely curve B_R displaced E units to the left. Therefore,

$$S = \left.\frac{dB_{RM}}{dH}\right|_0 = 1/2\left[\left.\frac{dB_R}{dH}\right|_E + \left.\frac{dB_R}{dH}\right|_E\right] = \left.\frac{dB_R}{dH}\right|_E, \qquad (8)$$

and thus the sensitivity for small signals is the slope of the B_R versus H curve at $H=E$.

2. *For small audio signals, when high-frequency excitation is added the sensitivity increases rapidly, goes through a peak, then decreases slowly, approaching zero as a limit.*

This is apparent from the first derivative of the B_R versus H curve (Fig. 8).

3. *For small audio signals, the sensitivity will be a maximum when the high-frequency excitation is adjusted to reach the point of inflection on the B_R versus H curve.*

The condition for maximum sensitivity is that dB_R/dH be a maximum. This is true when its derivative is zero:

$$\frac{d^2B_R}{dH^2} = 0, \qquad (9)$$

which corresponds to the point of inflection on the B_R versus H curve (see Fig. 8).

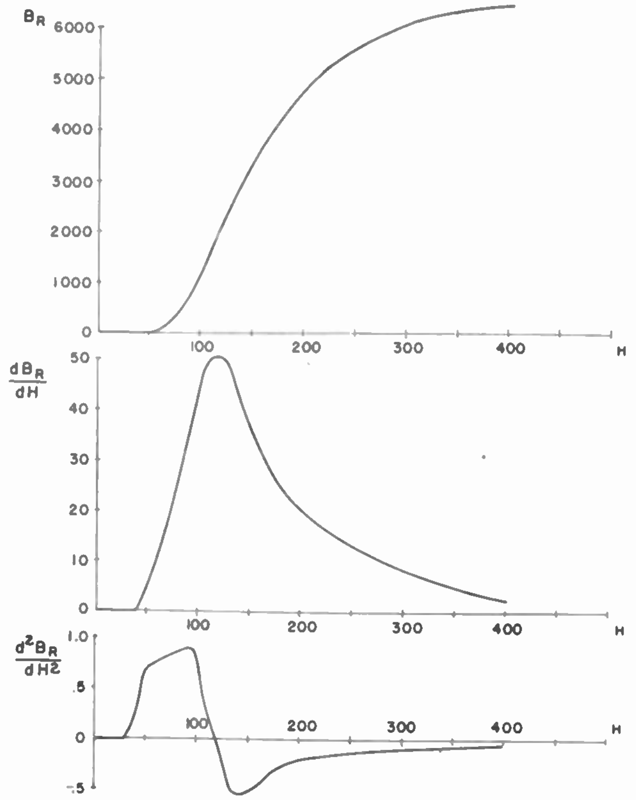

Fig. 8—First and second derivatives of B_R give sensitivity and linearity characteristics.

4. *For large audio signals approaching saturation, maximum output occurs with no high-frequency component. Increasing values of high frequency decrease the B_{RM}.*

Fig. 6 shows that, for $H_A=300$, the B_{RM} falls off as the high-frequency excitation increases. Positive peaks of high frequency do not add to the retained magnetism because of saturation; negative peaks decrease the retained magnetism. The net effect is a lower B_{RM}.

5. *Optimum value of amplitude of high frequency for maximum output will be different for each audio amplitude.*

According to Fig. 6, the high-frequency amplitudes giving greatest output for an audio signal of 25 oersteds are 100, 150, 200, 50, 300, 0, arranged in order. For a signal of 75 oersteds, the order is 150, 100, 200, 50, 300, 0; for a signal of 100 oersteds, 150, 200, 100, 50, 0, 300; at saturation, 0, 50, 100, 150, 200, 300. A crossing over of the recording characteristics is inevitable, since they are always in a certain order for initial sensitivity, and in a different order for saturation sensitivity.

6. *The recording characteristic is perfectly linear if E is chosen so that, over the operating range,*

$$\frac{d^2B_{R1}}{dH^2} + \frac{d^2B_{R2}}{dH^2} = 0.$$

The condition for linearity of the B_{RM} versus H characteristic is that its slope does not change:

$$\frac{dB_{RM}}{dH} = 1/2\left[\frac{dB_{R2}}{dH} + \frac{dB_{R2}}{dH}\right] = \text{constant}. \qquad (10)$$

Again taking the derivative with respect to H,

$$\frac{d^2B_{RM}}{dH^2} = 1/2\left[\frac{d^2B_{R1}}{dH^2} + \frac{d^2B_{R2}}{dH^2}\right] = 0, \qquad (11)$$

or

$$\frac{d^2B_{R1}}{dH^2} + \frac{d^2B_{R2}}{dH^2} = 0. \qquad (12)$$

The physical explanation is that, when (12) is satisfied, the curvature of B_{R1} exactly compensates for that of B_{R2} to give a linear B_{RM}.

7. *The recording characteristic B_{RM} versus H for small values of H_A is more linear than the B_R versus H characteristic when the operating point is at a curved portion of the B_R versus H characteristic.*

Examination of the B_{R1} and B_{R2} curves of Fig. 5 show that, because of symmetry about the origin, when $H=0$,

$$B_{R2} = -B_{R1}, \qquad (13)$$

$$\frac{dB_{R2}}{dH} = \frac{dB_{R1}}{dH}, \quad \text{and} \qquad (14)$$

$$\frac{d^2B_{R2}}{dH^2} = -\frac{d^2B_{R1}}{dH^2}. \qquad (15)$$

Equation (15) satisfies the condition of (12) for linearity. If variations in H are small, $\Delta^2 B_{R2}/\Delta H^2$ is of opposite sign to $\Delta^2 B_{R1}/\Delta H^2$, and nearly equal to it, so that

$$\frac{\Delta^2 B_{RM}}{\Delta H^2} = 1/2\left[\frac{\Delta^2 B_{R1}}{\Delta H^2} + \frac{\Delta^2 B_{R2}}{\Delta H^2}\right] \tag{16}$$

is closer to zero than either one of its terms.

8. *As the audio amplitude is decreased, the recording characteristic approaches linearity.*

This is the limiting case as ΔH in (16) approaches zero.

9. *The recording characteristic is perfectly linear if the operating range is over a section of the B_R versus H curve which can be represented by a second-degree (or lower) equation.*

Assume

$$B_R = AH^2 + BH + C. \tag{17}$$

Then

$$\frac{dB_R}{dH} = 2AH + B, \tag{18}$$

and

$$\frac{d^2 B_R}{dH^2} = 2A. \tag{19}$$

If the positive section of the B_R curve is displaced to give the B_{R1} characteristic, its second derivative will have a constant value $2A$, at all points within the second-degree range. The corresponding operating portion of the B_{R2} curve will have a second derivative, $-2A$, constant at all points. These conditions satisfy (12) for all points, giving a linear recording curve.

It is interesting to draw on tracing paper two sets of d^2B_R/dH^2 curves as given in Fig. 8, using both positive and negative values for H. These are slid over one another to test for the amount of cancellation obtained when various amounts of high-frequency excitations are used. A positive remainder indicates that the recording characteristic is concave upward; a negative remainder indicates a downward concave portion.

Although the analysis of this paper is primarily graphical, similar results may be obtained by mathematical analysis of the kind used for frequencies combined in nonlinear impedances. The B_R versus H characteristic is described by an equation of the form $B_R = AH + BH^3 + CH^5 + \cdots$. This was first suggested by R. E. Zenner, and later elaborated by W. W. Hansen, J. J. Fischer, and J. Markin (all of Armour Research Foundation) in unpublished work.

Q Measurements—Two- and Four-Terminal Networks*

M. C. PEASE†, MEMBER, IRE

Summary—Considering the equivalent circuit of a simple shunt-resonant four-terminal network, including loss, equations are derived permitting the calculation of unloaded and doubly loaded Q's, and the resonant frequency, from standing-wave-ratio or transmission-coefficient measurement at three arbitrary frequencies or wavelengths. The equations are then put in a convenient form for data taken on a "triple-pipper impedance bridge." Finally, the equations are given for the same type of measurement for a two-terminal network, giving the unloaded and loaded Q's.

IN THE NEIGHBORHOOD of its resonance, and so far as its behavior at the input or output terminals is concerned, many uhf shunt-resonant transmission elements, e.g., waveguide irises, can be completely described by the resonant frequency and the unloaded and doubly loaded Q's, together with an ideal transformer, if the structure is unsymmetrical.

A two-terminal network, such as a simple cavity, often may be described by its shunt-resonant frequency and its unloaded and loaded Q.

It is the purpose of this paper to develop convenient methods of measuring these parameters. We have simplified the four-terminal problem slightly by assuming symmetry, which eliminates the ideal transformer. Otherwise, the methods are limited only by the assumption of the validity of a lumped-constant equivalent circuit.

The usual methods of evaluating these parameters require some method of obtaining the half-power points. In many applications, however—e.g., resonant irises in waveguides—we may have a doubly loaded Q as low as 2 or 3.

Under such circumstances, the half-power points are very far from the center frequency. Not only is it difficult to find oscillators which can be tuned over a sufficiently wide range to obtain data at the half-power points, but a constant-valued equivalent circuit will not be valid for the range of frequencies required.

On the basis of the simplifying assumption of no loss (infinite unloaded Q), other methods have been evolved. We may calculate

$$Q_{2L} = \lim_{\lambda \to \lambda_0} \frac{\sigma - 1}{4\sqrt{\sigma}} \left|\frac{\lambda_0}{\lambda - \lambda_0}\right| \tag{1}$$[1]

* Decimal classification: R143×R244. Original manuscript received by the Institute, May 11, 1948; revised manuscript received, July 16, 1948.

† Sylvania Electric Products Inc., Boston, Mass.

[1] "Microwave Duplexers," edited by L. D. Smullin and C. G. Montgomery, vol. 14, "Radiation Laboratory Series," McGraw-Hill Book Co., New York, N. Y.; 1948.

or

$$= \lim_{\lambda \to \lambda_0} \frac{\lambda_0}{2} \frac{d|\Gamma|}{d\lambda} \tag{2}$$ [1]

from data in the immediate neighborhood of resonance.

The formulas of this paper are free from these limitations.

There are many experimental methods that can be used to obtain the necessary data. We shall, however, take as basic the measurement of voltage-standing-wave ratio as a function of frequency. If desired, the formulas can be readily converted to the use of transmission coefficients, or the independent variable can be changed to wavelengths.

We shall obtain, first, general formulas for measurements at any three points.

Second, the use of a "triple-pipper impedance bridge" permits presetting the three frequencies to any desired points. By maintaining a suitable relationship between them, the original formulas can be considerably simplified, particularly if an approximation is valid.

Finally, since the methods obtained may be useful for high-Q two-terminal cavities where speed of measurement and calculation is desired, we shall obtain the solutions for this case. It will be found that two solutions are obtained. We cannot determine, theoretically, which is correct without phase data.

I. Four-Terminal Networks

A. General Formulas

We shall approximate our four-terminal singly shunt-resonant network by the circuit of Fig. 1, where the load,

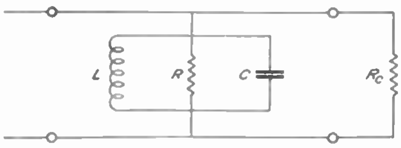

Fig. 1—Equivalent circuit of a simply resonant four-terminal network.

R_c, is the characteristic impedance of the line.

We define the doubly loaded Q as

$$Q_{2L} = \frac{1}{2} \frac{RR_c}{R + R_c} \omega_0 C. \tag{3}$$

This is equivalent to defining Q_{2L} as $\omega_0/(\omega_1 - \omega_2)$ where ω_1 and ω_2 are the frequencies at which half the incident power is reflected. Certain other authors[1] define Q_{2L} as $\omega_0/(\omega_1' - \omega_2')$ where ω_1' and ω_2' are the frequencies at which the susceptance equals plus or minus the total conductance. The two definitions are nearly equivalent, however.

Let

$$x = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} = \frac{f}{f_0} - \frac{f_0}{f} = \frac{\lambda_0}{\lambda} - \frac{\lambda}{\lambda_0}, \tag{4}$$

and

$$\sigma_0 = \frac{R + R_c}{R}. \tag{5}$$

We find that the normalized input admittance is

$$y = \sigma_0(1 + j2Q_{2L}x). \tag{6}$$

The complex reflection coefficient Γ is defined as

$$\Gamma = (y - 1)/(y + 1), \tag{7}$$

and the SWR(V), σ, as

$$\sigma = \{1 + |\Gamma|\}/\{1 - |\Gamma|\}. \tag{8}$$

Substituting (6) and (7) in (8), rationalizing the denominator, isolating the remaining radical, and squaring, we obtain the basic equation:

$$4Q_{2L}^2 x^2 \sigma_0 = \left(\sigma + \frac{1}{\sigma}\right) - \left(\sigma_0 + \frac{1}{\sigma_0}\right). \tag{9}$$

Let σ_1, σ_2, and σ_3 be the values at frequencies f_1, f_2, and f_3, respectively. Then let

$$k_i = \sigma_i + \frac{1}{\sigma_i} \quad (i = 1, 2, 3). \tag{10}$$

If

$$k_0 = \sigma_0 + \frac{1}{\sigma_0} \tag{11}$$

and

$$a = f_0^2/4Q_{2L}^2\sigma_0, \tag{12}$$

then we have

$$\left(\frac{f_i^2 - f_0^2}{f_i}\right)^2 = a(k_i - k_0) \quad (i = 1, 2, 3) \tag{13}$$

to be solved for f_0, k_0, and a.

Eliminating k_0 between two pairs of these, and solving each of the resulting two equations for f_0^4, a may be found. Then f_0^4 may be obtained explicitly.

The resultant equations, however, can be simplified by writing

$$k_1' = k_1 - k_2 \tag{14}$$

and

$$k_3' = k_3 - k_2.$$

Then

$$a = -\frac{(f_1^2 - f_2^2)(f_2^2 - f_3^2)(f_3^2 - f_1^2)}{f_1^2(f_2^2 - f_3^2)k_1' + f_3^2(f_1^2 - f_2^2)k_3'} \tag{15}$$

and

$$f_0^4 = f_1^2 f_2^2 f_3^2 \frac{(f_2^2 - f_3^2)k_1' + (f_1^2 - f_2^2)k_3'}{f_1^2(f_2^2 - f_3^2)k_1' + f_3^2(f_1^2 - f_2^2)k_3'}. \qquad (16)$$

From (13),

$$k_0 = k_2 - \frac{1}{a}\left(\frac{f_2^2 - f_0^2}{f_2}\right)^2. \qquad (17)$$

From (11),

$$\sigma_0 = \tfrac{1}{2}(k_0 \pm \sqrt{k_0^2 - 4}). \qquad (18)$$

It may be easily shown that, if σ_0 is greater than 1, the positive value must be taken.

Finally, from (12),

$$Q_{2L} = \frac{f_0}{2\sqrt{a\sigma_0}} \qquad (19)$$

and, from (3) and (5), the unloaded Q

$$Q_u = R\omega_0 C = 2\frac{\sigma_0}{\sigma_0 - 1}Q_{2L}. \qquad (20)$$

These are the equations desired.

These equations may be easily written for wavelengths if we use

$$a' = a/c^2$$

where

$$c = \lambda f.$$

Similar equations can also be obtained using the transmission factor, $K = $ (power in)/(power out), in place of σ. The set of equations to be solved[2] is

$$\left(\frac{f_i^2 - f_0^2}{f_i}\right)^2 = a^*(K_i - K_0) \qquad (21)$$

where

$$a^* = f_0^2/(2\sqrt{K_0} - 1)^2 Q_{2L}^2. \qquad (22)$$

Equation (21) is formally identical with (13).

B. SWR(V) Data Obtained at Three Preset Frequencies

Where the frequencies may be preset in any given relationship to each other, though in unknown relation to the center frequency—as, for instance, if use is to be made of a "triple-pipper impedance bridge"—a considerable simplification is possible. We choose the frequencies so that

$$\begin{aligned} f_1^2 &= f_2^2(1 - \Delta) \\ f_3^2 &= f_2^2(1 + \Delta) \end{aligned} \qquad (23)$$

where Δ is a positive constant <1. For Δ less than, say, 5 per cent, this is approximately equivalent to frequencies spaced by $f_2(\Delta/2)$.

Let

$$m = (k_1' - k_3')/(k_1' + k_3'). \qquad (24)$$

Substituting (23) in (15) and (16),

$$\frac{a}{f_2^2} = \frac{2}{k_1' + k_3'} \cdot \frac{\Delta^2}{1 - m\Delta} \qquad (25)$$

$$\doteq \frac{2}{(k_1' + k_3')}\Delta^2(1 + m\Delta) \quad \text{if} \quad m\Delta \ll 1. \qquad (25a)$$

Let

$$p = \left(\frac{f_0}{f_2}\right)^4 - 1 = \frac{m\Delta - \Delta^2}{1 - m\Delta} \qquad (26)$$

$$\doteq \Delta(m - \Delta)(1 + m\Delta) \quad \text{if} \quad m\Delta \ll 1. \qquad (26a)$$

Since p is usually a very small quantity if f_2 is taken near resonance,

$$f_0 \doteq f_2 + \tfrac{1}{4}pf_2 \qquad (27a)$$

and

$$k_0 \doteq k_2 - \frac{p^2}{4(a/f_2^2)}. \qquad (28a)$$

As before, (18) gives σ_0, and (19) gives Q_{2L}.

C. SWR(V) For Three Fixed Wavelengths

Similar equations can be obtained in terms of wavelengths, if we set

$$\begin{aligned} \lambda_1^2 &= \lambda_2^2(1 - \Delta) \\ \lambda_3^2 &= \lambda_2^2(1 + \Delta). \end{aligned} \qquad (29)$$

Then

$$a'\lambda_2^2 = \frac{2}{k_1' + k_3'} \frac{\Delta^2}{1 - \Delta^2} \doteq \frac{2\Delta^2(1 + \Delta^2)}{k_1' + k_3'} \qquad (30)$$

and

$$p' = \left(\frac{\lambda_0}{\lambda_2}\right)^4 - 1 \qquad (31)$$

$$= \frac{m\Delta - \Delta^2}{1 - m\Delta}$$

$$\doteq \Delta(m - \Delta)(1 + m\Delta)$$

and

$$\lambda_0 \doteq \lambda_2 + p\lambda_2/4 \qquad (32)$$

$$k_0 \doteq k_2 - \frac{p^2}{4a'\lambda_2^2}. \qquad (33)$$

Equation (18) gives σ_0, and Q_{2L} is given by the wavelength analogue of (19):

[2] Obtained, for example, by the methods of the paper by Paul I. Richards, "Applications of matrix algebra to filter theory," Proc. I.R.E., vol. 34, pp. 145P–151P; March, 1946.

$$Q_{2L} = \frac{1}{2\lambda_0\sqrt{a'\sigma_0}} \cdot \tag{34}$$

If transmission factors are given, the calculations are identical with I-B or I-C, except that (K_1-K_2) and (K_3-K_2) are used in place of k_1' and k_3'; and Q_{2L} is calculated from

$$Q_{2L} = \frac{(f_0/f_2)}{(2\sqrt{K_0}-1)\sqrt{(a^*/f_2^2)}} \tag{35}$$

or as

$$Q_{2L} = 1/(2\sqrt{K_0}-1)(\lambda_0/\lambda_2)\sqrt{a^*\lambda_2^2}. \tag{36}$$

D. Example

As an example, we show in Fig. 2 the experimental points obtained for a typical low-Q iris.

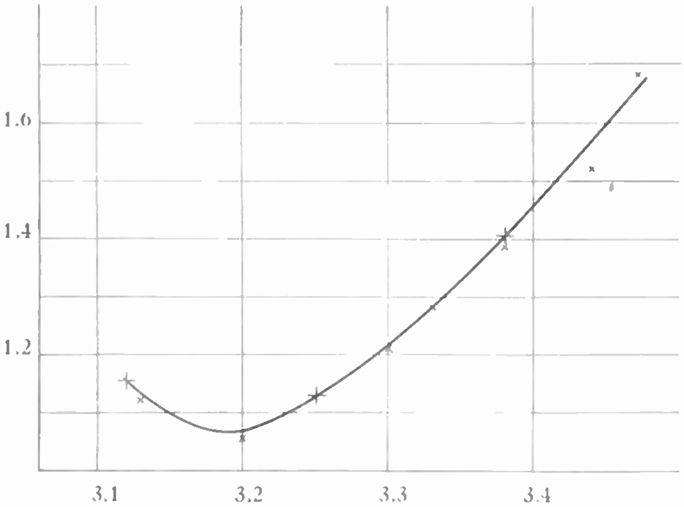

Fig. 2—SWR(V) versus λ of a typical window.

Reading the three points marked off in Fig. 2, at wavelengths calculated from $\Delta=0.08$ and $\lambda_2=3.25$ cm,

	λ	σ	k_i	
(1)	3.12	1.155	2.0209	$k_1'=0.0059$
(2)	3.25	1.130	2.0150	
(3)	3.38	1.405	2.1160	$k_3'=0.1010$

$$\Delta = .08 \qquad m = \frac{-0.0951}{0.1069} = -0.891.$$

Then

$$a\lambda_2^2 = 0.120$$
$$p = -0.0721$$
$$k_0 = 2.0042$$
$$\lambda_0 = 3.191 \text{ cm.}$$

Thus

$$\sigma_0 = 1.067 \quad \text{and} \quad Q_{2L} = 1.42.$$

It may be noted that, while the theoretical applicability of the formulas is not affected by taking the measurements far off resonance, the sensitivity of the results to experimental error requires, in practice, that the observations be centered in the immediate neighborhood of the center band.

II. Two-Terminal Networks

The equations for a two-terminal network are nearly identical to the four-terminal networks considered above. However, some modification is required.

We must distinguish two cases. If the input coupling is weak, there is zero net phase change as the frequency goes from zero to infinity, as with the four-terminal networks considered above.

If the coupling is strong, however, the phase of the complex reflection coefficient changes by 2π radians as the frequency goes from zero to infinity. At resonance its phase is opposite to that at either limit.

For this reason, we shall have to consider carefully how we define the center-band SWR.

Our equivalent circuit consists of L, C, and R in shunt, terminating a line of characteristic impedance R_c. We define

$$Q_u = R\sqrt{\frac{C}{L}} \cdot \tag{37}$$

Since, for such applications as magnetrons, etc., the coupling will normally be large, R, in general, will be greater than R_c. The SWR(V) measured, then, will be R/R_c at resonance, the reciprocal of σ_0. We shall denote this by σ_0', and

$$\sigma_0' = R/R_c. \tag{38}$$

Then the input admittance

$$y = (1 + jQ_u x)/\sigma_0'. \tag{39}$$

We observe that this is formally identical to (6) with $1/\sigma_0'$ replacing σ_0, and Q_u replacing $2Q_{2L}$.

In the calculation of f_0, k_0, and a, the SWR enters in only as $k=\sigma+1/\sigma$. These formulas, therefore, are invariant under the change from σ_0 to σ_0'.

The solution of k_0 for σ_0' requires careful consideration. It may be seen that in (18) the positive sign gives either σ_0 or σ_0', whichever is greater than 1, the negative sign giving the other.

With this ambiguity, then, the formulas for a, k_0, f_0, and σ_0 or σ_0' are the same as before. We then have

$$Q_u = f_0\sqrt{\sigma_0'/a} = f_0/\sqrt{\sigma_0 a}. \tag{40}$$

The loaded Q is

$$Q_L = \frac{RR_c}{R+R_c}\omega_0 C = \frac{Q_u}{1+\sigma_0'} = \frac{\sigma_0}{1+\sigma_0}Q_u. \tag{41}$$

The external Q is

$$Q_E = R_c\omega_0 C = Q_u/\sigma_0' = \sigma_0 Q_u. \tag{42}$$

The circuit efficiency η_c is given by

$$\eta_c = \frac{1/R_c}{1/R_c + 1/R} = \frac{\sigma_0'}{1 + \sigma_0'} = \frac{1}{1 + \sigma_0}. \quad (43)$$

Similar equations in terms of wavelength can be written immediately, if desired, from Section I-A. Also, if the frequencies or wavelengths are chosen in advance in accordance with (23) or (29), formulas like those in Section I-B or I-C can be found immediately.

These formulas lead to a fundamental ambiguity in the measurements. When σ_0 or σ_0' is obtained from (18), the question of which has been obtained cannot be answered without the introduction of some different types of data. The simplest of such data is phase. If the position of a minimum at resonance is identical with that at the asymptote, $\eta_c < 50$ per cent, and σ_0 is >1. If the position of the minimum is shifted $\lambda/4$ from the asymptotic position, $\eta_c > 50$ per cent, and σ_0', instead, is >1.

III. Conclusions

We have derived formulas for calculating the resonant frequency and Q's of simply shunt-resonant two- and four-terminal uhf components, from the measurement of voltage-standing-wave ratios or power-transmission factors at three frequencies or wavelengths.

If the frequencies or wavelengths can be preset in a definite mutual relationship, the formulas can be given in a form that permits much easier computations.

These formulas are general and exact within the range over which the equivalent circuits, including loss, are valid.

In the two-terminal case, two solutions are obtained, leaving an ambiguity that can be resolved only with some simple phase data. However, the practical problem will usually make obvious that one which is correct.

Contributors to Waves and Electrons Section

J. Alfred Davies

J. Alfred Davies was born in Atlantic City, N. J., on December 22, 1913. He received the A.B. degree in 1938 and the M.A. degree in mathematics in 1941, both from the University of Alabama. During the year 1940–1941, he was a graduate assistant in the mathematics department at the same University. From 1941 to 1942 he was an instructor in mathematics at the Engineering School, Mississippi State College, and also worked as a consulting statistician at the Mississippi Business Research Station.

In 1942, Mr. Davies was employed as a design engineer by the Ken-Rad Tube and Lamp Corporation, presently the Owensboro Tube Works, General Electric Company. Shortly after joining this company he became engaged in statistical quality control and product analysis, in which fields he is now working as a statistical engineer. He is a member of the American Society for Quality Control, Pi Mu Epsilon, Sigma Xi, and the Institute of Mathematical Statistics.

P. P. Eckersley

P. P. Eckersley (F'25) was born in January, 1892, in La Puebla, Mexico. He received the Certificate of Technology from the Manchester School of Technology in 1915, and the Higher Certificate from Bedales School, in 1910.

Mr. Eckersley was Vice-President of the IRE during 1936–1937. He was head of the experimental section, design department, of Marconi's Wireless Telegraph Company from 1919 to 1923, and chief engineer of the British Broadcasting Corporation from 1923 until 1929. He is now associated with the Telephone Manufacturing Company in Kent, England.

A member of the Institution of Electrical Engineers, he received the Premium Prize Award of that society in 1930.

T. T. Goldsmith, Jr.

Thomas T. Goldsmith, Jr. (A'38–VA'39–SM'46–F'49) was born in Greenville, S. C., on January 9, 1910. He attended the Furman University in Greenville, majoring in physics and mathematics. His graduate work was carried out at Cornell University, where he studied under, and acted as assistant to, Dr. Frederick Bedell. He worked as a physicist at the Biophysics Laboratory, Cold Spring Harbor, L. I., N. Y., for a period, where he completed his Cornell research work, and received the Ph.D. degree in September, 1936.

From November, 1936, to date Dr. Goldsmith has been associated with the Allen B. Du Mont Laboratories, Inc., as director of research. He is active on various committees of the Radio Manufacturers Association, the Society of Motion Picture Engineers, Television Broadcasters Association, and the AIEE. He has been active on the following IRE committees, among others: Admissions, Professional Groups, Television, and Electron Tubes.

Contributors to Waves and Electrons Section

Dr. Goldsmith holds membership in Sigma Pi Sigma, Sigma Xi, the American Physical Society, the Montclair Society of Engineers, and the Radio Club of America, in addition to those mentioned above.

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R. A. Millikan

Robert Andrews Millikan was born in Morrison, Ill., on March 22, 1868. He was awarded the A.B. degree from Oberlin College in 1891, and the Ph.D. degree from Columbia University in 1895. In 1895 and 1896 he also studied in the Universities of Jena, Berlin, and Göttingen. In 1896 he accepted a post at the University of Chicago as assistant in physics, and became successively instructor, assistant professor, associate professor, and professor of physics. He left the University of Chicago in 1921 to become director of the Norman Bridge Laboratory of Physics and chairman of the Executive Council at the California Institute of Technology.

In 1946 Dr. Millikan retired from his administrative position at the Institute, and in his new post as vice-chairman of the Board of Trustees is now acting in an advisory capacity while continuing his research work and writing. He is best known for his work on the isolation and measurement of the electron; the direct photoelectric determination of the fundamental radiation constant known as Planck's h; his study of Browning movements in gases; his discovery of the law of motion of a particle falling toward the earth after it enters the earth's atmosphere; the experimental study, with I. S. Bowen, of the spectroscopic properties of light atoms in all stages of "stripping"; the discovery, with Charles C. Lauritsen, of the laws governing the extraction of electrons from metals by fields alone; and the study of the nature and properties of cosmic rays.

Dr. Millikan is the author or joint author of several elementary and college text books in physics, and also in 1947 of a new and expanded edition of "Electrons (+ and −), Protons, Photons, Neutrons, Mesotrons and Cosmic Rays." In addition, he is the author of books of a philosophical nature, such as "Evolution in Science and Religion," "Time, Matter and Values," and "Science and the New Civilization." He is a member of the National Academy of Sciences, the American Philosophical Society, and the American Academy of Arts and Sciences, and is honorary member of many foreign scientific societies. He has been awarded, among others, the Nobel Prize in physics of the Royal Swedish Academy, the Hughes Medal of the Royal Society of Great Britain, and the Faraday Medal of the London Chemical Society.

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J. D. O'Neill

Jerry D. O'Neill was born in New York, N. Y., on January 31, 1918. From 1942 through 1945 he served in the U. S. Army Intelligence as a topographical expert. Since 1945 he has been associated with the Allen B. Du Mont Laboratories, Inc., in the capacity of field engineer. Mr. O'Neill is now engaged in work toward the B.S. degree at the Brooklyn Polytechnic Institute.

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J. H. Parsons

J. Howard Parsons (M'47) was born in Chicago, Ill., on February 1, 1920. He received his engineering education at Wilson Junior College, in Chicago, Ill., and at the University of Chicago. In 1941, he became a research assistant in the physics department at the University of Chicago, and joined the Metallurgical Laboratory of that University in February, 1942. In July, 1943, he transferred to the Clinton Laboratory, in Oak Ridge, Tenn., where he remained as a research associate until June, 1948. At that time he joined the staff of the University of California, at Los Alamos, N. M.

Mr. Parsons is chiefly engaged in the development of electronic instruments for radio-chemical assay.

M. C. Pease

Marshall C. Pease (M'47) was born in New York, N. Y., in 1920. He received the B.S. degree in chemistry from Yale University in 1940, and the M.A. degree in physical chemistry from Princeton University in 1943. He worked at the Radio Research Laboratory at Harvard University (Division 16, OSRD, a war laboratory for the development of countermeasures for radar), where he was principally concerned with the 1-kilowatt series of cw magnetrons. In 1945 he left to accept a commission in the U. S. Navy as a radar officer.

After his release from the Navy, Mr. Pease joined the electronic division of Sylvania Electric Products Inc. He is now in the tube development section as a theoretician and mathematician, analyzing present development programs. He is also in charge of test personnel of tube development.

❖

R. P. Wakeman

Robert P. Wakeman (M'46) was born in St. Johnsbury, Vt., on December, 16, 1913. He completed the ESMWT courses in ultra-high-frequency techniques I and II at Northwestern University. During 1942–1943 Mr. Wakeman was employed by the U. S. Army Signal Corps as a field engineer. In 1944 he joined the Radiation Laboratory at the Massachusetts Institute of Technology to work on pulse modulator development. From 1945 to the present time he has been in charge of field propagation studies for the Allen B. Du Mont Laboratories, Inc.

Mr. Wakeman has been active on several industry propagation committees.

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For a biography and photograph of Marvin Camras, see page 451 of the April, 1949, issue of the Proceedings of the I.R.E.

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Wave-Guide Modes—E. M. Wells. (*Marconi Rev.*, vol. 11, pp. 124–136; October to December, 1948.) A general elementary discussion, with perspective diagrams showing the field of configuration of various modes, for both rectangular and circular waveguides.

621.392.26† 946
On the Excitation of Waveguides: Part 3—A. A. Samarski and A. N. Tikhonov. (*Zh. Tekh. Fiz.*, vol. 18, pp. 971–985; July, 1948. In Russian.) The radiation resistance of a current is defined; methods of calculation are described and applied, with the aid of Poynting's theorem, to currents in a plane perpendicular to the waveguide axis and also to currents disposed arbitrarily. The results so obtained can also be derived by the method of induced electromotive forces; they are compared with the results of Schelkunoff and Slater. Part 2: 3335 of 1948.

621.392.26† 947
Reflection of Electric Waves at a Capacitive Diaphragm in a Rectangular Waveguide—K. Fränz. (*Arch. Elek.* (Übertragung), vol. 2, pp. 140–147; April and May, 1948.) The reflection coefficient can be derived by a simple frequency transformation from that for the corresponding case of a diaphragm between two conducting planes. The limiting frequency for the waveguide corresponds to the zero frequency for th plane case, for which the reflection coefficient is easily calculated from the electrostatic field.

621.392.26†:621.3.09 948
Wave Propagation in Nearly Circular Waveguides and Arrangements for Transmitting H_0 Waves round Bends—M. Jouguet. (*Câbles and Trans.* (Paris), vol. 2, pp. 257–284; October, 1948. With English summary.) Full paper. For the principal results see 2720 and 2724 of 1948.

621.396.67 949
On the Approximate Ground-Absorption Formula for Vertical Dipoles—K. F. Niessen. (*Physica, 's Grav.*, vol. 9, pp. 915–922; November, 1942. In German.) A formula previously given (19 of 1936) contained an error, as was shown by Sommerfeld and Renner (3583 of 1942). Corrected curves are now given.

621.396.67 950
On the Illustration of Dipole Radiation—F. Kiebitz. (*Arch. Elek.* (Übertragung), vol. 2, pp. 49–57; February and March, 1948.) The Hertzian solution of the field equations is expanded numerically and the results are used to obtain accurate pictures of the development of the radiation field near the dipole.

621.396.67 951
On the Choice of Integration Paths in the Problem relating to the Radiation of a Dipole above a Plane Earth—G. Eckart and T. Kahan. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 969–970; November 8, 1948.)

621.396.67 952
An Omnidirectional, Vertically Polarized, Four Element Antenna Array—R. S. Rettie. (*Canad. Jour. Res.*, sec. F, vol. 26, pp. 457–463; October, 1948.) Each element is a cylindrical $\lambda/2$ radiator surrounding a mounting pipe. A gain of 4 db over a $\lambda/2$ dipole is obtained. Operation is possible over a 10 per cent frequency band for a center frequency of 223 Mc.

621.396.67 953
Hallén's Theory for a Straight, Perfectly Conducting Wire, used as a Transmitting or Receiving Aerial—C. J. Bouwkamp. (*Physica, 's Grav.*, vol. 9, pp. 609–631; July, 1942. In English.) Hallén's general theory is simplified for the case of a straight uniform wire of infinite conductivity. From Maxwell's equations and the boundary conditions at the surface of the wire, the current distribution and hence the antenna field can be computed. The large constant $\Omega[=2 \text{ log } (2l/a)$, where $2l$ and $2a$ are, respectively, the length and diameter of the wire], is involved in a fractional formula for the antenna input impedance, both the numerator and denominator being given by power series in Ω^{-1}. Numerical values of the coefficients of Ω^{-1} and Ω^{-2} are given, so that for $0 \leq 2\pi l/\lambda \leq 5$, the antenna impedance can easily be computed ($\Omega > 10$). The wavelength and radiation impedance at the first and second point of resonance are given, the calculations including the second-order corrections for the finite diameter of the wire. If used as a receiving antenna, the wire is supposed to be parallel to the incoming electric field. An antenna form factor is defined. The loaded receiving antenna behaves as a generator with internal impedance equal to the input impedance of the wire when used as a transmitting antenna.

621.396.67 954
WEWS TV Antenna Installation—J. B. Epperson. (*Broadcast News*, pp. 66–71; October, 1948.) An RCA Type TT-3A Super Turnstile is fixed 1,526 feet above sea level, at the top of a 388-foot mast. Radiated power is 16.3 kw peak-to-peak for video and 10.3 kw for FM.

621.396.67 955
The Aerial in Television—A. Coudert and A. Orlowski. (*Radio Tech. Dig.* (Franç.), vol. 2, pp. 225–236; October, 1948.) General discussion of radiation patterns, antenna gain, feeders, and impedance matching, with illustrations of various types of television antennas. Part 2: 956 below.

621.396.67 956
The Aerial in Television—A. Coudert and A. Orlowski. (*Radio Tech. Dig.* (Franç.), vol. 2, pp. 287–294; December, 1948.) Discussion of a) multielement antennas, (b) a systematic study of reception conditions in and near Paris by means of a mobile receiver, (c) two types of French antenna, one being light and suitable for the Paris neighborhood while the other is designed for long-distance reception. Part 1: 955 above.

621.396.67 957
Multi-Slot Aerials—Ya. N. Fel'd. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1265–1272; October, 1948. In Russian.) Antennas consisting of nonsymmetrical slots cut in a closed metallic surface and excited from either the inside or the outside are considered. A system of integro-differential equations (7) is derived determining the voltage distribution along the slots for any method of excitation. A general method of solving these equations is proposed and the case of "tuned" antennas is discussed in detail. A formula (21) determining the radiated power of a transmitting antenna is also derived. For earlier work see 3344 of 1948.

621.396.671 958
The Patterns of Slotted-Cylinder Antennas—G. Sinclair. (Proc. I.R.E., vol. 36, pp. 1487–1492; December, 1948.) Formulas for the amplitude and relative phase of the field, and calculated patterns for one- and two-slot antennas of varying diameter are given. The results do not indicate relative gains. Increased directivity is obtained by increasing cylinder diameter or by suitable angular arrangement of two or more slots. Experimental results confirm calculated patterns.

621.396.671 959
Rhombic Aerial Design Chart—R. H. Barker. (*Radio Tech. Dig.*, (Franç.), vol. 2, pp. 263–271; December, 1948.) French version of 649 of April.

621.396.671 960
On the Theory of Coupled Antennae.—C. J. Bouwkamp. (*Phillips Res. Rep.*, vol. 3, pp. 213–226; June, 1948.) A treatment of coupling between two identical, parallel, perfectly conducting cylindrical wires, separated by a distance large compared with the radius of either, and each fed at the center by a "slice" generator. Approximate integral equations are obtained and solved for the currents in the two wires. Expressions are then derived for the input and mutual impedances. Some functions involved in the problem are tabulated. See also 3474 of 1941 (King and Harrison).

621.396.671 961
Aerial Absorption Surfaces and their Measurement for Decimetre and Centimetre Waves—R. Becker. (*Arch. Elek.* (Übertragung), vol. 2, pp. 120–123; April and May, 1948.) The absorption surface is defined and the formulas required for its determination are deduced. A method of measurement, based on principles given by Fränz (2898 of 1944), is described which only requires the measurement of two lengths and a single transmission-line voltage ratio. The absorption surface of a paraboloid of diameter 80 cm and focal length 20 cm, for $\lambda = 10$ cm, was found to be 2,800 cm², the so-called surface efficiency being 56 per cent.

621.396.675 962
Investigation of the Properties of a Ground Aerial—A. E. Pannenborg. (*Appl. Sci. Res.*, vol. B1, no. 3, pp. 213–240; 1948.) Theoretical results are compared with 3.5-Mc field-strength measurements. Measurements on ground waves agree very well with the theory, but the behavior of horizontally polarized waves is not fully understood.

621.396.676:621.396.93 963
Some Principles Underlying the Design of Aerial Systems for High-Frequency Radio Direction-Finders in H.M. Ships—Crampton, Struszynski, de Walden, and Redgment. (*See* 1075.)

621.396.67 964
Antenna Manual [Book Review]—W. Smith Editors and Engineers, Santa Barbara, Calif., 1948, 301 pp. (Proc. I.R.E., vol. 36, p. 1511; December, 1948.) ". . . recommended for those desiring a readable, elementary, and concise discussion of some of the more practical aspects of radio propagation antennas."

CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.3:621.392 965
On Transients in Homogeneous Ladder Networks of Finite Length—W. Nijenhuis. (*Physica, 's Grav.*, vol. 9, pp. 817–831; September, 1942. In English.) For all finite homogeneous ladder networks consisting of lumped elements, the solution can be written in such a way that the analogy with the solutions of

d'Alembert and Bernoulli is clearly apparent. By comparison of the two forms of solution, new approximations for some functions can be determined, such as Bessel functions in which the argument and order are equal. The formulas obtained are applied to a low-pass and a diffusion filter.

621.3.015.3:621.392.52 966
Repeated Integrals of Bessel Functions and the Theory of Transients in Filter Circuits—J. C. Jaeger. (*Jour. Math. Phys.*, vol. 27, pp. 210–219; October, 1948.) The solution for the simple high-pass filter, given by Carson and Zobel in 1923, involves a sum of repeated integrals of Bessel functions of order zero. This solution is regarded as a new "high-pass filter function," and is tabulated. Solutions for certain more complicated ladder networks can be expressed in terms of this function. The solutions of a number of transient problems in the ten simplest semi-infinite ladder networks can be expressed as single integrals involving tabulated functions.

621.3.018.4(083.74) 967
Ultrasonic Generator with Standard Frequency—A. Barone. (*Helv. Phys. Acta*, vol. 21, pp. 137 142; June 15, 1948. In German.) Details of a quartz-controlled 1-Mc generator with additional stages for 2 Mc and 4 Mc. The output from the 4-Mc stage is 18 to 20 watts. The variation of the fundamental frequency with the temperature of the crystal is determined by comparison of its tenth harmonic with WWV 10-Mc transmissions.

621.314.2 968
Methods for Designing High-Power Modulation Transformers—S. V. Person, M. A. Sobolev, and N. I. Eydlin. (*Radiotekhnika* (Moscow), vol. 3, pp. 3–23; September and October, 1948. In Russian.) Technical requirements imposed on modulation transformers are reviewed and two modulation circuits, one with and one without a modulation choke (Figs. 1 and 2), are considered. The design of transformers for each of the above circuits is discussed and a simple method is proposed for determining the maximum magnetic induction of the transformer so as not to exceed the permissible distortion. The design of the modulation choke is also examined. Numerical examples are given. The two modulation circuits are compared. A great saving in weight can be effected if a circuit with a choke is used in broadcasting transmitters. For commercial transmitters, the circuit without choke is preferable.

621.314.2:621.396.93 969
An Investigation of Symmetrical Screened Transformers for H.F. Radio Direction-Finders—W. Struszynski and J. H. Marshall. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 857–867; 1947.) Discussion of: (a) the equivalent circuit theory of such transformers, (b) the symmetry factor, (c) experimental evidence in support of the theory, (d) the principles of mechanical design, (e) methods of balancing, and (f) results of measurements of the symmetry factor for two typical transformers.

621.314.2.012 970
Diagrams for Output Transformer Design Calculations—J. Sommer. (*Funk und Ton*, vol. 2, pp. 549–563; November, 1948.) The diagrams enable transformer data to be found for a given power loss in the transformer and for a given lower limit of the frequency band to be used. In addition to the core dimensions, other details such as airgap, winding data, and size of wire are given. The characteristics of the output tube must be known. The diagrams apply to standard E and M stampings of grade-IV dynamo sheet and can also be used for filter-choke calculations.

621.314.3† 971
Magnetic Amplifiers—(*Elec. Times*, vol. 114, pp. 699–701; December 9, 1948.) Summary and discussion of two IEE papers entitled "Magnetic Amplifiers," by A. G. Milnes, and "A Theoretical and Experimental Study of the Series-Connected Magnetic Amplifier," by H. M. Gale and P. D. Atkinson.

621.316.727 972
On a Method of Phase-Shift Control for Auxiliary-Circuit Supplies—R. Matthäi. (*Brown Boveri Rev.*, vol. 35, pp. 157–161; May and June, 1948.) Essentially a simple bridge circuit for obtaining an ac voltage whose phase angle with respect to the supply can be made to have any value between 0 and 180°. Families of circle diagrams are derived for some characteristic cases, and their application is explained by means of numerical examples.

621.316.935.1:518.4 973
Graphical Iron Core Reactor Design—M. R. Whitman. (*Electronics*, vol. 21, pp. 136, 148; December, 1948.)

621.318.572 974
Coincidence Device of 10^{-8}–10^{-9} Second Resolving Power—Z. Bay and G. Papp. (*Rev. Sci. Instr.*, vol. 19, pp. 565–567; September, 1948.) For another account see 2471 of 1948.

621.318.572 975
Coincidence Circuit of Medium Resolution—H. L. Schultz and E. Pollard. (*Rev. Sci. Instr.*, vol. 19, pp. 617–620; October, 1948.)

621.318.572 976
Investigations on Various Coincidence Mixer Stages—P. Stoll, M. Walter, and W. Zünti. (*Helv. Phys. Acta*, vol. 21, pp. 177–179; August 10, 1948. In German.) Summary of Swiss Phys. Soc. paper.

621.318.572 977
On Scaling-Down Circuits—E. Baldinger and R. Casale. (*Helv. Phys. Acta*, vol. 21, pp. 117–130; June 15, 1948. In German.) All see-saw circuits can be used either as generators, flip-flops, or scaling-down circuits if the circuit constants are suitably chosen. The design of a simple multivibrator type of scaling-down circuit is discussed and approximate formulas are given for the calculation of the various circuit constants. An input stage is described suitable for counting voltage impulses of any wave form. A detailed circuit diagram is given for a scale-of-128 circuit with a resolving power of 1 μsecond.

621.319.4:551.57 978
The Effect of Humidity on the Calibration of Precision Air Capacitors—L. H. Ford. (*Jour. IEE* (London), part II, vol. 95, pp. 709–712; December, 1948.) Over the range of relative humidity 30 to 65 per cent, the total change in capacitance can be about 3 parts in 10^4 at room temperature.

621.392 979
On the Properties of the General Mesh Network with Structural Symmetry—G. Nasse. (*Compt. Acad. Sci.* (Paris), vol. 227, pp. 1350–1352; December 20, 1948.) Matrix theory for a system satisfying the following conditions: (a) it has two groups of $n+1$ terminals; (b) it contains $m \times n$ independent meshes and constitutes the extension, to a system of n phases, of a network of m geometrically independent meshes; (c) it possesses a symmetry of structure such that a circular permutation of the voltages applied to its terminals causes an identical permutation of the currents entering by these terminals.

621.392 980
The Gyrator, a New Electric Network Element—B. D. G. Tellegen. (*Philips Res. Rep.*, vol. 3, pp. 81–101; April, 1948.) The "gyrator" is a linear, constant, passive network element which violates the reciprocity relation. It "gyrates" a current into a voltage or vice versa. Network synthesis is much simplified by including such elements, which can be realized by means of a medium consisting of particles carrying both permanent electric and permanent magnetic dipoles, or by means of a type of gyromagnetic effect in a ferromagnetic medium.

621.392.33 981
A Universal Adjustable Transformer for U.H.F. Work—J. M. van Hofweegen and K. S. Knol. (*Philips Res. Rep.*, vol. 3, pp. 140–155; April, 1948.) A device for matching any two impedances by the proper adjustment of two shorting bridges. It consists of a screened 2-wire Lecher system asymmetrically loaded with respect to the screen. The device can also be used for rough impedance measurements. A universal adjustable waveguide transformer based on the same principle is also discussed.

621.392.43 982
A Single-Control Variable-Frequency Impedance-Transforming Network—A. Bark. (PROC. I.R.E., vol. 36, pp. 1535–1537; December, 1948.) Mathematical analysis of a coaxial-line network, based on the stub matching method, which will maintain perfect matching of generator to load over a wide range of frequencies by adjustment of a single reactance shunted across the load. Design formulas are derived and an equivalent circuit is illustrated. Transformers of this type have been built to match antenna cables to mixers of color television receivers operating over the range 480 to 920 Mc. Noise and losses due to sliding contacts are avoided. By ganging the shunt reactance to the local oscillator, matching was automatically achieved at any frequency in the tuning range.

621.392.5 983
The Calculation of Attenuators—K. Martin. (*Funk und Ton*, vol. 2, pp. 591–595; November, 1948.) Formulas and tables are given which greatly simply the determination of the values of the various components of Tee and pie-attenuators.

621.392.5 984
Insertion and Echo Loss for No-Load Quadripole—W. Herzog. (*Arch. Elek.* (Übertragung) vol. 2, pp. 84–87; February and March, 1948). Formulas for these losses are given and applied to a crystal band-filter.

621.392.52 985
Insertion Characteristics of Filters. Part 2 "Constant-*k*," M-Derived' Sections and Composite Filters—J. B. Rudd. (*AWA Tech. Rev.*, vol. 8, pp. 77–95; October, 1948.) Previous results (1697 of 1947) are modified to apply to multisection, constant-*k* filters terminated by resistances other than their design resistance. Expressions are developed for *m*-derived single-section filters and for composite filters consisting of prototype sections together with *m*-derived sections. Insertion loss and insertion phase-shift characteristics for some composite filters are shown graphically and formulas for the loss and phase shift are summarized and tabulated. The frequency variable used is such that the results obtained may be applied to low-pass, high-pass, and confluent band-pass filters.

621.396.611 986
On a Simple Statistical Property of an Ensemble of Linear Harmonic Oscillators—R. Kronig. (*Physica*, 's *Grav.*, vol. 9, pp. 113–116; January, 1942. In English.) For a system of oscillators of given frequency in thermal equilibrium, the distribution of the deviations from "rest" state is Gaussian.

621.396.611:621.316.726 987
Practical Notes on the Maintenance of Frequency Constancy for Spherical Resonators—K. F. Niessen. (*Physica*, 's *Grav.*, vol. 9, pp. 768–772; July, 1942. In German.) For the suppression of frequency variations due to expansion caused by Joule heating of the resonator walls, two alternative methods are recommended: (a) fixing the resonator between two diametrically opposed blocks whose separation does not vary with temperature, with the

halves of the dipole at the ends of the diameter joining the blocks; (b) fixing a ring of invariable diameter round the resonator, with the halves of the dipole at the ends of a diameter of the ring. See also 992 below.

621.396.611.1 988
On the van der Pol Oscillator—N. Minorsky. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 60–61; January 3, 1949.) Theoretical discussion of the oscillator defined by the equation

$$\ddot{x} - \epsilon(1 - x^2)\dot{x} + x = 0.$$

621.396.611.1:518.4 989
Graphical Determination of Oscillation Phenomena in Oscillatory Systems by Means of Complex Representation—H. Kleinwachter and H. Wojtech. (*Arch. Elek.* (Übertragung), vol. 2, pp. 69–75; February and March, 1948.) A particular advantage of the method is the fulfilment of the boundary conditions by the use of a logarithmic spiral as the solution function. When damping is absent, the spiral degenerates into a circle. Use of a sufficiently finely divided step function for the input signal enables an approximate solution of the differential equation to be obtained.

621.396.611.21 990
The Equivalent Circuit of the Curie Double Strip—E. J. Post. (*Appl. Sci. Res.*, vol. B1, no. 3, pp. 168–180; 1948.) A dynamic theory is given. The equivalent constants for the bending mode are closely related to those of the corresponding longitudinal vibration. Results based on the theory are compared with Dutch Post Office experimental results and with those of Lane (1582 of 1946).

621.396.611.21 991
Methods for Varying the Resonance Frequency of Crystal Oscillators—W. Herzog. (*Arch. Elek.* (Übertragung). vol. 2, pp. 153–163; April and May, 1948.) Discussion of the limits of the possible frequency variation and also of variation arrangements making use of capacitors or inductors, either in series or in parallel with the crystal. With some of these methods, the frequency variation can be extended considerably.

621.396.611.4 992
Mean Frequency Stability of Cavity Resonators—K. F. Niessen. (*Physica*, *'s Grav.*, vol. 9, pp. 145–157; February, 1942. In German.) A general discussion for the case where deformation of the resonator is limited to extension. For the sphere and cube, the stability differs very little from that for deformations such that the surface area remains constant. See also 3369 of 1948 and 993 below.

621.396.611.4 993
On the Frequency Stability of Some Cavity Resonators in an Electrical Circuit—K. F. Niessen. (*Physica*, *'s Grav.*, vol. 9, pp. 539–546; June, 1942. In German.) Calculations for the sphere and the cube show that, in general, the cube shows the smaller frequency variations for small irregular deformations. See also 992 above.

621.396.611.4 994
Nodal Planes in a Perturbed Cavity Resonator: Part 1—K. F. Niessen. (*Appl. Sci. Res.*, vol. B1, no. 3, pp. 187–194; 1948.) The resonator considered is a rectangular cavity of square cross section, with one of its walls rotated through a small angle about its edge, the electric vector being parallel to this edge. The change in resonance frequency, and the distortion of the electric field, for the fundamental vibration, due to the asymmetry of the cavity, are investigated mathematically.

621.396.611.4 995
On the Forced Electro-Magnetic Oscillations in Spherical Resonators—O. E. H. Rydbeck. (*Phil. Mag.*, vol. 39, pp. 633–644; August, 1948.) Electromagnetic oscillations in spherical systems are studied as functions of the exciting source, and the solutions obtained show the degree of excitation of the various modes and orders. Transverse electric *TE* waves and transverse magnetic *TM* waves are studied separately, and the equivalent network for a resonator with two current loops is shown.

621.396.611.4 996
A Method for Approximate Calculation of Natural Wavelengths of Cavity Resonators of Irregular Shapes—G. V. Kisun'ko. (*Radiotekhnika* (Moscow), vol. 3, pp. 24–35; September and October, 1948. In Russian.) The resonator under investigation is divided into a number of simple regions and the proper (eigen) functions of the resonator are expressed approximately in terms of the proper functions of these regions. Conventions necessary in using this method are discussed and a conception of average boundary conditions between the regions is introduced. The case of free electromagnetic oscillations in dielectric volumes bounded by ideally conducting cylindrical envelopes is considered in detail. The method is also applied to a system of parallel cylinders coupled by longitudinal slots.

621.396.611.4 997
Experiments on Cylindrical Cavity Resonators—D. D. Mansion. (*Rev. Telegr.* (Buenos Aires), vol. 37, pp. 631–634, 670; September, 1948.) Results of measurements are given for the resonance wavelength of a resonator as a function of the length and diameter of a central piston, and for the Q factor for various cylinder lengths. A double-beat method for measurement of the frequency excursion is also described.

621.396.615:537.525.92 998
Space-Charge Control in the Transit-Time Region with High Degree of Modulation—F. W. Gundlach. (*Funk und Ton*, vol. 2, pp. 407–419, 454–465, and 516–533; August to October, 1948.) Analysis for the plane diode, assuming that the initial velocity of electrons leaving the cathode is zero, that saturation does not occur, and that the current flowing across the discharge path is sinusoidal. For an oscillating diode, the optimum efficiency is only about 3 per cent, so that the resonance resistance of the attached resonator must be about 4 times greater than that necessary for oscillation. With space-charge control, the slope of the characteristic falls with increased transit angle and with increased modulation; the control power decreases, on the average, with increased transit angle. In the transit-time region, a diode can absorb real power when the plate voltage is made so negative that the plate dc vanishes. Theoretical results are supported by calculation of electron paths, agreement being exact in many cases and closely approximate in others. The degree of approximation is discussed and a complete review of the physical phenomena in in the space-charge diode is given. The behavior of the electron current and of the ac and dc voltages is shown graphically and the special case where the plate current has no dc component is discussed.

621.396.615.029.63 999
The Self-Excitation of a Triode Oscillator with Feedback in the Decimetre Wavelength Range—S. D. Gvozdover and V. A. Zore. (*Zh. Tekh. Fiz.*, vol. 18, pp. 1194–1206; September, 1948. In Russian.) The self-excitation of a triode is investigated mathematically, taking into account the cathode-grid transit time. General formulas are derived for determining the wave length of the oscillator, the condition of self-excitation, and the frequency correction necessary for ultra-short-wave operation. The theory is illustrated by an analysis of the self-excitation of the Esau circuit (Fig. 4).

621.396.619.13:621.396.621 1000
Crystal Discriminators—J. M. Barcala. (*Rev. Telegr.* (Buenos Aires), vol. 37, pp. 611–615, 648; September, 1948. (Discussion of their characteristics, with special reference to (a) operating voltages and their variation with frequency, (b) sensitivity, (c) effect of the Q factor of the tuned circuit, (d) effect of voltage ratio on output.

621.396.619.13:621.396.621 1001
A Phase Discriminator for Frequency-Modulation Reception—F. G. Newall and J. G. Spencer. (*Electronic Eng.*, vol. 21, pp. 25–26; January, 1949.) Description of a circuit which is easy to align and produces an output voltage high enough to drive a normal output tube. A circuit diagram is given and design data are discussed.

621.396.645+621.392.52 1002
The Properties of I.F.-Amplifier Networks—A. Lennartz. (*Funk und Ton*, vol. 2, pp. 579–590; November, 1948.) Equivalent circuits are used in discussion of various types of amplifiers and filters.

621.396.645 1003
Study and Construction of an Amplifier of Very High Quality: Parts 1 and 2—F. Gilloux. (*Radio Prof.* (Belgium), vol. 11, pp. 13–18 and 16–18; April and October, 1948.) Discussion of the design of the different stages of an amplifier in which phase distortion is low and which has an extended range of uniform response. Construction details are given.

621.396.645:621.316.722.4 1004
Low-Impedance Variable Voltage Tappings—M. G. Scroggie. (*Wireless World*, vol. 55, pp. 2–6; January, 1949.) A cathode follower can be controlled to provide a continuously variable output when shunted across a stabilized dc supply voltage. The cathode follower has an internal resistance which for most applications is >1 per cent of the load resistance into which it has to work. Circuit and operational details are discussed. The limitations on the output are those imposed by the characteristics of the tube used and its maximum loadings. For previous parts see 231 of February, 863 of April and 1200 below.

621.396.645:621.385:621.396.822 1005
On the Influence of the Noise of Vacuum Tubes on the Accuracy of Linear Amplifiers—Milatz and Keller. (*See* 1229.)

621.396.645:621.396.931 1006
Power Amplifier for the Citizens Transmitter—W. C. Hollis. (*Electronics*, vol. 21, pp. 84–87; December, 1948.) Construction and circuit details of a two-stage amplifier for use with the transmitter discussed in 855 of 1948. No machining is necessary. See also 3511 of 1948 and back references.

621.396.645.001.8:535.6 1007
Applications of Alternating Current Amplifiers to Optical Measurements—E. J. Harris. (*Electronic Eng.* (London), vol. 20, pp. 396–399; December, 1948.) An absorptiometer and two spectrometers are briefly discussed. Requirements concerning detector impedance, switching frequency of an interrupted beam, and tuning are considered.

621.396.645.029.4:621.314.2 1008
Note on A.F. Amplification: the Question of the Transformer—H. Gilloux. (*Radio Franç.*, pp. 10–13; October, 1948.) An extremely simple method of design is given for the transformer of an output stage using a single power tube or two smaller tubes in parallel. Examples show that for a corresponding push-pull stage, the transformer is very much heavier and about 4 times dearer. See also 1321 of 1948.

621.396.645.37 1009
Cascade-Connected Feedback Amplifiers—H. Mayr. (*Microtecnic* (Lausanne), vol. 11, pp. 174–178; August, 1948. In English.) The complex attenuation versus frequency characteristic of an n-stage amplifier is discussed, particu-

larly for $n \leq 4$. It is not possible to obtain a rectangular attenuation curve if all the stages are tuned to the same frequency, while detuning the stages involves a considerable reduction in gain. A better solution is offered by subdividing the whole amplifier into groups of one or two stages, each with its own feedback loop; this possibility is discussed in detail for 3- and 4-stage amplifiers.

621.396.645.37:621.3.011.3/.4 1010
High-Q Variable Reactance—J. N. Van Scoyoc and J. L. Murphy. (*Electronics*, vol. 22, pp. 118–122; January, 1949.) The basic circuit consists of a feedback amplifier with a reactance in the feedback loop. The Q of the circuit is high if the amplifier output impedance is small compared with the feedback reactance. The over-all reactance may be varied by changing the amplifier gain, which may be controlled by an applied signal, by a potentiometer, *or* by changing tube transconductance *or* load resistance in one *or* more stages of the amplifier. A double-triode cathode-follower circuit is discussed in which the second cathode follower is used as a variable load resistance for the first.

621.396.619.231:621.396.645.37 1011
Valve Characteristic Giving Linear Modulation when a Feedback Resistor Is Inserted in the Cathode Lead—Boelens. (*See* 1244.)

621.396.645.371 1012
Impedances in the Feedback Amplifier—P. M. Prache. (*Bull. Soc. Franç. Élec.*, vol. 8, pp. 531–535; November, 1948.) Discussion on 383 of 1948. See also 1329 of 1948.

621.396.645.371 1013
Corrector Circuits for Feedback Amplifiers—H. Cheireix. (*Bull. Soc. Franç. Élec.*, vol. 8, pp. 523–531; November, 1948.) Various circuits are considered, comprising both reactors and resistors, which can be used for control of phase, gain, and pass band.

621.396.645.371:621.317.715 1014
Influence of Reactive Feedback Networks on the Response of Galvanometers—P. Savic. (*Nature* (London), vol. 162, pp. 569–570; October 9, 1948.) The damping and/or natural frequency of a galvanometer can be altered to an extent proportional to the gain of an associated amplifier which has a suitable reactive feedback circuit.

621.396.69:06.064 1015
Amateur Radio Show—(*Wireless World*, vol. 55, pp. 13–14; January, 1949.) Brief description of some of the components exhibited.

621.392 1016
Microwave Transmission Circuits [Book Review]—G. L. Ragan (Ed.). McGraw-Hill, New York, N. Y., 1948, 716 pp., $8.50. (Proc. I.R.E., vol. 36, p. 1511; December, 1948.) Volume 9 of the MIT Radiation Laboratory series. "... deals with the problems of power transmission from one place to another at microwave frequencies, and its contents are applicable to almost all of the problems that come up in the design of microwave circuits."

GENERAL PHYSICS

53.081+621.3.081 1017
The Rationalized Giorgi System with Absolute Volt and Ampere as Applied in Electrical Engineering—P. Cornelius. (*Philips Tech. Rev.*, vol. 10, pp. 79–86; September, 1948.) The most important electromagnetic formulas are tabulated and discussed.

530.162:519.2 1018
A Problem on Random Vectors—R. D. Lord. (*Phil. Mag.*, vol. 39, pp. 66–71; January, 1948.) An analysis of the probability distribution of the component, along a given axis, of the sum of a number of equal coplanar vectors whose directions are random. The solutions are expressed in terms of Bessel functions and are generalized for a system in more than two dimensions. See also 3566 of 1947 (Horner).

535:52 1019
Some Recent Applications of Optics to Astronomy—E. H. Linfoot. (*Mon. Not. R. Astr. Soc.*, vol. 108, no. 1, pp. 81–93; 1948.) Includes a discussion of the Schmidt optical system.

535.61-15 1020
On the Range of Infra-Red Rays—W. Dechend. (*Elektron Wiss. Tech.*, vol. 2, pp. 255–259; November, 1948.) Discussion of experimental results on the relation between the range and (a) the power of the source, (b) the water-vapor content of the atmosphere, and (c) the reflecting power of objects.

536.3 1021
The Efficiency of Radiation Shields—A. E. De Barr. (*Rev. Sci. Instr.*, vol. 19, pp. 569–573; September, 1948.) "The effect of a number of thick radiation shields of infinite thermal conductivity and of a type realizable in practice is analyzed by matrix algebra, and a method for calculating the temperatures of the various shields is also given." See also 1022 below.

536.3 1022
Extension of De Barr's Analysis of Radiation Shielding—J. B. Garrison and A. W. Lawson. (*Rev. Sci. Instr.*, vol. 19, pp. 574–577; September, 1948.) Extension of De Barr's analysis (1021 above) to the case when the original source has arbitrary emissivity.

537.52 1023
Mechanism of Dielectric Breakdown—D. T. Hurd. (*Gen. Elec. Rev.*, vol. 51, pp. 26–33; December, 1948.) A theoretical discussion for gaseous, liquid, and solid dielectrics.

537.525.92 1024
Space-Charge Wave Amplification Effects—A. V. Haeff. (*Phys. Rev.*, vol. 74, pp. 1532–1533; November 15, 1948.) A search for more efficient means of generating and amplifying microwave energy has suggested that the mechanism of interaction between particles and associated space-charge waves is of primary importance in many natural phenomena. The theory of the effect is outlined and its application to effects such as solar noise, excess noise in electron beam tubes and magnetrons, and high temperatures of electron clouds in magnetrons, is considered.

538:061.3 1025
Magnetism—D.A.O. (*Nature* (London), vol. 162, pp. 799–801; November 20, 1948.) Summaries of 3 papers read at a British Association Symposium.

538.566.029.65/.66† 1026
The Limiting Region between Electromagnetic Millimetre Waves and the Long-Wave Infra-Red—F. X. Eder. (*Funk und Ton*, vol. 2, pp. 491–498; October, 1948.) A review of methods for generating such waves.

538.569.4:[546.441.26-145.1 +546.413.1-145.1 1027
Absorption of Ultra High Frequency Waves in Salt Solutions—S. K. Chatterjee and B. V. Sreekantan. (*Indian Jour. Phys.*, vol. 22, pp. 325–332; July, 1948.) Measurements at frequencies between 300 and 480 Mc, using aqueous solutions of $MgSO_4$ and $CaCl_2$ of different concentrations, show that absorption maxima generally shift to higher concentrations with increasing frequency. The reflection coefficient varies little with either concentration or frequency. The results agree with the theory outlined. Experimental details were discussed in 3099 of 1948.

538.569.4.029.65:546.21 1028
Atmospheric Absorption of Millimetre Waves—H. R. L. Lamont. (*Proc. Phys. Soc.*, vol. 61, pp. 562–569; December 1, 1948.) A brief account of this work was noted in 3398 of 1948. The maximum absorption observed was 15.7 db/km, after reduction to standard dry atmospheric conditions.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

521.1 1029
Review of Cosmology—H. Bondi. (*Mon. Not. R. Astr. Soc.*, vol. 108, no. 1, pp. 104–120; 1948.)

523.2 1030
The Origin of the Solar System—H. Jeffreys. (*Mon. Not. R. Astr. Soc.*, vol. 108, no. 1, pp. 94–103; 1948.)

523.53:621.396.9 1031
Determination of Meteor Radiants by Observation of Radio Echoes from Meteor Trails
J. A. Clegg. (*Phil. Mag.*, vol. 39, pp. 577–594; August, 1948.) By noting the variations in the ranges in addition to the variations of the hourly rates, it is possible to reduce the number of observing stations from three to one. Using a frequency of 70 Mc and an antenna system whose half-amplitude beam-width is $\pm 10°$, under favorable conditions the radiant can be located within a circle of radius 1°, but usual y the probable errors are greater. See also 3402 of 1948 (Lovell).

523.61 "1947" 1032
Comets [in 1947]—G. Merton. (*Mon. Not. R. Astr. Soc.*, vol. 108, no. 1, pp. 124–130; 1948.)

523.72:621.396.822:551.510.535 1033
Ionospheric Effects Noted during Dawn Observations on Solar Noise—R. Payne-Scott and L. L. McCready. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 429–432; December, 1948.) A small source of solar noise is observed from a high site at dawn. The direct wave and that reflected from the surface of the sea give an interference pattern from which the apparent elevation of the source can be calculated. The difference between the elevations on 200 and 60 Mc gives the ionospheric refraction on the lower frequency and is of the order of $\frac{1}{2}°$. The observed refraction is about three times that expected assuming a symmetrical F layer, but the change with elevation favors an asymmetrical F layer rather than a possible G layer. On several days, noise on 60 Mc completely disappeared for some minutes before returning with reduced amplitude. This is tentatively attributed to absorption or refraction effects. See also 412 of 1948 (McCready, Pawsey, and Payne-Scott).

523.745 1034
Maximum of the Current Cycle of Solar Activity—V. F. Chistyakov. (*Priroda*, p. 41; August, 1948. In Russian.)

523.746:621.396.822 1035
Changes in Ionization and Radio Reception during the Sunspot-Period 1944–1947—H. T. Stetson. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 449–454; December, 1948.) Discussion of the relation between sunspot number and field strength for 5-Mc and 10-Mc transmissions over a 373-mile path. The night field for 5 Mc had a maximum at a sunspot number of 105, while day fields decreased steadily with increasing sunspot number. For 10-Mc transmissions, both day and night fields increased with increasing sunspot number.

523.746 "1947" 1036
Solar Activity: Sunspots [in 1947]—H. W. Newton. (*Mon. Not. R. Astr. Soc.*, vol. 108, no. 1, pp. 122–123; 1948.

523.746 "1948.07/.09" 1037
Provisional Sunspot-Numbers for July to September 1948—M. Waldmeier. (*Terr. Mag. Atmo. Elec.*, vol. 53, p. 348; December, 1948.)

523.752 "1947" 1038
Solar Activity: Prominences [in 1947]—A. K. Das. (*Mon. Not. R. Astr. Soc.*, vol. 108, no. 1, pp. 123–124; 1948.)

538.12:521.15 1039
Magnetism of Celestial Bodies and Gravitation—J. Mariani. (*Nature* (London), vol. 162, pp. 612–613; October 16, 1948.) Discussion of Blackett's formula (3112 of 1947) and Wilson's relation $\sigma=\rho\sqrt{K}$, where the charge density σ is associated with the mass density ρ by the Newtonian gravitational constant K. The Newtonian theory of gravitation fails to explain Wilson's formula. A suggested explanation in terms of relativistic theory is given for the case of an isolated and limited fluid surrounded by a static gravitational field.

538.12:521.15:523.7 1040
Solar Magnetism and the Suggested Fundamental Magnetization by Rotation—S. Chapman. (*Mon. Not. R. Astr. Soc.*, vol. 108, no. 3, pp. 236–251; 1948.) The assumption (discussed by Blackett in 3112 of 1947) that a rotatory mass flux f produces a magnetic field as if it were a negative electric current of intensity $\beta G^{\frac{1}{2}}f/c$ is applied to the sun. Here β is a constant of order unity (0.3 for the earth), G is the gravitational constant, and c the velocity of light. The solar density distribution assumed is that of a point convective model having a nonuniform distribution of angular velocity. If the sun's polar magnetic intensity is of the order of 10 gauss, as suggested by Thiessen, β for the sun is about 0.4. The magnetic intensity at the center of the sun appears to be of the same order of magnitude as sunspot magnetic fields, but these fields may be due to some other mechanism, and this weakens the hypothesis that β is a universal constant. The nonuniformity of angular velocity introduces terms into the field potential which are too small to be observable at present.

550.38:523.78 "1947.05.20" 1041
Magnetic Effects Observed at Vassouras, Brazil, during the Solar Eclipse of May 20, 1047—L. I. Gama. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 405–428; December, 1948.)

550.38 "1948.04/.06" 1042
Selected Days, Preliminary Mean *K*-Indices and *C*-Numbers for Second Quarter, 1948—N. F. Eaton. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 477, 478; December, 1948.)

550.38 "1948.07/.09" 1043
Daily Magnetic-Activity Figures *C* and Three-Hour-Range Indices *K* and List of Sudden Commencements, July to September, 1948 at Abinger—H. Spencer Jones. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 479–480; December, 1948.)

550.38 "1948.07/.09" 1044
Cheltenham [Maryland] *K*-Indices for July to September, 1948—P. G. Ledig. (*Terr. Mag. Atmo. Elec.*, vol. 53, p. 404; December, 1948.)

550.384.3 1045
Pre-History of the Earth's Magnetic Field—E. A. Johnson, T. Murphy, and O. W. Torreson. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 349–372; December, 1948.) A study of anomalous deposits in glacial clays and of Pacific cores suggests that the earth's field has remained substantially constant for the last million years.

550.385 "1948.07/.09" 1046
Principal Magnetic Storms [July–September, 1948]—(*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 481–494; December, 1948.)

551.5:538.566 1047
Some Problems in Radio Meteorology—Booker. (*See* 1159.)

551.510.535 1048
On the Structure of the Ionosphere—J. Malsch. (*Arch. Elek.* (Übertragung), vol. 2, pp. 58–69; February and March, 1948.) If radio waves are reflected from the F layer, or from the E layer and affected by the lower part of that layer or by one still lower, path differences in the reflected beam give rise to interference effects which cause variations of the received field strength. The effect of limitation of the beam on zenith-reflection measurements of intensity by pulse methods is also considered.

551.510.535 1049
The Induction of Electric Currents in a Non-Uniform Ionosphere—A. A. Ashour and A. T. Price. (*Proc. Roy. Soc. A*, vol. 195, pp. 198–224; December 7, 1948.) Calculations are made of the distribution and the magnetic field of the currents induced in a nonuniformly conducting ionospheric shell by an external magnetic field, which is either periodic or subject to sudden changes. Assuming that the initial phase of magnetic storms is due to field changes outside the ionosphere, it is shown that its mean integrated conductivity is probably not much greater than 10^{-7} e.m.u. It is found that electromagnetic shielding by the ionosphere has an important effect on the distribution of field changes observed on the earth, and may lead to an apparent diurnal variation of frequency of occurrence of sudden commencements at a given station. Simple explanations are suggested for some known features of micropulsations, and for some well-known phenomena of magnetic disturbance, including Sangster's rotating disturbance vector.

551.510.535 1050
An Approximate Solution of the Problem of Path and Absorption of a Radio Wave in a Deviating Ionosphere Layer—J. E. Hacke, Jr., and J. M. Kelso. (PROC. I.R.E., vol. 36, pp. 1477–1481; December, 1948.) The method of approximations previously used for the case of vertical incidence (3115 of 1948) is extended to obtain solutions for oblique incidence, neglecting second-order absorption effects.

551.510.535:523.746 1051
Critical Frequencies, Sunspots, and the Sun's Ultra-Violet Radiation—C. W. Allen. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 433–448; December, 1948.) Monthly values of the ratio A of the critical frequency to the critical frequency for zero sunspot number have been averaged for a number of stations and tabulated for the 11-year period 1937 to 1947. Certain solar features have variations which are correlated with sunspot-number variations but lag behind them; the lag varies from 0.15 month for Ca flocculi to 0.87 month for coronal line.

The relationship between relative ultra-violet intensity S, sunspot number R and the ratio A is of the form $S=(1+bR)=A^n$ where b and n are constants. Values of b and n obtained from the 1937 to 1947 observations for the E, F_1, and F_2 regions are compared with those determined from the relation between critical frequency and the sun's zenith distance.

The recombination coefficients for $R=0$ are found for the E and F_1 regions; the decay coefficient for the F_2 region is found from observations at stations which are relatively free from F_2-region anomalies.

551.510.535:525.624 1052
Atmospheric Tides in the Ionosphere: Part 3—Lunar Tidal Variations at Canberra—D. F. Martyn. (*Proc. Roy. Soc. A*, vol. 194, pp. 429–444; November 9, 1948.) Semi-diurnal lunar variations have been found in the heights and critical frequencies of the E, F_1, and F_2 regions; the harmonic coefficients have been determined. It is deduced that (a) the ionospheric lunar variations are caused by ionic drift under the action of the "dynamo" electric forces, and not by simple tidal rise and fall of isobaric surfaces; (b) the lunar magnetic variations are not produced in the E, F_1, or F_2 regions. Part 1: 2421 of 1947. Part 2: 106 of 1948. Part 4: 1053 below.

551.510.535:525.624 1053
Atmospheric Tides in the Ionosphere: Part 4—Studies of the Solar Tide, and the Location of the Regions Producing the Diurnal Magnetic Variations—D. F. Martyn. (*Proc. Roy. Soc. A*, vol. 194, pp. 445–463; November 9, 1948.) Solar tidal effects are found in the E, F_1, and F_2 regions. For the F_2 region, the amplitudes and phases of certain seasonal semi-diurnal harmonics are determined and used to interpret the global distribution of F_2 ionization. The lunar magnetic variation appears to be produced mainly by currents in the D region, but is opposed by corresponding currents in the E and F_1 regions. This conclusion is checked by considering McNish's evidence of the effects of solar flares on the magnetic variation (10 of 1938). Part 3 1052 above.

551.510.535:551.524 1054
Temperature of the Upper Layers of the Atmosphere—V. A. Baranul'ko. (*Priroda*, pp. 34–35; May, 1948. In Russian.) Kessenikh and Bulatov suggested the possibility of a relationship between the temperature on the surface of the earth and that of the F_2 layer (2520 of 1945). This was confirmed by Seaton (3123 of 1947) who showed that, at noon, high temperatures are probable in the E layer (height about 100 km) and that much lower temperatures are probable in the F_1 and F_2 layers (respective heights 200 and 350 km). The F_2-layer temperature seems to be the nearest to that at the surface of the earth. Since there is a definite relationship between the temperature and the ionization density of the reflecting layer, this conclusion is of great practical importance in selecting optimum wavelengths for communication in particular directions.

551.594.11 1055
Studies on the Atmospheric Potential Gradient: Part 1—The Principle of Selection of Electrostatically Quiet Days—H. Israël and G. Lahmeyer. (*Terr. Mag. Atmo. Elec.*, vol. 53, pp. 373–386; December, 1948. In German.) General discussion. It is suggested that in addition to potential gradient measurements, the conductivity and/or vertical current should also be taken into account.

551.594.6:621.396.93 1056
Sferics—C. V. Ockenden. (*Met. Mag.*, vol. 76, pp. 78–84; April, 1947.) With four stations operating cathode-ray-tube direction-finding equipment at a frequency of about 10 kc, the British Meteorological Office is able to locate thunderstorms at distances up to 1,000 or 1,500 miles. The equipment is briefly discussed, and results for certain specific days are correlated with other synoptic information.

LOCATION AND AIDS TO NAVIGATION

621.396.9:523.53 1057
Determination of Meteor Radiants by Observation of Radio Echoes from Meteor Trails—Clegg. (*See* 1031.)

621.396.93 1058
Airborne Automatic Direction-Finders—K. F. Umpleby. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 693–704; 1947.) Discussion of the general principles of direction finders which automatically rotate their loops to the null position; three such systems are described. The American AN/ARN7 radio compass uses an iron-cored loop which is driven by a 2-phase motor supplied from a saturable-core transformer. The saturation is controlled through thyratrons, by the receiver output. The German Peilgerät VI system also uses an iron-cored loop, driven by a dc motor which is controlled, through a Ward-Leonard system and vibrator rectifier, by the receiver output. In the Royal Aircraft Establishment (R.A.E.) direction finder now being developed, the iron-cored loop is driven, through a differential gear, by two continuously running motors. The loop is rotated when the motor speeds differ; static friction is thus eliminated.

621.396.93 1059
Radio Direction-Finding by the Cyclical Differential Measurement of Phase—C. W.

Earp and R. M. Godfrey. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 705–721; 1947.) If a single vertical antenna element describes a circular path in a horizontal plane, the phase of the signal received will be modulated according to the position of the antenna relative to the direction of arrival of the signal. The same effect can be obtained by successive electronic switching to individual antennas of a set placed at regular intervals round the circumference of a circle. Several types of direction finder using this system, and their advantages over the simple Adcock system, are discussed. Site errors can thus be appreciably reduced.

621.396.93 1060
The Extension of Wireless Direction-Finding Techniques to Very High Frequencies for Naval Use—R. M. Griffith and W. Rosinski. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 727–740; 1947.) Discussion of: (a) a rotating H-Adcock system suitable, with slight modification of the antenna dimensions, for different bands in the range 30 to 600 Mc, (b) a fixed 4-antenna Adcock system for use with a goniometer for the range 30 to 100 Mc, (c) a similar system with visual indication, and (d) a rotating reflector system for frequencies above 150 Mc. Typical bearing correction curves are given for installations in a ship.

621.396.93 1061
An Analysis of the Performance of Multi-Aerial Adcock Direction-Finding Systems—P. G. Redgment, W. Struszynski, and G. J. Phillips. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 751–761; 1947.) A generalized modification of the conventional 4-antenna Adcock system is considered in which each antenna is replaced by twin elements connected in parallel, and n pairs of such twin elements are used in conjunction with an n-phase goniometer. Improvement upon the usual Adcock system can be obtained in respect of both spacing error and sensitivity, but an increase in the number of antennas above 8 is of little practical advantage since the spacing for small errors approaches the absolute limit of 1.22 λ. A star connection of antennas provides a correct sense signal only when the antenna spacing is <0.76 λ.

621.396.93 1062
A Simple Method of Reducing the Polarization Error of a U-Type Adcock Direction-Finder—H. Fletcher. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 771–782; 1947.) Tests with an Army transportable direction finder which had a satisfactorily small polarization error on a site with high conductivity showed that, on a site of very low conductivity, a simple counterpoise of eight radial elements could have an effectiveness comparable with that of more elaborate earthing arrangements. On an average type of unfavorable site, a fourfold reduction in polarization error was obtained, resulting in a performance comparable with that for a very good site. The use of the system described is recommended for all but the best sites, with or without a layer of crushed coke, which is suggested as an alternative to the large circular earth mat often used.

621.396.93 1063
The Performance of High-Frequency Direction-Finders in Various Types of H.M. Ships—C. Crampton, W. Struszynski, J. H. Marshall, and J. C. Woolley. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 798–808; 1947.) A survey of the problems and errors associated with high-frequency direction-finding in ships, including: (a) choice of antenna system; (b) siting and rigging requirements; and (c) calibration and estimation of accuracy. Calibration results and performance curves for typical installations in different ships are analyzed, and possible methods of improving performance are discussed.

621.396.93 1064
The Development of a High-Frequency Cathode-Ray Direction-Finder for Naval Use—S. de Walden, A. F. L. Rocke, J. O. G. Barrett, and W. J. Pitts. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 823–837; 1947.) The operation of the latest design of shipborne equipment for the frequency range 1 to 24 Mc is based on the familiar twin-channel principle. A crossed-loop antenna system is used and sense is determined by causing the output from an omnidirectional antenna to black out one end of the cathode-ray-tube trace. Development problems associated with the balance of the twin amplifiers, the simplification of alignment and operation and the visual presentation of sense are discussed. The performance under operational conditions is described.

621.396.93 1065
Medium-Frequency Direction-Finding in H.M. Ships—G. J. Burtt:R. T. P. Whipple. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 838–856; 1947.) Part 1, by G. J. Burtt: A review of the present state of medium-frequency direction finding in both surface vessels and submarines and of recent advances in equipment. Part 2, by R. T. P. Whipple: A general discussion of the effect on the direction finders of reradiation from the ship's hull and deck structures.

621.396.93 1066
Direction-Finding—W. Ross:C. Crampton. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 867–870; 1947.) Discussion on 1955 and 1956 of 1948.

621.396.93:519.283 1067
Statistical Plotting Methods for Radio Direction-Finding—R. H. Barfield. (*Jour IEE* (London), part IIIA, vol. 94, no. 15, pp. 673–675; 1947.)

621.396.93:519.283 1068
The Estimation of the Probable Accuracy of High-Frequency Radio Direction-Finding Bearings—W. Ross. (*Jour IEE* (London), part IIIA, vol. 94, no. 15, pp. 722–726; 1947.) The method is based on the probable value of the variance associated with the observation. Snap bearings are taken over a period of five minutes. The arithmetic mean is taken as the observational bearing. The probable variance is estimated from (a) the spread of the observed readings, (b) site errors based on the past history of the direction finder, (c) the ionospheric lateral deviation based on known data, and (d) the observational error based on the flatness of the bearing. In a practical trial of the method, good agreement was obtained between the estimated and the actual bearing errors.

621.396.93:519.283 1069
Statistical Theory of D.F. Fixing—R. G. Stansfield. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 762–770; 1947.)

621.396.93:551.508.5 1070
Direction-Finding and the Measurement of Wind by Radio—D. N. Harrison. (*Met. Mag.*, vol. 76, pp. 217–225; October, 1947.) A general review of methods used by the British Meteorological Office for heights up to about 20 km. The radiosonde transmitter and adcock direction-finding sets are being replaced by reflectors and radar sets, which give a greater accuracy but, at present, a reduced range of observations when wind velocities are high. Future developments will be directed toward attaining heights up to at least 30 km, higher instrumental accuracy and better automatic operation.

621.396.93:551.594.6 1071
Sferics—Ockenden. (*See* 1056.)

621.396.93:621.314.2 1072
An Investigation of Symmetrical Screened Transformers for H.F. Radio Direction-Finders—Struszynski and Marshall. (*See* 969.)

621.396.93:621.317.324† 1073
Field-Strength Estimation by Means of High-Frequency Direction-Finders in H.M. Ships—Crampton and Toczylowski. (*See* 1118.)

621.396.93:621.396.611.1 1074
The Errors in Bearings of a High-Frequency Direction-Finder Caused by Reradiation from a Nearby Vertical Mast—C. Crampton, R. T. P. Whipple, and A. H. Mugridge. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 815–822; 1947.) The largest errors occur when the mast is in resonance at the frequency used.

621.396.93:621.396.676 1075
Some Principles Underlying the Design of Aerial Systems for High-Frequency Radio Direction-Finders in H.M. Ships—C. Crampton, W. Struszynski, S. de Walden, and P. G. Redgment. (*Jour. IEE* (London), part III, vol. 95, pp. 437–453; November, 1948.) The chief design problems arise from the existence of the secondary field from the mast on which the direction-finding antenna must be placed, and the necessity for long feeders (up to 150 feet) to the receiver. Fixed crossed-loop antennas of the single-turn, screened type are used, directly connected to twin screened feeders, which are coupled to the receiver by means of a transformer. The sense antenna consists of a vertical rod coaxial with the loops and mast, and a counterpoise system immediately below the loops. A test signal for the antenna system is provided from a small loop placed inside the direction-finding loops and at 45° to each. A high degree of equivalence of the direction-finding loops and symmetry of the whole structure is required; the magnitudes of the errors introduced by departures from the ideal conditions are investigated. The mechanism of antenna effect (nondirectional response) and steps taken to reduce it are described. A detailed account of the principles and practice of sense determination is given.

621.396.93:621.396.677 1076
A Mobile Spaced-Loop Direction-Finder—F. Caplin and J. H. Bagley. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 676–682; 1947.) A direction finder covering the frequency range 2 to 20 Mc and giving a silent arc of ±5° for a field strength of 8 microvolts per meter at 2 Mc to 2 microvolts per meter at 20 Mc. It can be used when high-angle sky waves predominate. Direction and sense finding in one operation by rotating the loops and watching a meter is achieved by using electronic switches to couple a resistance to each loop alternately. The equipment can be transported in a jeep trailer. See also 2780 of 1947 (Ross).

621.396.93:621.396.677 1077
Developments in H.F. Direction-Finder Shore Stations Using Adcock Aerials—J. F. Hatch. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 683–692; 1947.) Experiments on finding the direction of the separate ray components of multipath telegraphy signals showed that there was no difficulty in obtaining a bearing on the first component to arrive, but for the later signals, the sidebands had to be used instead of the carrier; this gave a reduced sensitivity. An antenna balancing unit which enable adjustments to be made with an internal oscillator instead of a portable transmitter is described, also a balanced potentiometer for measuring goniometer errors within 0.1° and an electrostatic screen for a goniometer. The performances of four types of direction finders are compared, and a system using an aural null method is discussed. In this system, the goniometer can also be rotated and the null observed on a cathode-ray tube.

621.396.93.029.62 1078
The Development of Single-Receiver Automatic Adcock Direction-Finders for Use in the Frequency Band 100–150 Mc/s—R. F. Cleaver. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 783–797; 1947.) The signals

from the two directional elements of the elevated H-adcock antenna system are modulated at different frequencies with suppression of the carrier, which is later restored in constant phase by the addition of a signal from a central omnidirectional antenna. The combined signal is passed through the receiver, whose output contains two af components. Their amplitudes and phases give the bearing and sense of the received signal, which are indicated by a cathode-ray tube display. Two experimental models, and a naval model based on these, are discussed. The probable instrumental error in the naval model is about 0.5 to 1.25° after allowing for octantal error. The required field strength is 7 microvolts per meter or less. Instruments for land use are considered; a direction finder capable of unattended operation on two frequency channels and having full remote control facilities is now being developed.

621.396.932 1079
Compact Marine Radar—(*Wireless World*, vol. 55, pp. 16–17; January, 1949.) Details are given of the latest Kelvin-Hughes equipment. A rotating cheese antenna is mounted directly over a case containing the transmitter, receiver, and power supply. Power supply is from a motor generator giving ac at 500 cps. The transmitter frequency is between 9,434 and 9,524 Mc; peak power is 30 kw, pulse duration 0.2 microsecond, and repetition frequency 1,000 per second. All supervisory controls are placed near the display unit, which has a 9-inch cathode-ray tube with magnetic deflector coils rotated by servo motors coupled to the antenna scanner drive. Range scales give maxima of 5, 9, and 27 miles

621.396.933.2 1080
Considerations in the Design of a Universal Beacon System—L. B. Hallman, Jr. (Proc. I.R.E., vol. 36, pp. 1526–1529; December. 1948.) Specifications are given for a beacon to be installed in aircraft for radar range extension, for transmission of air traffic control information (including range, azimuth, altitude, and identity data) and for automatic transmission of intelligence required in the operation of the system. The beacon must operate whatever the operating frequency of the primary radar. Display systems for identity interrogation are outlined. Suitable arrangements of component units are illustrated by block diagrams

621.396.933.23 1081
Indication of Landing Courses Independent of Weather Conditions—K. F. Niessen. (*Philips Res. Rep.*, vol. 3, pp. 130–139; April, 1948.) Continuation of 102 of February.

629.135.052:621.317.733 1082
Double-Ratio Bridges—(*See* 1126.)

621.396.93 1083
Radar Aids to Navigation [Book Review]—J. S. Hall (Ed.). McGraw-Hill, New York and London, 1947, 389 pp., 30s. (*Nature* (London), vol. 162, pp. 633–634; October 23, 1948.) Vol. 2 of the MIT Radiation Laboratory series. "... describes the advantages and limitations of radar technique when applied to the problems of navigation and pilotage, whether the equipment is airborne, shipborne, or ground-based."

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788 1084
On the Limits of Use of Thermal Manometers for the Measurement of Low Pressures—L. Dunoyer. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 1147–1149; November 29, 1948.) A symmetrical bridge device is described which is sensitive to pressure variations of the order of 3×10^{-6} mm Hg.

538.221:621.318.323.2 1085
Ferrite H.F. Magnet Cores—A. Weis. (*Funk und Ton*, vol. 2, pp. 564–578; November, 1948.) The influence of the conductivity of ferromagnetic materials on their usefulness in circuit components is considered and the properties of semiconductor metal compounds, particularly the ferrites consisting of a divalent metal oxide with Fe_2O_3, are discussed. The results of investigations by the Hescho company on ferrites containing various proportions of Mn_3O_4 and Fe_2O_3 show that such materials have decided advantages, the loss angle being only a fraction of that of iron-dust cores of comparable permeability and the density little more than half the dust-core value. Ferrites can also be moulded easily into forms such as long thin cylinders, whose high permeability and low loss can be used with advantage in high-frequency permeability tuning. Other applications are suggested and the Philips Mn/Zn and Ni/Zn ferrites, known as ferroxcube, are briefly mentioned.

546.28 1086
Temperature- and Humidity-Constant Organic Compounds of Silicon and their Suitability for Use in Electrotechnics—W. M. H. Schulze. (*Funk und Ton*, vol. 2, pp. 622–632; December, 1948.) Discussion of the electrical and mechanical properties of various materials, including silicones.

546.431.82:537.228.1:621.395.61/.62 1087
New Synthetic Piezoelectric Material—G. N. Howatt, J. W. Crownover, and A. Drametz. (*Electronics*, vol. 21, pp. 97–99; December, 1948.) Pure $BaTiO_3$ ceramic can acquire and retain induced piezoelectric properties if a dc polarizing field is applied. The properties and production of such material, and its use in a transducer, are discussed. See also 935 above.

546.431.82:537.228.2 1088
Electrostrictive Effect in Barium Titanate Ceramics—W. P. Mason. (*Phys. Rev.*, vol. 74, pp. 1134–1147; November 1, 1948.) When a dc bias is applied to a multicrystalline $BaTiO_3$ ceramic, an ac voltage can excite resonances in 4 different modes in the ceramic. The amount of motion is greater than in magnetostrictive materials, and $BaTiO_3$ may be an important electromechanical transducing element. All the modes can be explained on the assumption that when a given domain becomes ferroelectric, it loses its cubic structure and becomes tetragonal.

546.431.82:621.315.612 1089
Domain Structure of $BaTiO_3$ Crystals—H. Blattner, W. Känzig, W. Merz, and H. Sutter. (*Helv. Phys. Acta*, vol. 21, pp. 207–209; August 10, 1948. In German.) Summary of Swiss Phys. Soc. paper.

546.431.82:621.315.612 1090
Anomalies of the Specific Heat of $BaTiO_3$—H. Blattner and W. Merz. (*Helv. Phys. Acta*, vol. 21, pp. 210–222; August 10, 1948. In German.) Summary of Swiss Phys. Soc. paper.

546.431.82:621.315.612 1091
Electrical Conductivity and Refractive Index of $BaTiO_3$—G. Busch, H. Flury, and W. Merz. (*Helv. Phys. Acta*, vol. 21, pp. 212–215; August 10, 1948. In German.) Summary of Swiss Phys. Soc. paper.

546.431.82:621.315.612.011.5 1092
Dielectric Properties of Titanates at Ultra-High Frequencies—J. G. Powles. (*Nature* (London), vol. 162, p. 614; October 16, 1948.) Curves are given for $BaTiO_3$ for frequencies of 1.5 Mc and 9,450 Mc and temperatures from 20° to 170° C. Complex-permeability data for the same two frequencies and a temperature of 21° C are tabulated for the titanates of Mg, Ca, and Sr, and also for $BaTiO_3$ at 24,000 Mc.

546.431.82:621.315.612.011.5 1093
Dielectric Properties of Mixed Barium and Strontium Titanates at 10,000 Mc/s—J. G. Powles. (*Nature* (London), vol. 162, p. 655; October 23, 1948.) Complex permittivity as a function of the percentage of $BaTiO_3$ is shown graphically.

620.197 1094
Protective Finishing of Electrical Equipment—F. Widnall and R. Newbound. (*Jour. IEE* (London), part II, vol. 95, pp. 695–702; December, 1948.) Discussion on 1057 of 1948.

621.3.032.53:533.5:666.1.037.5 1095
Glass-to-Metal Sealing—In future the U.D.C. number 666.1.037.5 will be used for this subject, instead of 621.3.032.53:533.5 as formerly.

621.314.63 1096
Theory of Rectification of an Insulating Layer—H. Y. Fan. (*Phys. Rev.*, vol. 74, pp. 1505–1513; November 15, 1948.)

621.315.59:061.3 1097
Semiconductors and their Applications—W. Grattidge and F. A. Vick. (*Nature* (London), vol. 162, pp. 624–626; October 16, 1948.) Brief details of the papers read at a conference of the Manchester and District Branch of the Institute of Physics.

621.315.59:621.383 1098
Photoelectric Emission and Contact Potentials of Semiconductors—L. Apker, E. Taft, and J. Dickey. (*Phys. Rev.*, vol. 74, pp. 1462–1474; November 15, 1948.) In spherical photo cells with interchangeable emitters, energy distributions of external photoelectrons from the semiconductors Te, Ge, and B were compared with those of several metals.

621.315.61 1099
Insulating Materials for U.S.W. Technics—W. M. H. Schulze. (*Elektrotechnik* (Berlin), vol. 2, pp. 273–279; October, 1948.) Discussion, with numerous tables and diagrams, of the electrical characteristics of about 50 materials, including most of the new synthetic materials, at frequencies up to 10^{10} cps and for a few materials up to 10^{12} cps. The 38 references include 20 from German sources.

621.315.611.011.5 1100
Breakdown of Solid Insulating Materials—P. Perlick. (*Arch. Elek.* (Übertragung), vol. 2, pp. 174–185; April and May, 1948.) Wagner's theory of thermal breakdown is satisfactory for problems in the thermal range; the dependence of the breakdown voltage on the ambient temperature and on the frequency can be estimated quantitatively from the material constants. The fundamentals of the principal theories of breakdown are reviewed and compared with experimental results. Breakdown in solids is also compared with breakdown in air or in oil. Breakdown with dc occurs in nearly every case at a voltage equal to or below the ac peak voltage. The ac breakdown voltage recently reported for high-voltage cables, which were low in comparison with the dc breakdown voltages, are not in agreement with measurements on thin samples and can only be attributed to internal or external edge effects.

621.318.22 1101
New Magnetic Alloy—"Diallist." (*Wireless World*, vol. 55, p. 38; January, 1949.) A Ni-Al-Co-Fe alloy containing a minute percentage of Nb has higher coercivity than that of any known alloy. It has been developed by the Permanent-Magnet Association in collaboration with the Electrical Research Association.

621.383.4 1102
New Photoconductive Cells—E. Schwarz. (*Nature* (London), vol. 162, pp. 614–615; October 16, 1947.) Various methods in which an electric discharge is active can be used to produce such cells from a number of substances which form two groups. The first group (which includes Pt, Ni, Sb, and Ge) has the properties that if the carrier is kept at or below room temperature during the production, the film has a high resistance, a high negative temperature

coefficient of resistance (β) and has its optimum sensitivity in the near infrared. Heating the film in air or oxygen afterwards reduces the sensitivity, the resistance, and β. The substances in the second group (which includes the sulphides, selenides, and tellurides of Pb, Sn, In, Tl, Cd, Bi, and Sb) have similar properties before heating. Heat treatment in air or oxygen at a low temperature temporarily reduces the sensitivity, the resistance, and β, but prolonged heat treatment at a higher temperature increases these quantities again. Most of the tellurides show high sensitivity only at −78° C or at the temperature of liquid air. Some of the sulphides are very sensitive even at room temperature.

666.1.037.5:669.018.47:536.413.2 1103
The Effect of the Melting Point and the Volume Magnetostriction on the Thermal Expansion of Alloys—J. J. Went. (*Philips Tech. Rev.*, vol. 10, pp. 87–94; September, 1948.) The thermal expansion coefficient for an alloy can be determined in terms of the melting points and magnetostrictive properties of the constituent metals. This is used to determine, for example, an alloy of Fe, Ni, Co, and Cu which can be sealed to hard glass.

669.14–15:621.365.5 1104
Surface Treatment of Steel by Means of H.F. Induction Heating—Kegel. (*See* 1143.)

533.583:621.386 1105
Production and Use of Getter Materials in German Radio Valves, Thermionic Devices Generally, and Electric Lamps [Book Notice]—B.I.O.S. Final Report No. 1834, H.M. Stationery Office, London, 29 pp., 4s. General information is arranged according to the firm supplying it, but detailed information according to the type of volatile or nonvolatile getter used.

MATHEMATICS

51:621.396 1106
Mathematics and Radio Problems—B. van der Pol. (*Philips Res. Rep.*, vol. 3, pp. 174–190; June, 1948.) The author discusses the lack of mutual understanding between technicians, physicists, and mathematicians. Other topics dealt with are: the relation between Dirac's delta function and Stieltjes integrals; Hurwitz's determinants characterizing the stability of linear systems; wave equation; diffraction around a sphere (propagation of radio waves); continued fractions applied to filter circuits; nonlinear differential equations as related to tube oscillators; modern electrical calculating machines.

518.5 1107
A Digital Computer for Scientific Applications—C. F. West and J. E. DeTurk. (Proc. I.R.E., vol. 36, pp. 1452–1460; December, 1948.) The machine consists of a central control, an arithmetic unit, and two memory devices; the electronic techniques used are briefly described. The high-speed Hg-pool memory has a capacity of 4,080 words, each of 45 binary digits. A magnetic tape is used as a permanent storage medium and has a capacity of 200,000 words.

518.5 1108
On a Principle of Connexion for Bush Integrators—O. Amble. (*Jour. Sci. Instr.*, vol. 23, pp. 284–287; December, 1946.) A Bush integrator is a precision means of interconnecting three shafts so that their rotations u, v, w satisfy the equation $w=\int v du$. A regenerative connection is one for which the output rotation w contributes to the input rotation u. By means of such connections, a logarithm or square root can be generated with one integrator, and any rational power with two. See also 1109 below.

518.5 1109
Extensions in Differential Analyser Technique—J. G. L. Michel. (*Jour. Sci. Instr.*, vol. 25, pp. 357–361; October, 1948.) Analysis and extension of Amble's work (1108 above). Examples are also given of (a) the generation of the integral of a quotient, (b) the solution of differential equations in which the highest-order derivative has a variable coefficient, and (c) the inversion of functions defined by a differential equation.

518.5:621.385.832 1110
An Electronic Memory—(*Elec. Times*, vol. 114, p. 575; November 11, 1948.) Brief summary and discussion of IEE paper entitled "A Storage System for Use with Binary-Digital Computing Machines," by F. C. Williams and T. Kilburn. A great deal of information can be stored on the face of a cathode-ray tube in the form of electrical charges, using a scanning system similar to that of a television raster and interrupting the beam to give charged areas. The charges so deposited are detected by means of a metal-foil covering on the face of the tube.

MEASUREMENTS AND TEST GEAR

531.761:621.385.832:539.16.08 1111
A Cathode-Ray Tube Chronoscope—D. Pitman. (*Electronic Eng.* (London), vol. 20, pp. 384–389; December, 1948.) A crystal-controlled oscillator drives two frequency-divider units which both have "staircase" wave form. Each input pulse from the oscillator charges a capacitor until a predetermined potential is reached, when the capacitor is discharged and the cycle repeats. The input signal is made to create a phase difference between the waves in the two units, which is proportional to the duration of the signal and can be displayed on a cro. For direct reading, the oscillator frequency must be 10^n cps, where n is an integer, and the frequency-divider units must divide by 10. In the present design, $n=5$ and intervals between 10 microseconds and 0.1 second can be measured to within ±10 microseconds. By means of additional circuits, the system can be adapted for counting. Circuit and operation details are discussed and various applications are suggested.

531.764.5:621.396.615.18 1112
An Experimental Piezoelectric Chronometer Employing Regenerative Frequency Division—A. R. Jarvis, E. Cowcher, R. Keith, and J. A. Poll. (*AWA Tech. Rev.*, vol. 8, pp. 49–67; October, 1948.) Outputs of 100 kc, 10 kc, 1 kc, 250 cps, and 50 cps are available from a 100-kc crystal oscillator. The amplified 50-cps output will drive synchronous clocks with an accuracy better than ±1 second per week. The design of suitable frequency dividers is discussed.

531.764.5:621.396.615.18 1113
A Compact Piezoelectric Chronometer—J. E. Benson and E. M. Dash. (*AWA Tech. Rev.*, vol. 8, pp. 69–75; October, 1948.) A frequency generator for operating standard 50-cps clocks from a precision 100-kc quartz crystal. Two decade regenerative divider stages are used for frequency division to 1 kc, followed by two counter-type stages. The equipment is built on standard carrier panels and occupies about 2 feet, 8 inches of rack space.

621.3.018.4(083.74) 1114
A Microwave Secondary Frequency Standard—R. R. Unterberger and W. V. Smith. (*Rev. Sci. Instr.*, vol. 19, pp. 580–585; September, 1948.) A 10-Mc crystal oscillator, a FM klystron oscillator and crystal mixers are used to supply frequency markers at 90-Mc intervals from 2,970 Mc to above 40,000 Mc. Interpolation between markers is possible by means of a calibrated receiver. The equipment may be used for measuring spectra or for calibrating wavemeters. Absorption lines providing reference frequencies between 23,000 and 40,000 Mc are tabulated.

621.3.092 1115
Phase and Group Velocities and their Measurement—W. Deutschmann. (*Funk und Ton*, vol. 2, pp. 607–621; December, 1948.) Discussion of the application of these two concepts in the theory of the propagation of signals through networks. In certain cases neither these concepts nor the "frequency velocity" defined by Bürck and Lichte (*Elek. Nach. Tech.*, vol. 15, pp. 78–101; March, 1938) suffice to characterize the propagation completely. Methods of measurement are described and their advantages and disadvantages mentioned.

621.317.2 1116
Laboratory Antenna Distribution System—F. Mural. (*Proc. Radio Club Amer.*, vol. 25, no. 1, pp. 3–12; 1948.) A system designed to provide for 8 television channels. Seven are those assigned to the New York, N. Y., area and the 8th is available for a signal generated in the laboratory. Each of the signals is distributed to 10 test positions in the laboratory.

621.317.3:621.385.001.4 1117
Quality Control for Receiving Valves, and Industrial Applications—R. Suart. (*Radio Franç.*, pp. 6–10; October, 1948.) Routine methods are described for the measurement of all electrical constants, background noise, and microphony, and also for testing physical characteristics, in order to ensure a uniformly good quality in mass production.

621.317.324†:621.396.93 1118
Field-Strength Estimation by Means of High-Frequency Direction-Finders in H.M. Ships—C. Crampton and H. S. Toczylowski. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 809–814; 1947.) The measurement of the field strength of signals received on two types of high-frequency direction finder and the application of the results to the estimation of the range of transmitters of known characteristics are described. The accuracy of the method and its useful range are given, together with details of experimental checks.

621.317.33:621.315.59 1119
On the Determination of the Electromagnetic Constants of Semiconductors at U.H.F.—P. Jacottet. (*Helv. Phys. Acta*, vol. 21, pp. 251–260; August 10, 1948. In German.) Formulas are given for obtaining the dielectric constant, loss angle, and reflection factor from measurements with a cylindrical resonator. The method is similar to that of Féjer and Scherrer (1176 of 1943). Results are given for a buna type of material.

621.317.336.029.64 1120
A Swept-Frequency 3-Centimeter Impedance Indicator—H. J. Riblet. (Proc. I.R.E., vol. 36, pp. 1493–1499; December, 1948.) The magnitude and phase of an impedance over a 12 per cent variation of frequency can be deduced from the display produced on a cathode-ray tube by the equipment. Method, operation, and performance are discussed.

621.317.372 1121
"Q" Meters—H. G. M. Spratt. (*Wireless World*, vol. 55, pp. 7–10; January, 1949.) Description, with complete circuit details, of a Dawe Instruments meter, and of its use for measurements on capacitors, resistors, and transmission lines.

621.317.374:621.316.8 1122
A Method for Measuring the Loss Angle of Resistors—J. Heffels. (*Arch. Elek.* (Übertragung), vol. 2, pp. 78–83; February and March, 1948.) A bridge method making use of a standard resistor of low loss.

621.317.71:621.385.3 1123
Fluctuations in Electrometer Triode Circuits—A. van der Ziel. (*Physica*, *'s Grav.*, vol. 9, pp. 177–192; February, 1942. In English.) Only the Brownian movement in the triode input circuit and the shot-effect of the grid dc produce any marked effect on the accuracy of small-current measurement with electrometer tubes. The rms errors are calculated for three different methods of measurement. When the input resistance is of the order of $10^{14}\ \Omega$, the rms error is only about 2×10^{-18} amperes.

621.317.715:621.396.645.371 1124
Influence of Reactive Feedback Networks on the Response of Galvanometers—Savic. (*See* 1014.)

621.317.729 1125
Automatic Plotting of Electrostatic Fields—P. E. Green, Jr. (*Rev. Sci. Instr.*, vol. 19, pp. 646–653; October, 1948.) The usual electrolyte-tank method has been extended by adding a servomechanism which causes the probe to trace out an equipotential automatically.

621.317.733:629.135.052 1126
Double-Ratio Bridges—(*Elec. Times*, vol. 115, pp. 73–74; January 20, 1949.) Brief summaries of two IEE papers: "Double-Ratio A.C. Bridges with Inductively Coupled Ratio Arms," by H. A. M. Clark and P. B. Vanderlyn, and "Direct-Capacitance Aircraft Altimeter," by W. L. Watton and M. E. Pemberton. The bridge discussed in the first paper was used in the altimeter described in the second; the altimeter was not suitable for heights above 200 feet.

621.317.74:621.315.2 1127
Two Test Sets for the Maintenance of Carrier-Current Systems on Balanced-Pair or Coaxial Cables—P. Herreng. (*Câbles and Trans.* (Paris), vol. 2, pp. 305–318; October, 1948.) With English summary.) Apparatus for measuring transmission constants and absolute voltage levels over a wide range of frequencies, and also the nonlinear distortion of repeaters.

621.317.755 1128
A Frequency Characteristic Analyzer—J. W. Sampsell. (*Rev. Sci. Instr.*, vol. 19, pp. 620–627; October, 1948.) For analyzing the frequency characteristics of filters, amplifiers etc., whose pass bands are within the range 500 cps to 50 kc. Special features are: (a) the width of the swept band and the rate of sweep can be varied independently, (b) the frequency is swept linearly, (c) the response of the pass-band circuit can be made to appear at any point in the sweep, and (d) the cro timebase can be interlocked with the generator frequency.

621.317.761 1129
A Direct-Reading Frequency-Measuring Set—F. C. F. Phillips. (Proc. I.R.E. (Australia), vol. 9, pp. 12–19; January, 1948.) Reprint of 156 of February.

621.317.763.029.63:621.396.611.4 1130
The Design and Use of Resonant Cavity Wavemeters for Spectrum Measurements of Pulsed Transmitters at Wavelengths near 10 cm—H. R. Allan and C. D. Curling. (*Jour. IEE* (London), part III, vol. 95, pp. 473–484; November, 1948.) The factors influencing the design of cylindrical resonators for frequencies near 3,000 Mc are discussed. The H_{011} mode is used and it is shown how the dimensions for optimum Q and freedom from unwanted resonances are obtained. Tuning is effected by the use of a coaxial plunger whose diameter is about $\frac{1}{3}$ of that of the resonator; an almost linear frequency variation with plunger position is obtained over a considerable range. Transmitter spectra are checked by observing the changes in output of a diode detector, coupled to the cavity, as the tuning of the cavity is varied. Errors arising in the method of coupling, and in the behavior of the cavity itself, are examined.

621.317.763.029.65†:535.33.071 1131
A Grating Spectrometer for Millimeter Waves—R. J. Coates. (*Rev. Sci. Instr.*, vol. 19, pp. 586–590; September, 1948.) A paraboloid of revolution, with a double waveguide feed at the focus, is used for illuminating a reflecting echelette grating and receiving the reflected signal. The grating angles corresponding to the intensity peaks can be read within ± 0.04 per cent for λ 3 to 12.5 mm. Observed and theoretical intensity curves are compared.

621.317.79:621.396.611.1 1132
An Instrument for Measuring the Resonant Frequency of Reradiating Structures—W. Struszynski, E. G. Robus, and J. C. Woolley. (*Jour. IEE* (London), part IIIA, vol. 94, no. 15, pp. 741–750; 1947.) The instrument consists of an oscillator whose output coil is inductively coupled to the structure, and an amplifying detector with a screened-loop pickup coil for measuring the induced current, which shows marked increases at the resonant frequencies of the structure. The circuit theory is analyzed and various applications are illustrated.

621.397.62.001.4 1133
Television Crosshatch Generator—(*Electronics*, vol. 22, pp. 154, 158; January, 1949.) An instrument requiring little power, which can be connected directly to a television receiver to produce on the receiver screen a test pattern of intersecting vertical and horizontal lines. A circuit diagram is given.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

535.61-15 1134
Present State of Knowledge and Technical Applications of Infra-Red Radiation—K. Grosskurth. (*Fernmeldetech. Z.*, vol. 1, pp. 169–174; October, 1948.) A general review, with tables and graphs showing the properties of various types of photo cells. A description is included of the electron-optical image converter developed in Germany and used in many war-time applications.

539.16.08 1135
Reduction of Dead Times in Geiger-Müller Counters—B. Collinge. (*Nature* (London), vol. 162, pp. 853–854; November 27, 1948.)

539.16.08 1136
The Geiger Discharge—D. H. Wilkinson. (*Phys. Rev.*, vol. 74, pp. 1417–1429; November 15, 1948.) The mechanism of formation of the space-charge sheath is analyzed, and the theory is used to calculate (a) the relation between starting potential and counter variables, (b) the amount of charge generated, (c) the shape of the plateau curve, (d) the velocity of propagation along the wire. All forms of counter behavior are shown to depend strongly on the ratio of charge generated in the counter to that originally on the wire.

539.16.08 1137
The Mechanism of the Geiger-Müller Counter—A. Nawijn. (*Physica, 's Grav.*, vol. 9, pp. 481–493; May, 1942. In English.)

539.16.08 1138
On Some Fluctuation Problems Connected with the Counting of Impulses Produced by a Geiger-Müller Counter or Ionization Chamber—H. A. van der Velden and P. M. Endt. (*Physica, 's Grav.*, vol. 9, pp. 641–657; July, 1942. In English.)

539.16.08 1139
Measurements on Self-Quenching Geiger-Müller Counters—A. G. M. van Gemert, H. den Hartog, and F. A. Muller. (*Physica, 's Grav.*, vol. 9, pp. 556–564 and 658–664; June and July, 1942. In English.)

539.16.08 1140
Origin of the Temperature Effect in Alcohol-Argon-Filled Geiger-Müller Tubes—G. Joyet and M. Simon. (*Helv. Phys. Acta*, vol. 21, pp. 180–183; August 10, 1948. In French.) Summary of Swiss Phys. Soc. paper.

539.16.08:531.761:621.385.832 1141
A Cathode-Ray Tube Chronoscope—Pitman. (*See* 1111.)

621.316.726:615.84 1142
Frequency Stabilization of Diathermy Units—C. K. Gieringer. (*Electronics*, vol. 21, pp. 78–80; December, 1948.) Discussion of the problems involved in maintaining the frequency of diathermy units within FCC limits, and description of a plug-in monitor which stops the oscillator and sounds a buzzer when the frequency drift exceeds a predetermined amount.

621.365.5:669.14-15 1143
Surface Treatment of Steel by Means of H.F. Induction Heating—K. Kegel. (*Elektrotechnik* (Berlin), vol. 2, pp. 285–291; October, 1948.) The advantages of the method are enumerated and suitable equipment is described, including a 30-kw high-frequency generator; practical examples include the hardening of (a) parts of complex shape, and (b) the teeth of gear wheels.

621.38.001.8 1144
Design for a Brain—W. R. Ashby. (*Electronic Eng.* (London), vol. 20, pp. 379–383; December, 1948.) Discussion of the Homeostat, a machine incorporating negative feedback which is claimed to be capable of automatically seeking the optimum adjustment of its controls to meet any change in operating conditions.

621.38.001.8:061.3 1145
Electronics in Industry—(*Electrician*, vol. 141, pp. 1593–1594; November 26, 1948.) Brief details of papers read at the first Electronics Symposium organized by the Scientific Instrument Manufacturers' Association.

621.38.001.8:539.17 1146
Electronics in Nuclear Physics—W. E. Shoupp. (Proc. I.R.E., vol. 36, pp. 1518–1526; December, 1948.)

621.383:535.61-15 1147
The Image Converter—H. Mahl. (*Elektron Wiss. Tech.*, vol. 2, pp. 260–268; November, 1948.) Basic principles are outlined. Various early types are shown and diode and triode arrangements are described which use electron-optical immersion objectives. Applications are illustrated.

621.384.6 1148
A Travelling-Wave Linear Accelerator for 4-MeV Electrons—D. W. Fry, R. B. R. S. Harvie, L. B. Mullett, and W. Walkinshaw. (*Nature* (London), vol. 162, pp. 859–861; November 27, 1948.) Basic principles and a first model of such an accelerator were discussed in 506 of 1948. The length of the accelerator here described is 2 meters and the rf power is 2 mw.

621.384.6 1149
Air-Cored Synchrotron—T. R. Kaiser and J. L. Tuck. (*Nature* (London), vol. 162, pp. 616–618; October 16, 1948.) Discussion of a method of accelerating electrons to extreme relativistic energies which may be specially applicable to this type of synchrotron. See also 1350 and 1432 of 1948 (Blewett).

621.384.6 1150
Electrostatic Deflection of a Betatron or Synchrotron Beam—E. D. Courant and H. A. Bethe. (*Rev. Sci. Instr.*, vol. 19, pp. 632–637; October, 1948.)

621.384.6 1151
Research on the Electron Cyclotron—H. Salow. (*Funk und Ton*, vol. 2, pp. 531–538; October, 1948.) Description of equipment and discussion of results for electron paths in resonance and out of resonance. The effects of varying operational parameters are also considered.

621.385.15 1152
The Determination of the Pulse Period of Electron Multiplier Tubes—G. Papp. (*Rev. Sci. Instr.*, vol. 19, pp. 568–569; September, 1948.)

621.39:578.088.7 1153
An Ink-Writing Cardiochronograph for the Study of the Activity of the Human Autonomic Nervous System—W. W. Loucks, S. S. Kostashuk, and A. C. Burton. (*Canad. Jour. Res.*, sec. F, vol. 26, pp. 447–456; October, 1948.)

621.396.645.001.8:535.6 1154
Applications of Alternating Current Amplifiers to Optical Measurements—Harris. (*See* 1007.)

621.396.9:623.26 1155
Development and Use of Magnetic Apparatus for Bomb and Mine Location—A. Butterworth. (*Jour IEE* (London), part II, vol. 95, pp. 645–652; December, 1948. Discussion, pp. 664–667. Summary, *ibid.*, part I, vol. 95, p. 407; September, 1948.) Full paper; summary noted in 2882 of 1948.

621.396.9:623.26 1156
Development of Locators of Small Metallic Bodies Buried in the Ground—B. Roston. (*Jour. IEE* (London), part II, vol. 95, pp. 653–664; December, 1948. Discussion, pp. 664–667. Summary, *ibid.*, part I, vol. 95, p. 408; September, 1948.) Full paper: summary noted in 2882 of 1948.

PROPAGATION OF WAVES

538.566+621.396.812.029.64 1157
Phase Difference between the Fields of Two Vertically Spaced Antennas—E. W. Hamlin and A. W. Straiton. (PROC. I.R.E., vol. 36, pp. 1538 1543; December, 1948.) The phase difference is considered as a function of transmitter and receiver heights, and of range, for line-of-sight microwave propagation. The results obtained have helped to interpret experimental results such as those noted in 1182 of 1947 (Sharpless), 3225 of 1948 (Straiton and Gerhardt) and 493 of February (Hamlin and Gordon).

538.566:534.222.1:621.392.26† 1158
On the Theory of Spherically Symmetric Inhomogeneous Wave Guides, in Connection with Tropospheric Radio Propagation and Under-Water Acoustic Propagation—H. Bremmer. (*Philips Res. Rep.*, vol. 3, pp. 102–120; April, 1948.) The atmosphere is regarded as an inhomogeneous medium forming a curved waveguide through which radio and sound waves can be propagated. Such propagation is, in many respects, similar to that of underwater sound waves, but essential differences are pointed out. These differences are due to the fact that the product of distance from the center of the earth and index of refraction has at least one minimum for atmospheric propagation and one maximum for oceanic propagation. In consequence, the interval between the arrival of consecutive rays originating from a point source is, for atmospheric propagation, least for the rays coming latest, while for oceanic propagation it is least for those coming earliest. Arbitrary spherically symmetrical waveguides are discussed, and the concept of cutoff frequency is considered from the point of view of the modes as well as from that of geometrical optics.

538.566:551.5 1159
Some Problems in Radio Meteorology—H. G. Booker. (*Quart. Jour. R. Met. Soc.*, vol. 74, pp. 277–315; July to October, 1948.) Discussion of atmospheric refraction of radio waves. Results obtained at war-time radar stations and the general nature of superrefraction are considered; see also 516 of 1947. Propagation curves were given in 2892 of 1947 (Booker and Walkinshaw) for calculating the field strength due to a given transmitter under specified conditions of superrefraction; the problem here considered is the determination of these conditions of superrefraction from the ordinary data available in synoptic meteorology. The conjugate power-law theory of eddy-diffusion is tentatively used for study of the profiles of temperature and humidity involved in certain types of superrefraction; theoretical and observed profiles are compared. The limitations of the theory and possible methods of overcoming them are considered. See also 3487 of 1948 (Macfarlane).

621.396.11 1160
Work of Soviet Scientists in the Field of Propagation of Ultra-Short Radio Waves—B. A. Vvedenski. (*Bull. Acad. Sci.* (*URSS*, pp. 835–852; June, 1948. Bibliography, pp. 852 854. In Russian.) A brief survey of investigations carried out by Soviet scientists during the last 25 years at m, dm, and cm wavelengths.

621.396.11 1161
Range of Low-Power Radiocommunication—M. V. Callendar. (*Jour. IEE* (London), part III, vol. 95, pp. 425–435; November, 1948. Summary, *ibid.*, part I, vol. 95, p. 506; November, 1948.) Data available for calculating the ground-ray field strength are summarized and simplified field-strength formulas derived. Power losses at the transmitting antenna and the field strength required to overcome noise at the receiver are considered in typical cases. Maximum-range curves are deduced for a 1-watt transmitter and wavelengths from 0.5 to 2,000 meters with antennas at heights up to 15,000 feet.

621.396.11:535.312 1162
Reflections from Flat Sheet and Angle Reflectors—L. Lewin. (*Jour IEE* (London), part III, vol. 95, pp. 485 488; November, 1948.) "The reflection diagram from a flat sheet is the same as the transmission polar diagram of an aperture of double the dimensions of the sheet. A small deviation from flatness considerably alters the diagram, a sag of half a wavelength at the center producing a zero where there would otherwise have been a maximum. The right-angled corner reflector is also investigated, and the broad reflection diagram expected from geometrical optics is found. The effect of a small departure of the included angle from a right angle is also discussed."

621.396.11:551.510.535 1163
The Variation of the Height of the F-Layer, Cause of the Changes of the Frequency of Radio Waves during Propagation—K. Rawer. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 1149–1151; November 29, 1948.) Systematic measurements of the angle of arrival of the WWV transmissions show that the angle decreases as the frequency increases. The results correspond to a 5-hop path for the 5-Mc signals, with 4 and 2 hops respectively for the 10-Mc and 20-Mc signals. Calculations based on the daily variation of the height of the *F*-layer give a frequency variation, for a path with 5 reflections, of the order of 20×10^{-8}. This is in good agreement with the measurements of Decaux (1725 of 1948). The changes of *F*-layer height at sunrise and sunset correspond also to the sign of the observed frequency changes.

621.396.11:551.510.535 1164
Distribution of the Field Reflected by the Ionosphere in the Absence of Absorption—P. Lejay and D. Lepechinsky. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 997–1000; November 15, 1948.) Methods are described for determining graphically the virtual trajectories of propagation (a) for plane earth and ionosphere, and (b) for curved earth and ionosphere. These methods are applied to find the field at a point distant from a transmitter, neglecting absorption. With a perfectly reflecting earth there is a considerable augmentation of the field at a given point, due to reflected radiation. At points near the limit of the skip zone, considerable amplitude variations occur.

621.396.11:551.510.535 1165
On Attenuation Phenomena in the Ionosphere—B. Beckmann. (*Arch. Elek.* (Übertragung), vol. 2, pp. 124 135; April and May, 1948.) A theoretical discussion, with an account of the principal results of echo and reception measurements carried out during 1943 and 1944 at the German Post Office Research Establishment, München. During the day, "penetration attenuation" and *D*-layer attenuation are superposed, but after sunset only the former is found. Penetration attenuation varies approximately linearly with frequency. The variations of the maximum day-time attenuation are considerable; at the high and the low frequencies of the short-wave region, the behavior is fundamentally different, corresponding to the different origin of the attenuation at these frequencies. After sunset, when the *D*-layer attenuation disappears, the effects at the high and the low frequencies are similar. The seasonal variation of the attenuation is quite different for the high and the low frequencies. *D*-layer attenuation increases with the increased ionization from winter to summer and the penetration attenuation probably also increases in the summer. A decrease is, however, observed and is probably accounted for by a reduction of *F*-layer reflection due to an abnormal *E* layer. With transmission paths of the order of 1,000 km, for 20-m waves, only 25 per cent of a large number of field-strength measurements showed the so-called sunset effect. In the majority of cases, the field strength sank gradually below the sensitivity limit of the receiver. With oversea transmission paths, sunset effects are not observed.

621.396.812:551.510.535 1166
Periodic or Rhythmic Variation of the Intensity of Short Wave Radio Signals—S. S. Banerjee and R. N. Singh. (*Indian Jour. Phys.*, vol. 22, pp. 413–422; September, 1948.) Measurements of the angle of arrival of downcoming waves at sunrise and sunset show that periodic fading may be due to the reflection of waves from one or two ionospheric layers which have slow vertical movement and adequate electron density. Such movement is presumably caused by rapid changes of electron density during the transition periods of ionization. Slow periodic fading can also occur if the frequency used approaches the maximum usable frequency for the transmission path in question. In this case, the interference is caused by magneto-ionic components of reflected waves, as suggested by Appleton and Beynon (2895 of 1947). Observations of periodic fading of short-wave signals transmitted from Delhi at various seasons and times of day are also discussed.

621.396.812.029.64 1167
Low-Level Atmospheric Ducts—J. S. McPetrie and B. Starnecki. (*Nature* (London), vol. 162, p. 818; November 20, 1948.) Correlation of meteorological data with microwave field-strength measurements at the ends of a 60-mile oversea path (see 518 of 1947) suggests as expected (see 516 of 1947) that high field strengths, when the air is more than 5° warmer than the sea are due to ducts with a height of 80 to 120 feet, caused by a temperature inversion and/or a lapse in water vapor content near sea level. But when the air is more than 5° colder than the sea, ducts with a height of 20 to 50 feet are formed. This unexpected result may be due to the presence of sea spray whipped up by the wind.

RECEPTION

621.395.521.3:621.395.813 1168
Hysteresis Distortion of Pupin Coils—H. Mermoz and M. Troublé. (*Câbles and Trans.* (Paris), vol. 2, pp. 319–346; October, 1948. With English summary.)

621.396.621 1169
F.M. Receiver Design Problems—E. C. Freeland. (*Electronics*, vol. 22, pp. 104–110; January, 1949.) A survey of design and production techniques. The relative merits of limiter-discriminator, synchronized-oscillator, and ratio detectors are discussed. Hum reduction and the tracing and elimination of regenerative effects in if and rf stages are considered, particularly for ac/dc receivers.

621.396.621 1170
F.M. and P.M. Demodulator—J. A. Sargrove and R. E. Blaise. (*Electronics*, vol. 22, pp. 165, 171; January, 1949.) The newly developed phasitron circuit is described, in which a single multigrid tube is used as a demodulator. Any conventional tube having two control grids is

suitable. The incoming signal is applied to one grid; the other grid is connected to a high-Q circuit tuned to the mean carrier frequency. Because of the electron coupling between the two grids, oscillations are induced in the tuned circuit about 90° out of phase with the carrier oscillations. As the incoming signal deviates, because of its modulation, from the mean frequency, the oscillations in the tuned circuit vary in phase and thus the resultant plate current varies about its mean and the incoming FM signal is converted into an af current. Various industrial applications are suggested.

621.396.621:621.396.619.13 1171
A Phase Discriminator for Frequency-Modulation Reception—Newall and Spencer. (*See* 1001.)

621.396.621:621.396.9 1172
An Anti-Clutter Radar Receiver—R. V. Alred and A. Reiss. (*Jour. IEE* (London), part III, vol. 95, pp. 459–465; November, 1948.) Summary, *ibid.*, part I, vol. 95, pp. 507–508; November, 1948.) The loss of target echoes due to receiver saturation by responses from neighboring land, rain storms or sea, can be much reduced by the use of a receiver in which the output is proportional to the logarithm of the input. The design and performance of such a receiver are described and other applications mentioned.

621.396.621.54 1173
The Tracking of Superheterodyne Receivers—K. J. Coppin. (*Jour. Brit. I.R.E.*, vol. 8, pp. 265–284; November and December, 1948.) The "2-point" method is shown to be generally unacceptable even when the two frequencies with zero tracking error are chosen as well as possible. Limitations of existing formulas for calculating oscillator circuit parameters for the commercially established "3-point" method are noted. The error form of this method is analyzed and it is shown that, in general, within a tuning range, exact alignment may be obtained at three frequencies which are uniquely determined by the circuit constants. An expression is derived for calculating the tracking error at frequencies between these points. The three frequencies should be chosen so that, throughout the tuning range, maximum percentage tracking errors are numerically equal rather than maximum absolute errors. General equations for calculating oscillator circuit parameters are derived and a numerical example is included. Errors of the method are indicated.

621.396.621.54.029.6 1174
Developments in Radio-Receiver Circuits for the Ultra-Short-Wave Range—A. van Weel (*Philips Res. Rep.*, vol. 3, pp. 191–212; June, 1948.) Frequency converters are described in which the signal voltage is applied in push-pull whereas the local-oscillator voltage is applied in parallel. The input circuit is tuned to both frequencies at the same time. Methods for blocking the local-oscillator power from the antenna are discussed. Designs for diode, triode, and self-oscillating mixers are considered in detail.

621.396.622:621.396.8 1175
Rectification of a Sinusoidally Modulated Carrier in the Presence of Noise—D. Middleton. (Proc. I.R.E., vol. 36, pp. 1467–1477; December, 1948.) The low-frequency output, signal and noise, is determined for an nth-power-law half-wave rectifier and for modulation up to 100 per cent. Special attention is given to linear and square-law detectors, for which audio signal-to-noise ratios are calculated. The noise passed by the audio filter depends on the spectral shape of the if; three types of filter are considered. The degree of modulation has little effect on output noise power if the mean input signal-to-noise power ratio $p>1$; for $p\rightarrow\infty$ the audio signal-to-noise ratio is proportional to the input carrier amplitude, is independent of the if filter bandwidth and is only slightly dependent on filter shape.

621.396.8 1176
Signal-to-Noise Ratio in A.M. Receivers—E. G. Fubini and D. C. Johnson. (Proc. I.R.E., vol. 36, pp. 1461–1466; December, 1948.) Experiments confirm the theory of the effect of a linear detector on the signal and signal-to-noise ratio obtained by demodulation of an rf carrier. When a carrier is present, the output noise of a detector increases by 4 to 7 db, according to the shape of the if filter. For sine-wave AM and if bandwidths at least 3 or 4 times larger than the af bandwidth, a universal curve can be given that shows the relation between the signal-to-noise ratio at the output and the carrier-to-noise ratio at the input of a second detector. If two AM carriers are simultaneously present at the(input of a linear second detector, this discriminates against the modulation of the weaker carrier.

621.396.813:621.396.619.13 1177
The Necessary Bandwidth in Reception of F.M. Signals for Eliminating Nonlinear Distortion—E. I. Manaev. (*Radiotekhnika* (Moscow), vol. 3, pp. 54–61; September and October, 1948. In Russian.) Approximate formulas (6) and (11) "sufficiently accurate for practical purposes" are derived for determining the bandwidth of FM oscillations in the cases of sinusoidal and square-wave signals respectively. The components of the FM oscillations are assumed to be not smaller than 1 per cent of the unmodulated carrier. The nonlinear distortion occurring in the tuned circuits of a receiver when the FM oscillations are being received is discussed. A formula (35) is derived for determining the coefficient of nonlinearity k_f which is equal to the ratio of the amplitude of the third harmonic of the frequency deviation to that of the first harmonic. Formula (30) is deduced for determining the necessary bandwidth of high-frequency and if amplifiers for a given k_f. Two numerical examples are included.

621.396.822:523.746 1178
Changes in Ionization and Radio Reception during the Sunspot-Period 1944–1947—Stetson. (*See* 1035.)

631.397.828 1179
The Reduction of Interference in Television Sound Reception—H. Fairhurst. (*Jour. Telev. Soc.*, vol. 5, pp. 126–131; December, 1947.) The effect on the sound channel is more serious than that on the picture. At 45 Mc the main sources of interference are the inition system of motor cars. The relative merits of different types of noise limiter are discussed. A series-diode-following circuit used in Murphy receivers is described, with full explanation of the reasons for choosing various component values.

STATIONS AND COMMUNICATION SYSTEMS

621.394.441 1180
Carrier-Frequency-Shift Telegraphy—R. Ruddlesden, E. Forster and Z. Jelonek. (*Jour. IEE* (London), part III, vol. 95, pp. 454–458; November, 1948). Discussion on 2354 of 1948.

621.396.1 1181
Geographical Distribution of the Frequencies Allocated by the Copenhagen Plan—(*Radio Tech. Dig.* (Franç.), vol. 2, pp. 303–311; December, 1948.) See also 832 of April.

621.396.3 1182
Some Developments in Commercial Point-to-Point Radiotelegraphy—J. A. Smale. (*Jour. IEE* (London), part III, vol. 95, pp. 454–458; November, 1948.) Discussion on 2358 of 1948.

621.396.5 1183
Modern Single-Sideband Equipment—C. T. F. van der Wyck. (Proc. I.R.E., vol. 36, p. 1505; December, 1948.) Comment on 3510 of 1948. The oscillator there discussed is included in United States patent No. 2,321,354 of 1943.

621.396.619.16 1184
Pulse Communication—D. Cooke; Z. Jelonek and E. Fitch; A. J. Oxford. (*Jour. IEE* (London), part III, vol. 95, pp. 465–466; November, 1948.) Discussion on 2079 of 1948.

621.396.65.029.64:621.397.743 1185
6,000-Mc Television Relay System—W. H. Forster. (*Electronics*, vol. 22, pp. 80–85; January, 1949.) A 2-way system operating between New York and Philadelphia. Repeaters and terminal equipment are described. See also 2921 of 1948.

621.396.712 1186
Planning the New KOMO Studios—F. J. Brott and S. Bennett. (*Broadcast News*, pp. 8–21; October, 1948.) An illustrated description of the general arrangement of studios, offices, etc. For further details see 1187 to 1189 and 1222 below.

621.396.712 1187
Constructing the KOMO Studios—S. Bennett. (*Broadcast News*, pp. 22–27; October, 1948.) Discussion of sound isolation, studio design and performance, reverberation measurements, etc. See also 1186 above.

621.396.712 1188
Equipment for the New KOMO Studios—M. E. Gunn. (*Broadcast News*, pp. 32–44; October, 1948.) An illustrated description, with block diagrams, of the master control room system, studio consoles, etc. See also 1186 above.

621.396.712:697 1189
Heating and Ventilating the New KOMO Studios—J. K. Gannett. (*Broadcast News*, pp. 28–31; October, 1948.) See also 1186 above

621.396.712 (44) 1190
The Allouis OCII Transmitting Centre for Broadcasting on Decametre Waves—M. Matricon. (*Rev. Tech. Comp.* (Franç.), pp. 5–15; September, 1948. In French, with English summary.) For other accounts see 553 and 2922 of 1948.

621.396.931 1191
Some Australian Developments in F.M. Mobile Communication Equipments—H. A. Ross: A. J. Campbell. (*AWA Tech. Rev.*, vol. 8, pp. 1–48; October, 1948.) Part 1, by Ross, discusses the design of experimental equipment working on frequencies of 43.2, 75.8, and 160.4 Mc, with special reference to discrimination against interference and fidelity of reproduction. Results of field tests on the 160.4-Mc equipment are given and a 3-dimensional contour map of the service area is discussed.

Part 2, by Campbell, outlines general considerations for the design of commercial equipment and some standard units are described and illustrated.

621.396.1 1192
International Radio Regulations [Book Notice]—H.M. Stationery Office, London, 1947, 336 pp., 3s. 6d. (*Govt. Publ.* (London), p. 13; November, 1948.) Contains the appendices and additional radio regulations annexed to the International Telecommunicatoin Convention, Atlantic City, 1947. This edition is a photostat copy of the English text published by the Bureau of the International Telecommunications Union, Berne, omitting the facsimile signatures of the original.

SUBSIDIARY APPARATUS

621-526+621.316.7 1193
Servomechanisms and Regulators. Stability Criteria. Application Examples—C. Galmiche. (*Rev. Gén. Élec.*, vol. 58, pp. 19–30; January, 1949.) General stability criteria are applied in a detailed discussion of (a) speed regulation of dc motors by means of a Leonard combination with "rototrol" excitation, (b) voltage regulation by electronic methods. The analogy between servomechanisms and electrical feedback devices is discussed in a short appendix.

621-526:621.3 1194
The Contribution of Electricity to the Technique of Servomechanisms—G. Lehmann. (*Bull. Soc. Franç. Élec.*, vol. 8, pp. 496–500; October, 1948.)

621.3.027.3:539.16.08 1195
High-Voltage Supplies for G-M Counters—A. Thomas. (*Electronics*, vol. 21, pp. 100–103; December, 1948.)

621.316.722.4:621.396.645 1196
Low-Impedance Variable Voltage Tappings—Scroggie. (*See* 1004.)

621.316.99 1197
The Design of Earthing Devices in Communication Equipment—E. A. Alekhin. (*Vestnik Svyazi*, no. 11, pp. 12–13; 1948. In Russian.) Nomograms are given for earthing devices using pipes, wires, or plates.

621.385.832:535.247 1198
A Recording Photometer and its Use in Studies of Cathode-Ray Screen Displays—F. Hamburger, Jr., and E. J. King. (*Jour Opt. Soc. Amer.*, vol. 38, pp. 875–879; October, 1948.) A 931-A photo-multiplier tube is combined with suitable optical and electronic auxiliaries in an instrument sensitive to either transient or slow brightness changes at any selected point of a cathode-ray screen.

621.396.68 1199
Very High Voltage Supply without H.V. Transformer—M. Alixant. (*Radio Tech. Dig.* (Franç.), vol. 2, pp. 239–249; October, 1948.) A review of high-frequency oscillator and pulse methods suitable for use in television receivers, and a short description of some practical circuits.

621.396.68:621.316.722 1200
Stabilized Power Supplies: Part 3—Extension of Output Voltage Range—M. G. Scroggie. (*Wireless World*, vol. 54, pp. 453–456; December, 1948.) A detailed circuit diagram is given and discussed, showing how output voltage can be varied between 0 and 500 volts. See also 1004 above.

771.3:621.317.755 1201
A New, Versatile Camera for the Cathode-Ray Oscillograph—H. P. Mansberg. (*Oscillographer*, vol. 10, pp. 2–14; October to December, 1948.) General description of the Du Mont Type 314 camera.

TELEVISION AND PHOTOTELEGRAPHY

621.397.2 1202
Electronics in the Service of the Press. The Transmission of Pictures to a Distance—E. Belin. (*Radio Tech. Dig.* (*Franç.*), vol. 2, pp. 209–223; October, 1948.) Discussion of methods suitable for (a) telephony circuits, and (b) radio links, with examples of the results obtained.

621.397.26 1203
Ultrafax—(*Electronics*, vol. 22, pp. 77–79; January, 1949.) The material to be transmitted is photographed on 35-mm film; high-speed scanning of the image is achieved by using as light source a cathode-ray tube with a special low-persistence phosphor screen. By means of a multiplier photo cell modulator, the message is made to modulate a 7,000-Mc carrier wave. Normal television technique is used for reception and the image on the cathode-ray tube is photographed on 16-mm film which can be developed in a few seconds. A transmission speed of about half a million words per minute was achieved at a demonstration of the equipment.

621.397.3 1204
The 'Reverse' Method for Designing the Output Stage of a Scanning System with a Magnetic Field—K. V. Saprykin. (*Radiotekhnika* (Moscow), vol. 3, pp. 34–46; July and August, 1948. In Russian.) The shape of the output current curve is usually determined from the given input voltage curve and the parameters of the circuit. Here the value and shape of the input voltage is determined, so that a sawtooth output current is obtained with given parameters of the circuit. The method is discussed in detail and a general equation (3) determining the input voltage is derived. The equation is then simplified by omitting the distributed capacitance of the coils in the circuit. The necessary jump in the input voltage and its total swing are determined by the scanning frequency, the duration of the flyback, and the time constants of the coils.

621.397.335 1205
Frame Synchronising Signal Separators—A. W. Keen. (*Electronic Eng.* (London), vol. 21, pp. 3–9; January, 1949.) Shortened version of 2386 of 1948.

621.397.335 1206
Locked Oscillator for Television Synchronization—K. Schlesinger. (*Electronics*, vol. 22, pp. 112–117; January, 1949.) 1948 National Electronics Conference paper. Flywheel circuits are discussed as a simple alternative to automatic phase control.

621.397.5 1207
The Part Played by Russian Scientists in the Development of Television—G. I. Golovin. (*Priroda*, pp. 73–80; August, 1948. In Russian.)

621.397.5 1208
Television Waveforms: Some Comparisons between British and American Standards—(*Wireless World*, vol. 54, pp. 439–440; December, 1948.) Discussion with special reference to the sense of modulation and the inclusion of equalizing pulses in the synchronizing signals.

621.397.5:629.135 1209
Airliner Television—(*Electronics*, vol. 21, p. 158; December, 1948.) Good reception from stations up to 180 miles away can be obtained with a commercial receiver. Alternative dipole antennas are provided.

621.397.5(083.74) 1210
Comparison of Television Standards in Germany, England and America—E. Schwartz. (*Arch. Elek.* (Übertragung), vol. 2, pp. 88–101; February and March, 1948.)

621.397.5(73) 1211
Impressions of American Television—T. M. C. Lance. (*Jour. Telev. Soc.*, vol. 5, pp. 132–138; December, 1947.) Discussion, p. 139. Report on the author's visit to the United States during March and April, 1947.

621.397.6:778.55 1212
Shutterless Television Film Projector—L. C. Downes and J. F. Wiggin. (*Electronics*, vol. 22, pp. 96–100; January, 1949.)

621.397.62 1213
Television Receiver Design, Engineering, Manufacture—(*Tele-Tech*, vol. 7, pp. 61–94; November, 1948.) A general survey of Philco methods.

621.397.62 1214
Televisor—(*Electronic Eng.* (London), vol. 21, p. 27; January, 1949.) Amendments to the booklet mentioned in 577 of 1948 (Flach and Bentley), with some additional notes.

621.397.62:535.88 1215
Projection-Television Receiver: Part 1—The Optical System for the Projection—P. M. van Alphen and H. Rinia. (*Philips Tech. Rev.*, vol. 10, pp. 69–78; September, 1948.) For another account see 2387 of 1948.

621.397.62:621.398 1216
Television Remote Viewers—V. Zeluff. (*Electronics*, vol. 21, pp. 90–93; December, 1948.) Video and audio signals from a receiver may be fed to a remote viewing unit which contains only power, audio, and scanning circuits. Circuits, based upon United States television standards, are suggested for operating electrostatic or magnetic cathode-ray tubes.

621.397.743:621.396.65.029.64 1217
6,000-Mc Television Relay System—Forster. (*See* 1185.)

621.397.5 1218
Principes Fondamentaux de Télévision 'Book Review'—H. Delaby. Eyrolles, Paris, 200 pp. (*Radio Tech. Dig.* (Franç.), vol. 2, p. 285; December, 1948.) Part of the Collection du Centre d'Enseignement de la Radiodiffusion Française. A further volume devoted to basic techniques of broadcasting is in preparation. "An excellent course of instruction in the principles of television."

TRANSMISSION

621.396.61 1219
The Design of a 500 Watt M.F. General Purpose Transmitter—W. J. Morcom. (*Marconi Rev.*, vol. 11, pp. 112–123; October to December, 1948.) A transmitter corresponding to that noted in 553 of March (Cooper), but covering the frequency range 275 to 550 kc.

621.396.61 1220
An Experimental Frequency-Modulated Broadcast Transmitter—J. B. Rudd, W. W. Honnor, and W. S. McGuire. (*Jour. IEE* (Australia), vol. 20, pp. 107–117; September, 1948). Description of a 250-watt transmitter for the range 88 to 108 Mc, including the modulator unit, power amplifier unit, antenna array, and station monitor. The operation of the modulator depends upon the time delay when a FM signal is passed through a multisection band-pass filter. A signal from a fixed-frequency source is mixed with a FM signal from a modulated oscillator; the FM difference-frequency component is passed through a time-delay network and mixed with the original FM signal. The difference-frequency component of the output of this second mixing has the same center frequency as the fixed source and is modulated in phase. Pure FM is obtained if the audio input is passed through an inverse-frequency network before application to the modulator.

621.396.61:621.316.726 1221
On the Stabilization of the Mean Frequency of F.M. U.S.W. Transmitters—F. Kirschstein and D. Weber. (*Funk und Ton*, vol. 2, pp. 499–515; October, 1948.) Discussion with special reference to the Crosby circuit (see 3294 of 1941) the Seeley discriminator (see 3802 of 1940) and the push-pull frequency modulator.

621.396.61:621.396.712 1222
KOMO's New 50,000 Watt Transmitter—F. J. Brott and C. E. Miller. (*Broadcast News*, pp. 45–50; October, 1948.) A general illustrated description. The end-fire antenna system consists of 3 antennas approximately 0.3 λ apart, mounted on 500-foot masts. See also 1186 above.

621.396.61:621.398 1223
The Keying of Radio Transmitters from Great Distances—S. E. Kuteynikov. (*Vestnik Svyazi*, no. 9, pp. 21–22; 1948. In Russian.) Discussion of keying from a distance of several hundred kilometers by using the SMT-34 carrier system.

621.396.619.23 1224
Some Aspects of the Design of Balanced Rectifier Modulators for Precision Applications—D. G. Tucker. (*Jour. IEE* (London), part III, vol. 95, p. 436; November, 1948.) Discussion on 3542 of 1948.

VACUUM TUBES AND THERMIONICS

621.383 1225
On the Residual Current in Photoelectric Receivers of Very High Sensitivity—M. Duchesne. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 1155–1157; November 29, 1948.)

621.383.4 1226
New Photoconductive Cells—Schwarz. (*See* 1102.)

621.385 1227
Miniature Battery Valves—O. P. Herrnkind. (*Funk-Technik* (Berlin), vol. 3, pp. 452–453; September, 1948.) Complete electrical data for the Tungsram high-frequency/medium-frequency pentode 1T4T, pentagrid frequency changer 1R4T, diode-low-frequency pentode 1S5T, and output pentodes 1S4T and 3S4T. Base connections and the circuit diagram of a receiver using these tubes are also given.

621.385 1228
Quadrature Operation of Filamentary Gas Tubes—V. L. Holdsway. (*Bell Sys. Tech. Publ. Monogr. B*-1552, 4 pp.) Reprinted from *Trans. AIEE*, vol. 67; 1948.) With plate and filament voltages in phase and the currents of comparable value, the nonuniformity of filament temperatures may result in short tube life. With the voltages in quadrature, the plate current may be increased to approximately double that permissible with in-phase operation.

621.385:621.396.822:621.396.645 1229
On the Influence of the Noise of Vacuum Tubes on the Accuracy of Linear Amplifiers—J. M. W. Milatz and K. J. Keller. (*Physica*, 's *Grav.*, vol. 9, pp. 97–112; January, 1942. In English.) The noise of an RCA57 tube with floating grid is measured as a function of frequency. The source of this noise, and its effect the precision of α-particle ionization measurement with this tube, are discussed.

621.385.001.4:621.317.3 1230
Quality Control for Receiving Valves, and Industrial Applications—Suart. (*See* 1117.)

621.385.032.21 1231
Cathodes with High Emission—R. A. Oñativia. (*Rev. Telegr.* (Buenos Aires), vol. 37, pp. 635–638; September, 1948.) Discussion of the preparation, activation, and characteristics of cathodes using (a) oxides of the alkaline earths (b) ThO₂, and (c) thoriated tungsten.

621.385.032.29 1232
The Calculation of Electrode Temperatures in the Radio Valve—I. A. Harris. (*Jour. Brit. I.R.E.*, vol. 8, pp. 288–312; November and December, 1948. Discussion, pp. 312–315.) The theory of radiation equilibrium of typical electrode systems is developed and, in conjunction with the theory of conduction, the results are applied to the calculation of the temperature of each electrode. An example is worked out for the case of an output tetrode, which gives results in close agreement with typical measured temperature values for such a tube.

621.385.1 1233
The Consequences of an Electron-Inertia Effect in Valves- Part 2—M. J. O. Strutt and A. van der Ziel. (*Physica*, 's *Grav.*, vol. 9, pp. 65–83; January, 1942. In German, with English summary.) Electrons, moving near electrodes, induce variable charges on them and hence currents in their leads. These currents are calculated for cases in which the electrodes are arranged so as to obtain maximum currents at very high frequencies impressed on the electron stream. The application of such electrode systems in vht amplifying tubes to obtain high ac output is discussed; comparison is made with known arrangements giving lower output currents. By making use of electron oscillations about an electrode, the output currents can be further increased. The considerable input loss at high frequency can be compensated or even changed to a negative resistance by means similar to those used for increasing the output. Numerous curves illustrate the results of the discussion. Part 1, 254 of February.

621.385.1 1234
European Receiving Tubes—H. A. S. Gibas. (*Electronics*, vol. 22, pp. 156, 162; February, 1949.) The system of numbering generally adopted for different types of tube is explained and the special features of certain tubes, including the Telefunken double tetrode VEL 11, are described.

621.385.1:621.316.726.078 1235
Tube Engineering News, the Transitrol—J. Kurshan. (*Communications*, vol. 29, pp. 15, 33; January, 1949.) Summary of the IRE-RMA Rochester Fall meeting paper. The "Transitrol" is a tube in which transit time effects, instead of being reduced to a minimum, are used to control the frequency of an oscillator automatically. Some electrons leave the cathode and go directly to the plate; their transit angles are small. These electrons contribute only to the transconductance of the tube. Others reach the plate indirectly after reflection and have large transit angles; these electrons contribute to the transsusceptance as well as the transconductance. By varying the reflector potential, the transsusceptance of the tube can be altered, and hence the frequency of the oscillations. The tube can also be used to generate a FM signal directly, by applying modulation to the reflector.

621.385.3 1236
Electrometer Valves—B. M. Tsarev. (*Uspekhi Fiz. Nauk.*, no. 2, pp. 251–270; 1948. In Russian.) A detailed survey giving data on tubes manufactured in Russia and elsewhere.

621.385.3/.4:537.291 1237
On the Phenomena in the Cathode–Grid Space of Triode and Tetrode Oscillators at Ultra High Frequencies—M. S. Neiman. (*Radiotekhnika* (Moscow), vol. 3, pp. 7–25; July and August, 1948. In Russian.) Discussion of: (a) deviation of the emission current from the "$\frac{3}{2}$ power" voltage versus current relationship, (b) the passage through the control grid of only those electrons which have left the cathode at the beginning of the emission interval, (c) the return of emitted electrons to the cathode, (d) the large ratio of the time interval during which the electrons pass through the grid to the interval during which they are emitted from the cathode, (e) variations in electron velocities after passage through the grid, and (f) acceleration of electrons by the grid. The simplifying assumption is made that all electrostatic lines of force from the space charge are directed toward the cathode; a number of simple relationships are derived but these should be regarded only as approximate. Other phenomena due to the inertia of electrons in oscillators with flat electrodes are mentioned.

621.385.3:621.317.71 1238
Fluctuations in Electrometer Triode Circuits—van der Ziel. (*See* 1123.)

621.385.38 1239
Recent Trends in the Construction of Thyratrons—J. Bell. (*Bull. Soc. Franç. Élec.*, vol. 8, pp. 489–495; October, 1948.) Discussion, mainly from the physical point of view, of thyratrons with Hg vapor, Ar, or X as gas filling. The characteristics for these types are compared.

621.385.5:621.396.822 1240
Measurements of Noise Factors of Pentode at 7.25-m Wavelength—A. van der Ziel and A. Versnel. (*Philips Res. Rep.*, vol. 3, pp. 121–129; April, 1948.)

621.385.831:621.318.25 1241
Residual Hum in Valves—W. I. Heath: L. H. Light: W. Grey Walter, H. W. Shipton, and W. J. Warren. (*Electronic Eng.* (London), vol. 20, p. 406; December, 1948.) Comment on 3559 of 1948 and the authors' reply. Heath suggests that the effects reported may be due to magnetic fields produced by a mains transformer or a mains-operated heater. Light considers that demagnetization cannot reduce the magnetic hum in the EF37 unless the tubes have previously been accidentally magnetized. Grey Walter, Shipton, and Warren note that the difference between pentode and triode connection may be significant, but have observed much higher hum levels than Light.

621.385.832 1242
Three-Dimensional Representation on Cathode-Ray Tubes—C. Berkeley. (Proc. I.R.E., vol. 36, pp. 1530–1535; December, 1948.) The procedure described may be applied to any regular system of co-ordinates or any number of variables. The representation may take the form of an oblique perspective picture. The procedure consists in: (a) setting the form of the representation, (b) deriving the position of the spot on the cathode-ray tube as a function of its true position in space, and (c) making the indicated corrections electrically. Applications in various fields are discussed. See also 577 of March (Parker and Wallis).

621.385.832 1243
The Negative-Ion Blemish in a Cathode-Ray Tube and Its Elimination—R. M. Bowie. (Proc. I.R.E., vol. 36, pp. 1482–1486; December, 1948.) Removal of initiating substances, provision of a thin metallic backing layer for the screen, and particularly the use of ion traps, are suggested remedies.

621.396.615.142.2 1244
Multifrequency Bunching in Reflex Klystrons—G. Hok. (Proc. I.R.E., vol. 36, p. 1505; December, 1948.) Comment on 2988 of 1948 (Huggins).

621.396.615.142.2 1245
Elementary Theory of the Klystron—A. V. J. Martin. (*Radio Tech. Dig.* (*Franç.*), vol. 2, pp. 199–207; October, 1948.)

621.396.619.231:621.396.645.37 1246
Valve Characteristic Giving Linear Modulation when a Feedback Resistor is Inserted in the Cathode Lead—W. W. Boelens. (*Philips Res. Rep.*, vol. 3, pp. 227–234; June, 1948.) The linearity of grid modulation is improved by the use of a feedback resistor in the cathode lead. Suitable plate-current versus grid-voltage characteristics for given feedback resistances are derived theoretically.

621.396.822 1247
On the Theory of Electrical Fluctuations—D. K. C. MacDonald. (*Proc. Roy. Soc. A*, vol. 195, pp. 225–230; December 7, 1948.) Comment on 2419 of 1948 (Fürth). MacDonald regards Nyquist's own derivation of his theorem as valid rather than Fürth's suggested alternative. Fürth's method of applying Nyquist's theorem directly to a thermionic tube to obtain the current fluctuations is also criticized. See also 2418 of 1947 (Fürth) and 953 of 1947.

533.583:621.385 1248
Production and Use of Getter Materials in German Radio Valves, Thermionic Devices generally, and Electric Lamps [Book Notice]—(*See* 1105.)

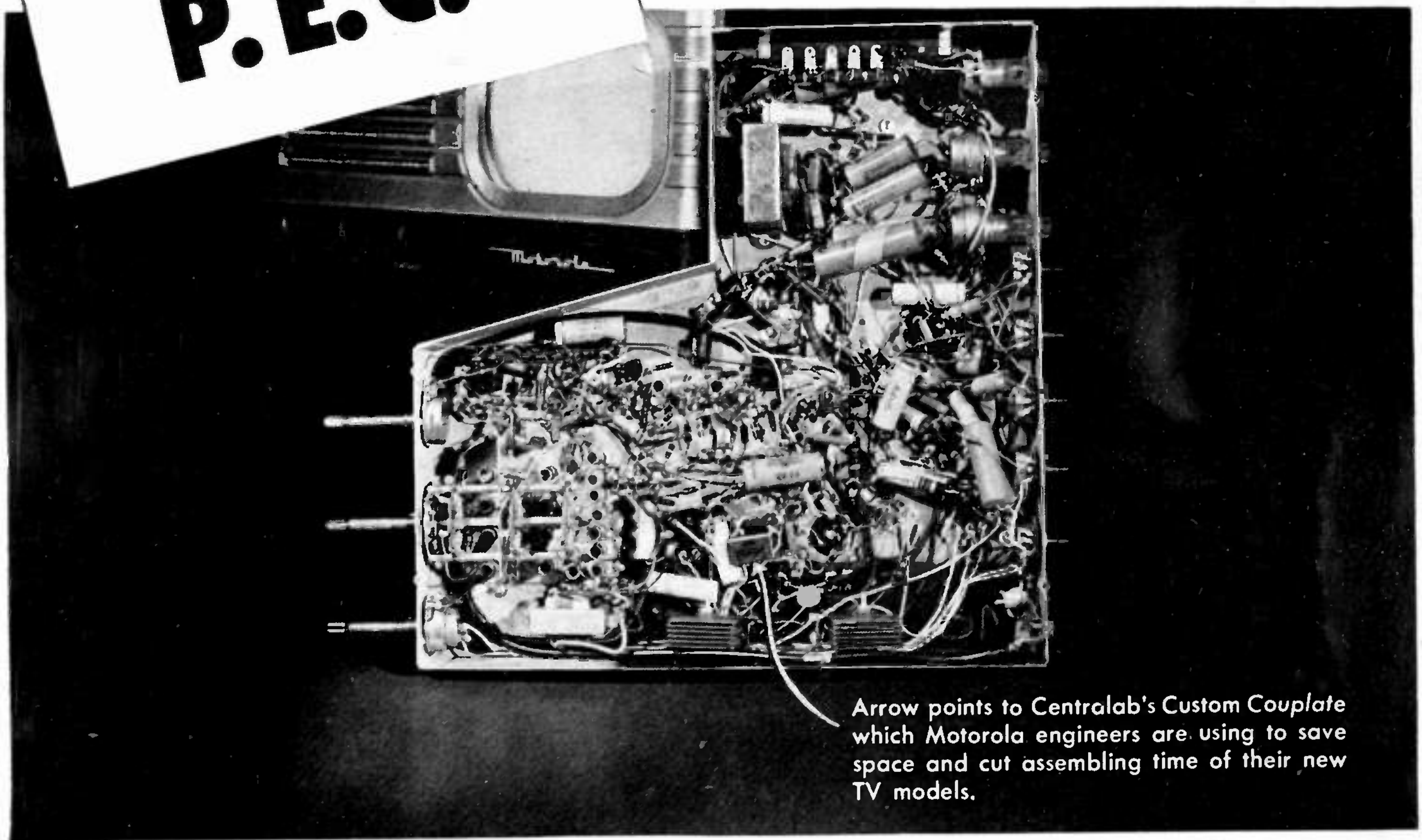
621.396.615.142 1249
Velocity-Modulated Thermionic Tubes [Book Review]—A. H. W. Beck. Cambridge University Press, London, 1948, 180 pp., 15s. (*Phil. Mag.*, vol. 39, p. 1005; December, 1948.) One of the Modern Radio Technique series edited by J. A. Ratcliffe. ". . . a very useful addition to the literature on the subject, presenting much that has only been available in scattered, recent papers."

MISCELLANEOUS

522.1 1250
Proceedings of Observatories—(*Mon. Not. R. Astr. Soc.*, vol. 108, no. 1, pp. 54–80; 1948.) Brief discussions of recent work at various observatories all over the world.

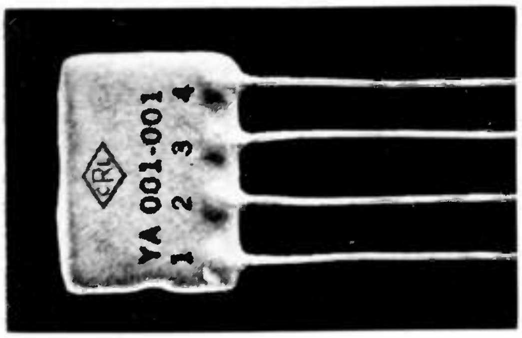
PROGRESS REPORT ON P.E.C.*

How Motorola is using Centralab's "P. E. C." to help build more and finer television receivers than ever before!

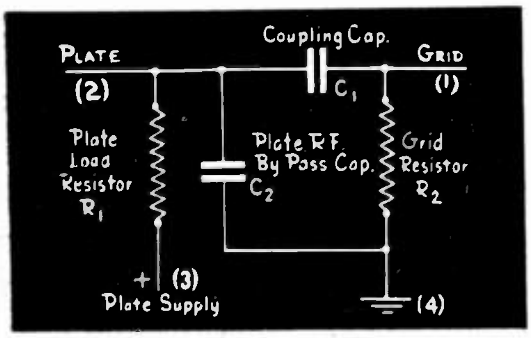


Arrow points to Centralab's Custom *Couplate* which Motorola engineers are using to save space and cut assembling time of their new TV models.

Chassis courtesy of Motorola Corp.



"**COUPLATE**" is made of high dielectric Ceramic X to give long life, low internal inductance, positive resistance to humidity and vibration. A circuit diagram of CRL's *Couplate* is shown below.



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

SPEEDED production and finer products go hand in hand where Centralab's amazing *Printed Electronic Circuit* is concerned. Take the case of Motorola's new television receivers. Engineers for Motorola find that CRL's *Couplate* — a printed interstage coupling plate — saves production time by cutting in half the number of connections to be soldered . . . that it speeds assembly by simplifying wiring operations. They also find that *Couplate* helps them produce finer TV receivers by practically eliminating loose or broken connections — from plate load resistor to coupling capacitor.

Integral Ceramic Construction: Each *Printed Electronic Circuit* is an integral assembly of *Hi-Kap* capacitors and resistors closely bonded to a steatite ceramic plate and mutually connected by means of metallic silver paths "printed" on the base plate.

For complete information about Couplate as well as other CRL *Printed Electronic Circuits*, see your nearest Centralab representative, or write direct.

LOOK TO **Centralab** IN 1949!

Division of **GLOBE-UNION INC., Milwaukee**

SECTION MEETINGS

Atlanta

"The Translator," by W. H. Brattain, Bell Telephone Laboratories; January 21, 1949.

Election of Officers; January 21, 1949.

"Electronic Arithmetic," by C. N. Hoyler, RCA Laboratories, Princeton, N. J.; February 25, 1949.

Baltimore

"The Transistor," by J. N. Shive, Bell Telephone Laboratories; February 23, 1949.

Buffalo-Niagara

"A High-Quality Loudspeaker System," by R. T. Bozak (formerly, Wurlitzer Company) January 19, 1949.

Cedar Rapids

"What Kind of Engineers," by J. D. Ryder, Iowa State College; January 19, 1949.

"Annual Report," by T. A. Hunter, Regional Director of Region 5; January 19, 1949.

Election of Officers; January 19, 1949.

"An Analysis of Distortion Resulting from Two-Path Propagation," by I. H. Gerks, Collins Radio Company; February 16, 1949.

"A Low-Drag Aircraft Antenna for Reception of Omnidirectional Range Signals in the 108 M to 122 M Band," by J. P. Shanklin, Collins Radio Company; February 16, 1949.

Chicago

"The Radio Engineer Looks at Industrial Electronics," by E. D. Cook, General Electric Company; January 21, 1949.

"Sensory-Prosthesis," by N. Wiener, Massachusetts Institute of Technology; February 18, 1949.

Cincinnati

"The Airborne Transponder," by L. B. Hallman, Jr., Wrightfield Communication & Navigation Laboratory; February 15, 1949.

Cleveland

"The Transistor, A New Semiconductor Amplifier," by J. N. Shive, Bell Telephone Laboratories; March 17, 1949.

Connecticut Valley

"New Developments in Cathode-Ray Test Equipment," by W. G. Fockler, A. B. DuMont Laboratories; February 17, 1949.

Dallas-Fort Worth

"A Consideration of the Factors Affecting the Design of Turnstile Antennas," by G. H. Brown, RCA Laboratories, Princeton, N. J.; February 23, 1949.

Dayton

"Meteors and the Earth's Upper Atmosphere," by F. L. Whipple, Harvard College Observatory; March 5, 1949.

Denver

"Cyclotrons and Atom Smashing," by D. B. Harris, Collins Radio Company; March 3, 1949.

Election of officers; March 3, 1949.

Detroit

"The Role of Electronics in Nucleonics," by I. H. McLaren, Bendix Aviation Research Laboratories; February 18, 1949.

Indianapolis

"Intercarrier Sound Systems for Television," by S. W. Seeley, Radio Corporation of America; January 28, 1949.

"Xerography (Electrophotography)," by L. M. Slater, P. R. Mallory & Company, Inc.; February 25, 1949.

(Continued on page 36A)

COSMALITE

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Of course COSMALITE is used by the STANDARD COIL PRODUCTS CO., INC. in their Television and Auto Set Tuning Assemblies. It is the first choice of those who insist on precision and quality products at prices that are right!

The advantages of COSMALITE SHELLS FOR TELEVISION DEFLECTION YOKES are many. Of prime importance is the fact that we have available, without charge, the tools for punching and notching many types of Cosmalite shells and coil forms. This means attractive prices with quicker deliveries.

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* Reg. U. S. Pat. Off.

SECTION MEETINGS

(Continued from page 34A)

LOS ANGELES

"Kinescope Recording," by L. D. Grignon, Twentieth Century Fox Film Corporation and R. M. Fraser, National Broadcasting Company; February 15, 1949.

"Radio—The Hard Way," by G. R. Sams, U. S. Navy Air Missile Test Station; February 15, 1949.

LOUISVILLE

"Tunnel Oscillator," by H. R. Warren, Girdler Corporation; March 11, 1949.

"Radio Aids to Air Navigation," by P. Glasscoe, Civil Aeronautics Administration; February 11, 1949.

NEW MEXICO

"The Transistor," by R. B. Gibney, Los Alamos Scientific Laboratory; February 18, 1949.

OTTAWA

"Receiving Tube Performance at Ultra-High Frequencies," by J. A. Loutit, Canadian Army; February 24, 1949.

PHILADELPHIA

"Some Recent Developments in the Application of Germanium Crystals," by S. T. Martin; Sylvania Electric Products Inc.; March 3, 1949.

PRINCETON

"The Reduction of Common Channel and Multipath Interference by Special FM Receiver Design," by L. B. Arguimbau, and J. Granlund, Massachusetts Institute of Technology; February 10, 1949.

SACRAMENTO

"Stratovision Television," by J. H. Landells, Westinghouse Electric Company; February 8, 1949.

ST. LOUIS

"Stratovision—Airborne Television and FM Broadcasting," by C. E. Nobles, Westinghouse Electric Company; January 27, 1949.

SAN DIEGO

"Design of Regulated Power Supplies," by D. C. Kalbfell, Kalbfell Laboratories Inc.; March 1, 1949.

SALT LAKE

"The Engineer Looks at Industrial Electronics," by E. D. Cook, General Electric Company; February 11, 1949.

SAN ANTONIO

Election of Officers; January 20, 1949.

SAN FRANCISCO

"Experimental Television at 600 Mc," by L. M. Reed and C. N. Winningstad, Television; February 9, 1949.

SEATTLE

"Television Equipment," by J. Frost, Engineering Products Division, RCA, America; January 25, 1949.

TOLEDO

"The Radio Engineer Looks at Industrial Electronics," by E. D. Cook, General Electric Company; February 28, 1949.

TORONTO

"Microwave Relaying," by J. McLean, Philco Corporation; January 24, 1949.

"Television Test Equipment," by C. W. Finnegan, Stromberg-Carlson; February 7, 1949.

WASHINGTON

"Fundamental Limitations of Wide-Band Amplifiers," by H. A. Wheeler, Wheeler Laboratories, Inc.; January 10, 1949.

(Continued on page 37A)

SECTION MEETINGS

(Continued from page 36A)

"Twenty Year's Progress in Radio Diriction Finding," by R. L. Smith-Rose, Junior Past Vice-President of the IRE; March 14, 1949.

"A Circuit Which Picks the Peak of a Low-Frequency Asymmetrical Wave," M. J. Parker, Naval Ordnance Laboratory; February 14, 1949.

"Low-Frequency Synchronized Sawtooth Generator Providing Constant Amplitude Sweep with a Periodic Synchronization Input," P. Yaffee, Naval Ordnance Laboratory; February 14, 1949.

SUBSECTIONS

Amarillo-Lubbock

"New Developments in the VHF Band," by G. F. Reed and J. S. Stover, General Electric Company and Communications Engineering Company, respectively; February 16, 1949.

Hamilton

"Factors Underlying Reproduction of Sound," by H. Goldin, Gaumont-Kalee Ltd.; March 14, 1949.

Lancaster

"Contribution of Electricity to the Safety of Air Navigation," by W. L. Anderson, Pennsylvania Aeronautics Commission; February 17, 1949.

"Some Recent Developments in the Application of Germanium Crystals," by S. T. Martin, Sylvania Electric Products Inc.; March 2, 1949.

Long Island

"The Traveling-Wave Tube and Other Electron Beam Amplifier Tubes," by R. G. E. Hutter, Sylvania Electric Products Inc.; January 26, 1949.

Monmouth

"Television Receiving Antennas," by D. C. Gleckner and A. E. Joost, Antenna Research Laboratory, Inc.; February 16, 1949.

Northern New Jersey

"Infrared Electron Telescopes," by G. L. Krieger, RCA Laboratories, Princeton, N. J.; February 9, 1949.

"Three-Speed Magnetic Tape Recording," by R. H. Ranger, Rangertone Inc.; February 9, 1949.

STUDENT BRANCH MEETINGS

Alabama Polytechnic Institute—IRE Branch

"Dynamic Noise Suppressors," by R. M. Steere, Associate Professor, Alabama Polytechnic Institute; January 31, 1949.

Election of Officers January 31, 1949.

"Radar," by C. Cadell, Student; February 14, 1949.

University of Alberta—IRE Branch

"The Position of the Engineer in Society," by Mr. McDougall, Registrar of the Association of Professional Engineers of Alberta; February 22, 1949.

California State Polytechnic College—IRE Branch

"Instruments in Industry," by H. M. Whittenton, General Electric Company; February 23, 1949.

(Continued on page 38A)

STUDENT BRANCH

MEETINGS

(Continued from page 37A)

College of the City of New York—IRE Branch

"The Coming Convention of the IRE," by E. K. Gannett, Assistant Secretary, The Institute of Radio Engineers; March 1, 1949.

"Quality Control in the Manufacturing of Components for Television," by L. Lutzker, Dumont Laboratories Inc.; March 15, 1949.

Columbia University—IRE-AIEE Branch

"Panel Discussion on Employment," by W. A. Curry, J. A. Balmford, W. LaPierre, L. O'Neill and L. Mason, staff members of the Electrical Engineering Department of Columbia University; March 4, 1949.

University of Dayton—IRE Branch

"Composite Video Signals," by G. Share, Student; February 22, 1949.

University of Florida—IRE-AIEE Branch

"Effective Bandwidth as a Measure of Noise Transmission Properties of Electrical Networks," by W. J. Kessler, Assistant Research Engineer, University of Florida; February 24, 1949.

"The Transistor, A New Semiconductor Amplifier," by J. A. Becker, Bell Telephone Laboratories; March 10, 1949.

The George Washington University—IRE Branch

"Radio Wave Propagation," by J. H. Dellinger, National Bureau of Standards; January 5, 1949.

University of Illinois—IRE-AIEE Branch

"Hallmarks of an Engineer," by W. L. Everitt, Head of Electrical Engineering Department, University of Illinois; February 15, 1949.

Iowa State College—IRE-AIEE Branch

"Industrial Electronics," by T. E. Johnty, General Electric Company; February 21, 1949.

"Mount Palomar Telescope," by L. B. Spinney, Professor of Physics, Iowa State College; March 2, 1949.

State University of Iowa—IRE Branch

"Forward Echo Ranging Transmission Systems," by D. Harris, Collins Radio Corporation; February 23, 1949.

Lafayette College—IRE-AIEE Branch

"Operation and Applications of Electrical Indicating Instruments," by L. F. Parachini, Weston Electrical Instrument Company; February 24, 1949.

University of Maine—IRE-AIEE Branch

"FM Rebroadcasting," by R. W. Hodgkins, Guy Gannett Broadcasting Services; February 17, 1949.

Manhattan College—IRE Branch

"High Fidelity Amplification," by H. T. Sterling, Instructor, Manhattan College; February 23, 1949.

Massachusetts Institute of Technology—IRE-AIEE Branch

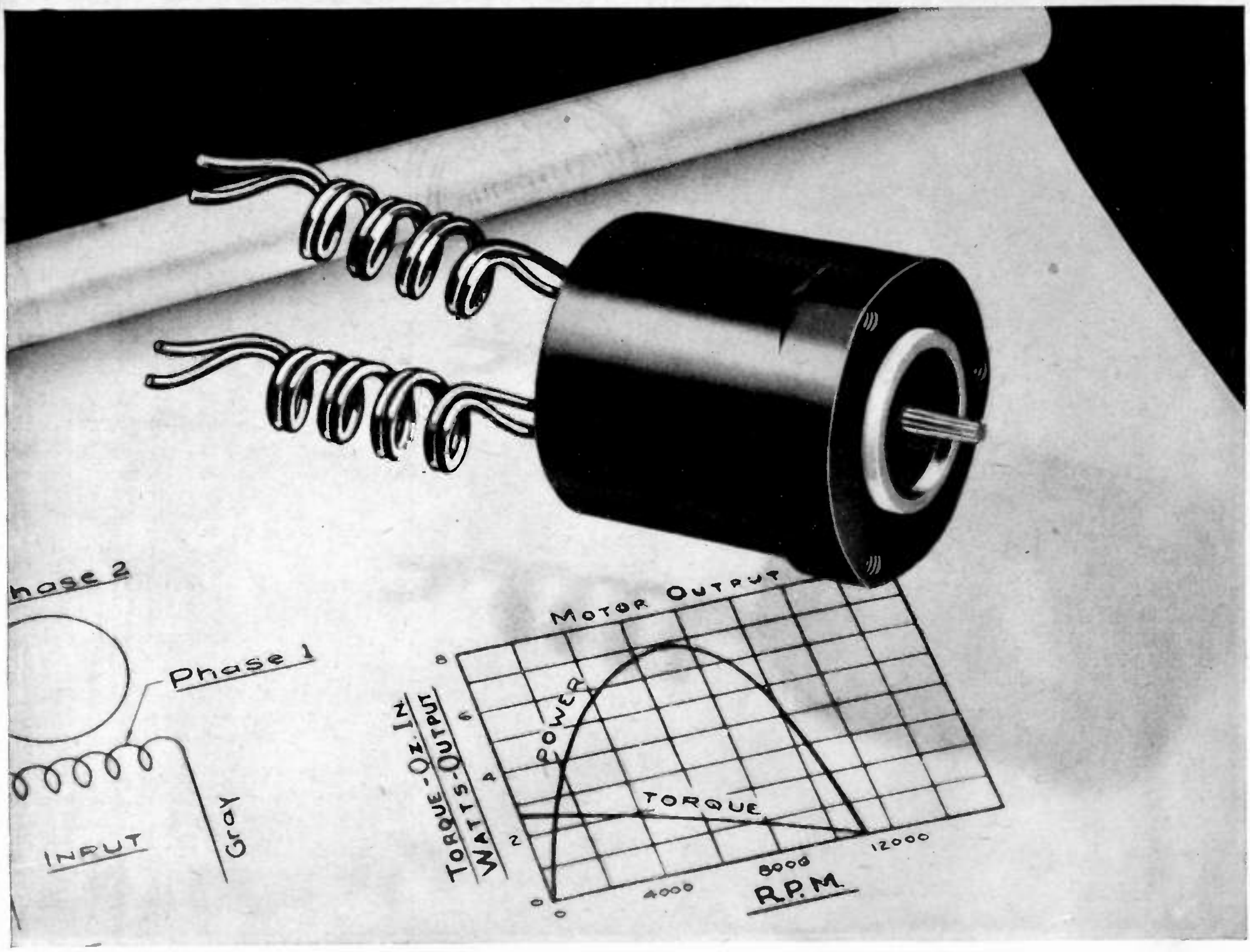
"Planning Your Career," by D. B. Smith, Philco Corporation; February 23, 1949.

Michigan State College—IRE-AIEE Branch

"Television and Microwave Relay," by W. Lawrence, Radio Corporation of America; February 23, 1949.

Election of Officers; March 9, 1949.

(Continued on page 40A)



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This newest addition to the Kollsman line of special-purpose motors is a two-phase, low-inertia induction unit. It is designed for use in 400-cycle servo (null follow-up) systems which require a small motor with an unusually high torque/inertia ratio.

The Model 1318-0460 delivers maximum torque at stall, has a low moment of inertia and will not run single phase. Its frame is fully enclosed. Units with either plain or pinion shaft are available.

The Model 1318-0460 is but one of a complete line of special-purpose motors developed by Kollsman for remote indication and control applications. Complete information concerning any or all of these units is available by addressing: Kollsman Instrument Division, Square D Company, 80-08 45th Avenue, Elmhurst, N. Y.

Characteristics

Frequency (c.p.s.)	400
Phases	2
Speed (r.p.m. — no load)	11,500
Torque (oz./in. — stalled)	2.5
Torque/inertia ratio (radians/sec./sec.)	26,340
Torque/inertia ratio (in. — oz./oz.—in.^{2})	68
Size (dia. x length)	1¾" x 1-47/64"
Weight (ounces)	6.1

KOLLSMAN INSTRUMENT DIVISION

SQUARE D COMPANY

ELMHURST, NEW YORK — GLENDALE, CALIFORNIA

STUDENT BRANCH

MEETINGS

(Continued from page 38)

UNIVERSITY OF MICHIGAN—IRE-AIEE BRANCH

"The Selective-Sequence Electronic Calculator," by H. R. J. Grosch, Columbia University; March 2, 1949.

MISSOURI SCHOOL OF MINES—IRE-AIEE BRANCH

"Electric Welding," by J. D. Vallier, General Electric Company; February 16, 1949.
"What the Electronic Engineer Can Expect of Industry," by W. Bennetsen, Emerson Electric Company; March 2, 1949.

UNIVERSITY OF NEBRASKA—IRE-AIEE BRANCH

"Demonstration of Magic," by Professor Westgate, University of Nebraska; March 2, 1949.

NEW YORK UNIVERSITY—IRE BRANCH

"The Engineer's Part in Television Broadcasting," by O. S. Freeman, Television Station WPIX; March 1, 1949.
"Ways and Means of Getting a Better Paying Job," by W. D. Ingham, Engineering Employment Service, Inc.; March 8, 1949.

PURDUE UNIVERSITY—IRE BRANCH

"Electronic Digital and Analogue Computers," by D. C. Ross, staff member, Purdue University; February 24, 1949.

ST. LOUIS UNIVERSITY—IRE BRANCH

"Presenting the Technical Report," by M. Cochran, staff member, St. Louis University; February 24, 1949.

STANFORD UNIVERSITY—IRE-AIEE BRANCH

"Through the Sonic Barrier," by W. G. Vincenti, Ames Laboratory; January 15, 1949.
"Carnival of Measurements," by J. M. Whittenton, General Electric Company; February 7, 1949.

TEXAS AGRICULTURAL & MECHANICAL COLLEGE —IRE-AIEE BRANCH

"D-C Calculator," by H. R. McKenzie, Student; March 3, 1949.

UNIVERSITY OF TOLEDO—IRE-AIEE BRANCH

"The Radio Engineer Looks at Industrial Electronics," by E. D. Cook, General Electric Company; February 28, 1949.
"Operation of a TV Transmitter," by W. Stringfellow, Radio Station WSPD-TV; March 16, 1949.

UNIVERSITY OF VIRGINIA—IRE-AIEE BRANCH

"Techniques of Cosmic-Ray Measurements Using Rockets," by C. Y. Johnson, Naval Research Laboratories; February 22, 1949.

WAYNE UNIVERSITY—IRE-AIEE BRANCH

"D-C Control Circuits," by Mr. Lee, Department of Street Railways of Detroit; March 3, 1949.

MEMBERSHIP

The following transfers and admissions were approved and will be effective as of May 1, 1949:

Transfer to Senior Member

Ashbrook, F. M., 111-A Ellis St., China Lake, Calif.
Atwood, N. A., U.S.N., New York Naval Shipyard, Brooklyn 1, N. Y.
Bellare, D., 1245 E. Ninth St., Brooklyn 30, N. Y.
Condom Sastre, A., Calleb #56, Vedado, Havana, Cuba
DeSoto, C. B., 1 East 79 St., New York 21, N. Y.
Gegenheimer, W. C., 2402 Fairway Dr., Winston-Salem, N. C.
Jones, R. E., Mayo Clinic, Section on Engineering, Rochester, Minn.
Kennedy, E. D., Eng. & Indust. Experiment Station, University of Florida, Gainesville, Fla.
Kramer, A. W., 639 Lincoln St., Waltham, Mass.
MacAdam, W. K., 25 Bar Beach Rd., Port Washington, L. I., N. Y.
Maki, G. J., Box 1198, R.F.D. 7, Sacramento, Calif.
Massa, E. A., 28 Brookhouse Dr., Marblehead, Mass.
Pickering, W. H., California Institute of Technology, 1201 E. California, Pasadena 4, Calif.
Pierce, R. J., Box 2200, Honolulu, Hawaii
Price, W. E., 61 Nevada Ave., Haddonfield, N. J.
Shank, R. J., Florence Ave. at Teale St., Culver City, Calif.
Sherman, H., 348 Cedar Ave., Long Branch, N. J.
Speller, J. B., 154 Greenridge Ave., White Plains, N. Y.
Stewart, C., 51 Fort Dr., Alexandria, Va.
Summerhayes, H. R., Jr., 1212 Oxford Pl., Schenectady 8, N. Y.
Tittle, C. W., Box 5004, T. C. Station, Denton, Tex.
Toporeck, E. R., 104 B Forrestal, China Lake, Calif.
Wang, C. C., 223-20 65 Ave., Bayside, L. I., N. Y.
Wooldridge, A. C., 67 Prospect Ave., Montclair, N. J.

Admission to Senior Member

Bastedo, G. R., 24 Farragut Pl., N.W., Washington 11, D. C.
Callick, E. B., 1910 K St., N.W., Washington, D. C.
Davis, W. C., 4166 Ardmore Rd., South Euclid, Ohio
Gardiner, J. G., 1014 Madison Ave., Winston-Salem, N. C.
Hogan, C. M., 31 Merzen Ct., Cincinnati 17, Ohio
Sieger, J., Freed Radio Corp., 200 Hudson St., New York 13, N. Y.
Vos, M., Telefon AB L M Ericsson, Stockholm 32, Sweden
Wamsley, D. H., Radio Corporation of America, Lancaster, Pa.

Transfer to Member

Aristel, J., 749 Silver Spring Ave., Silver Spring, Md.
Bartlett, F. E., Radio Station KSO, Des Moines 9, Iowa
Boyer, B. E., 202 Via Alameda, Redondo Beach, Calif.
Chow, W. T., 141-30 Pershing Crescent, Jamaica 2, L. I., N. Y.
Coatney, A. E., 2106 Newton St., Columbus, Ind.
Coursey, J. R., 4700 Monarch St., Dallas 4, Tex.
DeBarro, F., 12, Rue Delta, Heliopolis, Egypt
French, I. F., 646 Humphrey St., Swampscott, Mass.
Gordon, W. E., Franklin Hall, Cornell University, Ithaca, N. Y.
Grdseloff, L., c/o Green's Commercial Agencies, P.O.B. 600, Cairo, Egypt
Hall, G. G., c/o Tecnico Ltd., Box 12, Marrickville, N.S.W., Australia

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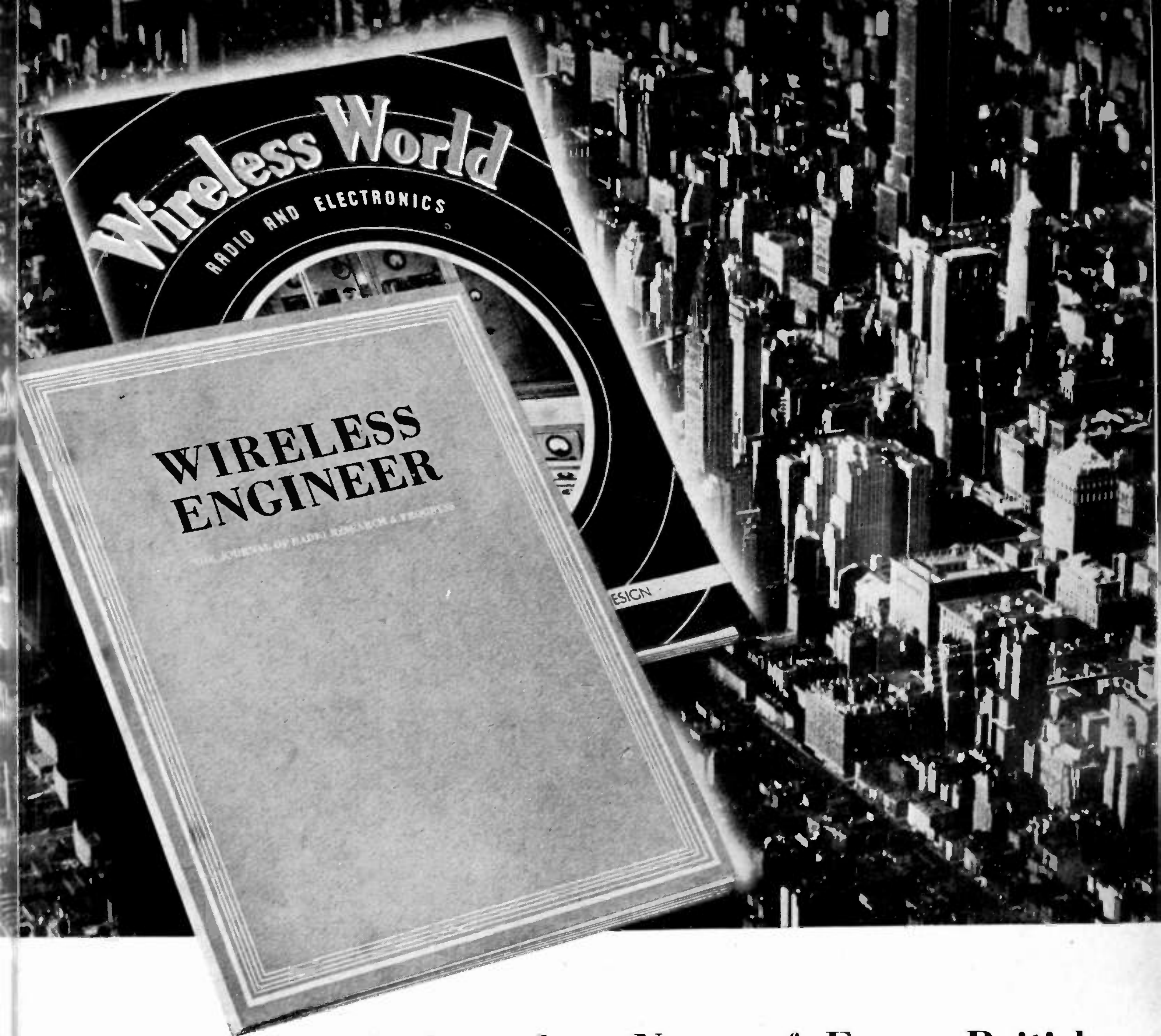
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RADIO AND ELECTRONICS
WIRELESS ENGINEER

MEMBERSHIP

(Continued from page 42A)

Israel, L. J., U.S.A.F., Hdqs. 15 Air Force, Colorado Springs, Colo.
Kafafian, J., 1544 Union St., San Francisco 23, Calif.
Martin, J. L., 4226 W. 13 St., Amarillo, Tex.
McCarley, H. H., U. S. Navy, U.S.S. Hugh Purvis (DD708), c/o Fleet Post Office, New York, N. Y.
McLeod, J. W., 7549 S. Wabash Ave., Chicago 19, Ill.
Merchant, J. H., 2 Cedar St., Binghamton 47, N. Y.
Orwin, R. J., 4518 36 St. S., Arlington, Va.
Overmire, M. O., 12147 Huston St., North Hollywood, Calif.
Papenfuss, C. A. C., 6845 13 Avé., S., Minneapolis 19, Minn.
Parker, A. B., 3835 Voltaire St., San Diego 7, Calif.
Pattinson, T. H., 47, Rodenhurst Rd., London, S.W. 4, England
Richelieu, C. C., 23 Bennett Rd., Gardner, Mass.
Ruble, H. E., 3011 Athens Ave., Dayton 6, Ohio
Smith, D. P., Washington St., Sparkill, N. Y.
Stanton, L. J., 151 Eighth Ave., New York 11, N. Y.
Tschannen, R. F., 412 E. Maple St., Lombard, Ill.

Admission to Member

Acosta, R. J., #872 Cabrera St., Cabrera Urbanization, Rio Piedras, Puerto Rico
Arbuckle, F. M., RCA Princeton Laboratories, Princeton, N. J.
Bryson, H. C., Jr., 5428 Norfolk St., Philadelpha 43, Pa.
Covington, A. E., Box 388, Billings Bridge, Ont., Canada
Desi, G. R., Hazeltine Electronic Corp., Little Neck, L. I., N. Y.
Elliott, H. R., Apt. 3, 10 Bacon, Waltham 54, Mass.
Ferrara, G., 236 Withington Ave., Ferndale 20, Mich.
Ferre, M. D., Schlumberger Well Surveying Corp. Box 550, Old Quarry Rd., Ridgefield, Conn.
Fox, J. C., 2140 Alta, Louisville 5, Ky.
Godbey, J. K., 3120 Dutton Dr., Dallas 11, Tex.
Gustinelli, J. L., Vaubien, Orsay, Seine-et-Oise, France
Hayes, T. J., 36 N. Horton St., Dayton 3, Ohio
Hughes, R. W., 200 Lorraine Ave., Upper Montclair, N. J.
Johnson, R. R., Box 76, HAFB, c/o North American Aviation, Inc., Alamogordo, N. Mex.
Lescarbeau, R. F., 31 Suffolk Dr., East Hartford, Conn.
Matthews, R. D., R-145, Sandia Base Branch, Albuquerque, N. M.
Melhose, A. E., 635 Glen Ave., Westfield, N. J.
Morgan, H. L., 2312 El Cajon Blvd., San Diego 3, Calif.
Norris, W. E., 5401 Madison St., 2-C, Richmond, Calif.
Piecha, L. M., 2218 S. Leavitt St., Chicago, Ill.
Rao, K. L. N., Department of Radio Technology, S. J. Occupational Institute, Bangalore 1, India.
Rifkin, M. S., 144-46 69 Ave., Kew Gardens Hills, L. I., N. Y.
Roberts, G. A., 408 College Ave., Swarthmore, Pa.
Schlicke, H. M., 54 Juniper Rd., Port Washington, L. I., N. Y.
Simons, F. A., c/o CAA, International Services Division, A-36, Washington 25, D. C.
Tellez, J. M. B., 147 Aguascalientes, Mexico D. F., Mexico
Thompson, W. C., Jr., B.O.Q. B-3, China Lake, Calif.
Tolbert, C. W., 502 Oakland, Austin, Tex.
Uslan, S. D., 2202 East 28 St., Brooklyn 29, N. Y.
Warner, A. H., 603 Essex Circle, China Lake, Calif.
Weber, M., 516 N. Adams St., Albuquerque, N. M.
Woodyard, O. C., 113 Atkins Ave., Neptune, N. J.

(Continued on page 45A)

MEMBERSHIP

(Continued from page 44A)

The following admissions to Associate grade were approved to be effective as of April 1, 1949:

Alberts, H. C., 112-45 150 St., St. Albans 12, L. I., N. Y.
Alexander, R., c/o Sylvania Electric Products Inc., 40-22 Lawrence St., Flushing, L. I., N. Y.
Alston, W. H., 747 Webster Pl., Plainfield, N. J.
Alter, R. S., 59 Vreeland Ave., Clifton, N. J.
Aman, O. B., 49 W. 167 St., New York 52, N. Y.
Anderson, D. E., C.R.E.I., 16 St. and Park Rd. N.W., Washington 10, D. C.
Applebaum, A., 15 Elmira St., S.E., Washington 20, D. C.
Archer, L. A., 2300 W. Davis St., Dallas 11, Tex.
Arimura, T., 4436 S. Woodlawn Ave., Chicago 15 Ill.
Asdal, M. K., 463 Page Ave., Lyndhurst, N. J.
Atlee, R. Y., 50 Doyer Ave., White Plains, N. Y.
Bailey, E. M., 6 Home Ct., Stamford, Conn.
Baird, J. A., 38 Valley View Rd., Rockaway, N. J.
Bardsin, S., 1050 Stratford Ave., New York 59, N. Y.
Basole, G. M., 391, Sultan Bazar, Hyderabad, Hyderabad (DN), India
Baum, L., 634 Montgomery St., Brooklyn 25, N. Y.
Baumann, L. S., 525 Jennings Ave., Bridgeport 4, Conn.
Berardi, C. S., 3010 N. Washtenaw Ave., Chicago 18, Ill.
Beresford, D. R., 200 A Brynes, N.O.T.S., China Lake, Calif.
Berliner, J., 31 Cedar Ave., Apt. 29, West End, N. J.
Bernstein, B., 590 Clinton Ave., Newark 8, N. J.
Bestich, J. P., 4670 Grant St., Gary, Ind.
Billig, L. S., 140 Edgewater Apts., Essex 21, Md.
Bocchieri, A. J., 1504 Monroe St., Washington 10, D. C.
Bonney, H. W., 125-20 82 Ave., Kew Gardens 15, L. I., N. Y.
Boogers, N. H., 59 Gresham Rd., Toronto 12, Ont., Canada
Booker, C. A., Jr., 8001 Susquehanna St., Wilkinsburg 21, Pa.
Bremmer, L. F., 14269 24 Ave., S.W., Seattle 66, Wash.
Budd, W. E., 136-39 39 Ave., Flushing, L. I., N. Y.
Butenhoff, R. L., 132 Taylor Rd., Oak Ridge, Tenn.
Calvarese, J. F., Box 148, Valley Forge Rd., Wayne, Pa.
Canfora, A. E., 6501 17 Ave., Brooklyn 4, N. Y.
Cantor, T. E., 3015 Riverdale Ave., Riverdale 63, N. Y.
Caquelin, M. W., 123 17 St., N.E., Cedar Rapids, Iowa
Carter, W. R., 208-20 Cross Island Pkwy., Bayside, L. I., N. Y.
Casey, T. J., 5300 W. 84 St., Minneapolis 20, Minn.
Chambers, R. N., 182 Oakridge Ave., Summit, N. J.
Coffe, S. A., 1592 Carlyon Rd., E. Cleveland 12, Ohio
Coonley, K. E., 825 President St., Brooklyn, N. Y.
Cope, R. H., Jr., 23 S. Washington St., Port Washington, L. I., N. Y.
Cornet, J. F., 281 Grove St., Montclair, N. J.
Cottam, R. W., 68 Hillcrest Rd., Mount Vernon, N. Y.
Cotton, S., 72 Barker Ave., Eatontown, N. J.
Court, H. R., Jr., 159 Hoffman Ave., Elmont, N. Y.
Crawford, C. A., 1136 Milwaukee Ave., Chicago 22, Ill.
Cutler, M. L., 2412 Colston Dr., Silver Spring, Md.
Daniels, T., Prairie View A & M College, Prairie View, Tex.
Davis, D. P., 1408 Commonwealth Ave., West Newton 65, Mass.

(Continued on page 46A)

MEMBERSHIP

(Continued from page 45A)

Dawson, R. K., C.R.E.I., 16 St. & Park Rd., N.W., Washington, D. C.
Dechovitz, H., 1619 Washington Ave., New York 57, N. Y.
Decker, K. M., 4706 Nichols Ave., S.W., Washington 20, D. C.
Deerhake, W. J., Watson Laboratories, 612 W. 116 St., New York 27, N. Y.
Dembinski, L. J., 4314 Parkwood Ave., Baltimore 6, Md.
Diorio, J., 27 Euclid Ave., New York 8, N. Y.
Donnell, W. C., Jr., 138 W. 65 St., New York 23, N. Y.
Doran, J. W., 2936 Bainbridge Ave., New York 58, N. Y.
Dunson, P. M., 1978 Dartmouth Ave., Columbus 3, Ohio
Edmonds, J. B., Apt. 1-B, Geo. Legare Homes, Naval Base 54, S. C.
Ehrlich, N., Rm. 867A, 463 West St., New York 14, N. Y.
Ellis, A. H., 172 Lawn Ave., Stamford, Conn.
Evans, T., 7 Maesquarre Rd., Ammanford, Carm., England.
Fairman, H. E., 41-41 41 St., Long Island City 4, L. I., N. Y.
Fannin, B. M., 226 Eddy St., Ithaca, N. Y.
Fanwick, C., 518 Watchung Ave., Plainfield, N. J.
Feher, E., 226 W. 79 St., New York 24, N. Y.
Fenster, A., 35-50 82 St., Jackson Heights, L. I., N. Y.
Filo, J., Jr., 5653 N. Kenmore Ave., Chicago 40, Ill.
Fisher, E. C., 9 Southwood Ave., Highcliffe, Christchurch, Hants., England
Fogel, L. J., 77-12 35 Ave., Jackson Heights, L. I., N. Y.
Foley, W. V., 594 Edgewater Apts., Baltimore 21, Md.
Forsythe, E. H., 10118 Curtis Rd., Detroit 21, Mich.
Fulwiler, R. J., 3000 Windsor Ave., Waco, Tex.
Fury, A. E., 9-23 124 St., College Point, L. I., N. Y.
Gardiner, H. E., 2830 Fifth St., Port Arthur, Tex.
Gellert, Max E., 1406 Eutaw Pl., Baltimore 17, Md.
Gillette, F. N., General Precision Laboratory, Pleasantville, N. Y.
Gledhill, A. P., E/S Homan Ave., Blue Point, L. I., N. Y.
Gluck, S. E., Moore School of Electrical Engineering, 200 S. 33 St., Philadelphia 4, Pa.
Goetz, J. A., 87 Madison Ave., Nutley 10, N. J.
Goodman, D. M., 55 Mill Lane, Levittown, Hicksville, L. I., N. Y.
Gouban, T. T. E., 93 Columbia Ave., Long Branch, N. J.
Gramhich, H. F., 808 Superior, Toledo 4, Ohio
Grant, C. R., 2131 Suitland Ter., S.E., Washington 20, D. C.
Grant, M. S., 2037 Hoyt Ave., Fort Lee, N. J.
Green, E. H., Department of Physics, Brooklyn College, Brooklyn 10, N. Y.
Green, W. F., 312 Webster Ave., Brooklyn 30, N. Y.
Greenwood, J. H., 166 N. Sprague Ave., Pittsburgh 2, Pa.
Gridley, D. H., 3926 First St., S.W., Washington 20, D. C.
Grosvenor, H. D., 1332 Maple Ave., Fort Wayne 6, Ind.
Gruenloh, G. J., 3721 N. Kostner St., Chicago 41, Ill.
Grundy, A. B., Jr., 325 Main St., Cromwell, Conn.
Gurviz, A. A., Calle San Gervasio 93, Barcelona, Spain
Guzzardo, B. A., 2700 Live Oak, Waco, Tex.
Haas, D. L., 2215 Adams Ave., Norwood 12, Ohio
Hales, E. B., Box 422, Mt. Kisco, N. Y.
Hampshire, J. R., Rochester Gas & Electric Corp., East Ave., Rochester 4, N. Y.

(Continued on page 47A)

MEMBERSHIP

(Continued from page 46A)

Harmatuk, S. N., 1575 Odell St., New York 62, N. Y.
Hassler, N. A., 215 Middle Neck Rd., Great Neck, L. I., N. Y.
Heritage, J. L., 1669½ Sunset Cliffs Blvd., San Diego 7, Calif.
Herman, L. C., 85 Nassau Rd., Great Neck, L. I., N. Y.
Hicks, J. M., 304 N. George St., Goldsboro, N. C.
Hirschel, L. R., 503 Chillum Rd., Chillum, Md.
Hoffman, R. R., Jr., 2714 Bainbridge Ave., New York 56, N. Y.
Holdman, M., 160 Vassar Ave., Newark 8, N. J.
Hubbard, C. L., 205 Humble Bldg., Houston, Tex.
Irick, J., 1838 Edge Hill Rd., Huntington Valley, Pa.
Jacob, F., 1620 Chilton St., Baltimore 18, Md.
Jacobson, R. S., 31 Kenwood Rd., Garden City, L. I., N. Y.
Jednacz, J. E., 125 Monroe Ave., Merchantville, N. J.
Jouppi, A. F., 30 Wilwade Rd., Lake Success Hills, Great Neck, L. I., N. Y.
Judin, H., 4200 Atlantic Ave., Brooklyn 24, N. Y.
Jurow, S., 192 Pembroke St., Brooklyn, N. Y.
Katz, G. M., 870 Riverside Dr., New York, N. Y.
Keen, H. S., 697 Waypark Ave., E. Hempstead, L. I., N. Y.
King, P. D., 125-06 Francis Lewis Blvd., Laurelton, L. I., N. Y.
King, R. W., Jr., 16 Tompkins Rd., Scarsdale, N. Y.
Klizas, P. A., 328 West Fourth, Emporium, Pa.
Knight, W. C., 854 Kimball Ave., Westfield, N. J.
Kram, W., 1487 E. 37 St., Brooklyn 10, N. Y.
Kuder, M., 3211 Wheeler Rd., S.E., Washington 20, D. C.
Kulinyi, R. A., 167 Newman Springs Rd., Red Bank, N. J.
Laitinen, P. O., 119 Kinship Rd., Baltimore 22, Md.
Lane, A. L., 811 Otis Pl., Washington, D. C.
Langberg, E., 59 Dean St., Aberdare, Glam, Great Britain
Lanxon, R. C., 625 A. C. & W. Sqn., A.P.O. 942, c/o Postmaster, Seattle, Wash.
Leavitt, J. N., 8 King George Crescent, Batawa, Ont., Canada
Lefgren, A. H., George Westinghouse Vocational H. S., 49 Flathush Ave., Ext., Brooklyn 1, N. Y.
Leporati, R. A., 13414 Kelso Ave., Cleveland, Ohio
Lieberman, H., 62 Hadley Ave., Clifton, N. J.
Lilly, J. L., 138-18 97 Ave., Jamaica 4, L. I., N. Y.
Loffredi, A. J., R.D. 6, Washington, Pa.
Loiacono, J. V., 110 E. Fourth St., New York 3, N. Y.
Loth, P. A., 8500 107 St., Richmond Hill 18, L. I., N. Y.
Lowe, M., 1941 82 St., Brooklyn 14, N. Y.
Lowenstein, M., 208 E. 72 St., New Yok 21, N.Y.
MacCallum, J. M., Jr., 23 S. Washington St., Port Washington, L. I., N. Y.
Madaris, F. W., Timken Roller Bearing Co., c/o Steel & Tube Div. Library, Canton 6, Ohio
Magee, T. T., 111 Gregory Ave., West Orange, N. J.
Maltz, J. A., 1011 Rutland Rd., Brooklyn 12, N. Y.
Mani, K. S., Krishnaswami Subramanian, c/o V. S. Rajagopalan & Co., Ltd., Avanashi Rd., Coimbatore, South India
Marcus, D. H., 614 Fifth Ave., Asbury Park, N. J.
Marin, F. L., 649 E. 14 St., New York 9, N. Y.
Marks, T. P., U. S. Navy, U.S.S. Albany CA123, c/o FPO, New York, N. Y.
Martin, W. H., 133 Prince St., Marblehead, Mass.
Masse, C. J., 170 10 St., Oakland, Calif.
Mataresе, J., 4144 Paulding Ave., New York 66, N. Y.
Matthews, J. G., 70 Old Army Rd., Bernardsville, N. J.

(Continued on page 48A)

Membership

(Continued from page 47A)

Mayo, N. D., 83-30 Kew Gardens Rd., Kew Gardens, L. I., N. Y.
McKnight, B. E., 230 E. 32 St., New York 16, N. Y.
Mekota, J. E., 64 Oxford St., Apt. 1, Cambridge, Mass.
Mesh, T. J., 66 Garfield Ave., Easthampton, Mass.
Moore, E. N., 6337 N. Kenmore St., Chicago 40, Ill.
Mott, E. D., Radio Station CKGB, Timmins, Ont., Canada
Muchmore, D. E., 214-07 47 Ave., Bayside, L. I., N. Y.
Najmy, F. A., 17 Strong Pl., Brooklyn 2, N. Y.
Obsharsky, P., 165 Broad St., Red Bank, N. J.
O'Donnell, N. A., 48 Drexel Rd., Buffalo 14, N. Y.
Owyoung, J. T., 1400 Jackson St., San Francisco 9, Calif.
Pack, E. L., 98 Tyler St., Pittsfield, Mass.
Paulin, W. P., 155 St. George St., Toronto, Ont., Canada
Phelps, R. J., 255 Raymond St., Rockville Center, L. I., N. Y.
Pierce, M. E., Radio Station KSOO, 317 S. Phillips Ave., Sioux Falls, S. Dak.
Plotkin, A. A., 422 W. Fifth St., Plainfield, N. J.
Powell, F. P., 1847 Marshall Rd., Baltimore 22, Md.
Powell, H. H., 338 Windermere Ave., Lansdowne, Pa.
Prest, T. A., Box 20, Goderich, Ont., Canada
Rao, D. V. R., Wireless College, Colwyn Bay, Wales
Ralph, D. J., 1817 Keerny St., N.E., Washington, D. C.
Rawson, R. J., 144-C Wallworth Park Apts., Haddonfield, N. J.
Reeves, D. B., 49 Prospect St., Madison, N. J.
Rehnke, C. H., 260-17 75 Ave., Floral Park, L. I., N. Y.
Ritchey, N. B., 164 E. Rogues Path, Huntington Station, L. I., N. Y.
Rizos, J. G., 120 Greene St., Osborn, Ohio
Robb, D. T., 5 Oak Knoll Rd., Summit, N. J.
Roberts, A. K., 131 Audley St., (82 Drive), Kew Gardens, L. I., N. Y.
Roberts, J. D., 1230 Wood Ave., Waco, Tex.
Robertson, G. H., 730 Ridgewood Rd., Maplewood, N. J.
Rodman, M. D., RCA Bldg. 13-7, Camden, N. J.
Roethle, R. H., 8030 N. Division Rd., Milwaukee 11, Wis.
Rosenson, A. H., 420 Lincoln Rd., Miami Beach, Fla.
Rotman, W., 86 Franklin Ave., Revere 51, Mass.
Rowland, H. L., British Embassy, Washington, D. C.
Rudolph, I. W., 5344 Vancouver Ave., Detroit 4, Mich.
Sage, T. C., 1420 W. Third St., Topeka, Kan.
Sakaguchi, H. K., R.F.D. 2, Brighton, Colo.
Sanchez, H. M., 30-25 72 St., Jackson Heights, L. I., N. Y.
Sanders, M., 3115 Parker Ave., Silver Spring, Md.
Sanders, R. W., Stanford University, Stanford, Calif.
Sard, E. W., 49 Somerset Dr., Great Neck, L. I., N. Y.
Sarkissian, H. H., 1802 Harper Ave., Hermosa Beach, Calif.
Seymour, J. W., 1527 New Hyde Park Dr., New Hyde Park, L. I., N. Y.
Shapiro, L., Free Acres, Scotch Plains, N. J.
Shapiro, R. I., 401 Gregory Ave., Wilmette, Ill.
Sheldon, G. N., 813 Dakin St., Chicago 13, Ill.
Simpson, C. A., 3 Park St., Norwalk, Conn.
Sims, R. U., 311 Barnes St., Pittsburgh 21, Pa.
Smith, E. D., Box 251, Lexington Park, Md.
Sprung, R. A., Robinson Aviation, Inc., Teterboro, N. J.
Statsinger, J., 6 W. 77 St., New York 24, N. Y.
Steinberg, W. A., 3256 Bedford Ave., Brooklyn 10, N. Y.
Stell, J. P., 1707 Kenyon St., Washington 10, D. C.
Sulkowski, E. L., 513 Euclid Ave., Canonsburg, Pa.
Suszek, E. J., 29-31 Sunset Ct., Passaic, N. J.
Swedloff, B., 2162 N. Eighth St., Philadelphia 22, Pa.

(Continued on page 49A)

Membership

(Continued from page 48A)

Theodoseau, T., 85-70 144 St., Jamaica 2, L. I., N. Y.
Thomas, T. R., E. County Line Rd., Hatboro, Pa.
Thornton, C. H., 2577 Edsel Dr., Trenton, Mich.
Townsend, C. K., Jennings Radio Mfg. Co., Box 1278, San Jose, Calif.
Traver, L. L., 1780 Broadway, San Francisco 9, Calif.
Tuck, E., Bldg. 1-021, Naval Ordnance Laboratory White Oak, Silver Spring, Md.
Tyrolf, E. O., Jr., 2226 Bartholomew St., New Orleans 17, La.
Van Zant, R. R., 7306 16 Ave., Takoma Park 12, Md.
Vaughan, E. M., 4314 15 St., N.W., Washington 11, D. C.
Walsh, J. V., 501 E. 234 St., Apt. 4E, New York 66, N. Y.
Ward, J. C., Manville Lane, Pleasantville, N. Y.
Weitz, S., 1034½ Sewall Ave., Asbury Park, N. J.
Wellock, R. R., 214-07 47 Ave., Bayside, L. I., N. Y.
Whitlock, M. C., 5234 W. Cuyler Ave., Chicago 41, Ill.
Williams, R. J., III, 1586 Gerry Walk, Camden, N. J.
Winenow, R. E., 2500 Chestnut St., San Francisco 23, Calif.
Woodward, C. P., 1664 Wakeling St., Philadelphia 24, Pa.
Zimmerman, C. W., 4623 Newmoore Ave., Dallas 9, Tex.

The following transfers to Associate grade were approved to be effective as of February 1, 1949:

Ball, L. D., 152 School House Lane, Philadelphia 44, Pa.
Beck, E. A., 1220 Emerson St., N.W., Washington 7, D. C.
Bell, A. C., 416 W. Sixth, Emporium, Pa.
Bellew, F. N., R.F.D. 1, Tompkins Ave., South Plainfield, N. J.
Eikelman, J. A., Jr., 325 Winnipeg Pl., Long Beach 4, Calif.
Elice, S. S., 252 Grand St., Brooklyn, N. Y.
Fleming, E. W., 1220 Quincy St., N.W., Washington, D. C.
Forbath, F. P., 4296 Vineland Ave., North Hollywood, Calif.
Gildersleeve, R. E., Baldwinsville, N. Y.
Glomb, W. L., 448 Prospect St., South Orange, N. J.
Grantham, R. B., 415 Greenwood Ave., Topeka, Kan.
Greenwood, R. W., 1334 Hale St., Philadelphia 11, Pa.
Harper, E. B., Jr., 830 N. Harrison Ave., East Lansing, Mich.
Harris, B. L., 3013 22 N., Seattle, Wash.
Hartman, R. B., University Club, Golden Hill St., Bridgeport, Conn.
Hoffer, G. J., 205 E. 32 St., Brooklyn 26, N. Y.
Ketchum, J. R., 12588 Lake Ave., Lakewood 7, Ohio
Kolb, A. F., 421 Abert Pl., Milwaukee 12, Wis.
Latham, N. D., 2010 Narberth Ave., Haddon Heights, N. J.
Perry, B. A., Port Murray, N. J.
Portnoy, A., 3130 Brighton 7 St., Brooklyn 24, N. Y.
Rivers, T. A., 37 Waverly Way, N. E., Atlanta, Ga.
Rogell, P., 38-07 Little Neck Pkwy., Little Neck, L. I., N. Y.
Rubin, R., 90 Fuller St., Brookline 46, Mass.
Slavin, M., 1718 Yakona Rd., Towson 4, Md.
Steffen, E. W., Jr., c/o B.A.R., East Akron Station, Akron 5, Ohio
Tanana, E. J., 2854 W. 21 St., Chicago 23, Ill.
Turner, T. D., 2[illegible]9 Broadway, Fort Wayne, Ind.
Walder, B., 1937 Loring Pl., New York 53, N. Y.
Williamson, J. C., 39-D Oak Grove Dr., Baltimore 20, Md.
Wood, A. G., Apt. 18D, Irvington Gardens, Irvington, N. Y.

The continued expansion of the

NATIONAL UNION RESEARCH DIVISION

has created many fine positions for men interested in

VACUUM TUBE RESEARCH

Our laboratories are devoted entirely to research and development work on vacuum tubes. These include cathode ray, microwave, receiving, radial beam, subminiatures, and various special tube types. We are in a position to offer an interesting job, a stable future, and ideal working conditions to men who are qualified. Men with vacuum tube or similar experience and with degrees in Physics or Engineering are needed at the present time. Recent graduates without experience but with degrees in Physics or Electrical Engineering, as well as tube or circuit technicians, are invited to apply.

Before you decide on your future connection, be sure to look into the opportunities National Union has to offer.

Send resume to: Divisional Personnel Manager
National Union Research Division
350 Scotland Road
Orange, New Jersey

WANTED PHYSICISTS ENGINEERS

Engineering laboratory of precision instrument manufacturer has interesting opportunities for graduate engineers with research, design and/or development experience on radio communication systems, Servomechanisms (closed loop), electronic & mechanical aeronautical navigation instruments and ultra-high frequency & microwave technique.

WRITE FULL DETAILS TO EMPLOYMENT SECTION

SPERRY GYROSCOPE COMPANY
DIVISION OF SPERRY CORP.
Marcus Ave. & Lakeville Rd.
Lake Success, L.I.

TEC

ENGINEERS

TELEVISION EQUIPMENT CORPORATION offers an unusual opportunity for qualified engineers to participate in the growth of the rapidly expanding Industrial Television field.

Experience in video, pulse, electronic display and circuit techniques with good educational background required. Several positions are available ranging from junior to project engineers.

Compensation is commensurate with ability. Opportunity to participate in company profit sharing plan.

Apply by letter to Leonard Mautner, Vice President:

TELEVISION EQUIPMENT CORPORATION
238 William Street
New York 7, 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.

VACUUM TUBE ENGINEER

Vacuum tube engineer (starting salary $6235.20) to head well equipped and well staffed vacuum tube development shop in Government laboratory. Requires engineer or physicist with several years broad experience in vacuum tube development and construction and a genuine interest in developing new tube making techniques. Box 559.

ELECTRONIC ENGINEER

To work on challenging control projects of an electrical, electro-mechanical and electronic nature. Good pay with excellent future with an established New England company of top rating for young man with necessary background who has initiative and imagination. Box 560.

DESIGN ENGINEERS

We have several immediate openings for design or development engineers. Mechanical and electrical engineers with considerable experience in design of instruments or control preferred. Excellent opportunities with leading manufacturer of automatic controls. Attractive salary. Location: Minneaolis. Contact J. A. Johnson, Employment Office, Minneapolis-Honeywell Regulator Co., 4th Ave., and 28th St., Minneapolis 8, Minn.

RADIO PROJECT ENGINEER

Graduate engineer with at least 5 years recent experience in design and development of low power oscillator and amplifier circuits as used in signal generators in the very high frequency range. Must be familiar with theoretical concepts and calculations of the circuit components, as well as practical design and layout work. Applicant should have initiative and supervisory ability and be capable of assuming full responsibility for project. Apply by letter only. Address Personnel Dept. Federal Mfg. & Engineering Corp., Brooklyn 5, N.Y.

ELECTRONIC AND SALES ENGINEERS

Recently established and rapidly expanding company has openings for Junior electronic engineers in development and field engineering. Also openings for sales engineers in south, southeast, Pennsylvania and New England. Attractive remuneration. Company specializes in industrial electronics, instrumentation and control. Write giving full details to Fielden Electronics Inc., Huntington Station, N.Y.

ELECTRONICS RESEARCH ENGINEER

Electronics research engineer experienced in electronics as well as engineering sciences and physics for research and de-

(Continued on page 51A)

POSITIONS OPEN

(Continued from page 50A)

velopment work in connection with motor and generator control equipment. Excellent opportunity with large manufacturer in New York City. Please write, giving complete education and experience. Replies will be held in confidence. Our employees know of this ad. Box 561.

ENGINEERS

Location, Phoenix, Arizona. Excellent working conditions. Housing available. Motorola, Inc. announces a Research Laboratory devoted to armed service contract and company research in microwave, mobile communications, supervisory control, telemetering, miniaturization, and aviation electronics. Only fully qualified experienced inventors, engineers and scientists should apply. Send detailed statement of education and experience to Motorola Inc., D. E. Noble, 4545 Augusta Blvd., Chicago 51, Ill.

SENIOR AND JUNIOR ENGINEERS

Senior and Junior engineers needed with experience on SCR-584 radar or similar equipment. Location about 50 miles from Los Angeles. Electronic Engineering Co., 2008 W. 7th St., Los Angeles 5, Calif.

ELECTRONICS ENGINEER

Radio and industrial electronics instructor for two year technical college—Extension Division of Georgia Institute of Technology, Atlanta, Georgia. Write The Technical Institute, Chamblee, Georgia.

ENGINEERS

Men needed immediately for permanent positions on an experimental, development, and production program of complex electronic and electro-mechanical equipment. Work covers computers, servos, amplifiers, instrumentation, small mechanisms in aircraft simulation for complex training equipment. Applicants must have college degree, equivalent experience or both. Apply Personnel Mgr., Link Aviation, Binghampton, N.Y.

ELECTRO-ACOUSTIC ENGINEER

An attractive opportunity for an experienced engineer with thorough electro-acoustic training (transducers) in a small, sound rapidly growing concern. Permanent position with future opportunity as a research director. Write giving full details to Box 563.

RESEARCH SCIENTISTS

Experienced research scientists with advanced degrees and experience in physics, aerodynamics, electronics, optics, mathematics, chemistry, metallurgy, or meteorology to perform supervisory research and act as permanent consulting group to engineering laboratories. Excellent opportunity for men with right qualifications. Salary \$8,000-\$14,000 bracket. Write Box 564.

DEVELOPMENT ENGINEER

Newly formed and rapidly expanding aeronautical instrument company located in New York Metropolitan area is seeking an instrument development engineer who is an electronics and servomechanism

(Continued on page 52A)

Electronic Engineers

BENDIX RADIO DIVISION
Baltimore, Maryland
manufacturer of
RADIO AND RADAR EQUIPMENT
requires:

PROJECT ENGINEERS

Five or more years experience in the design and development, for production, of major components in radio and radar equipment.

ASSISTANT PROJECT ENGINEERS

Two or more years experience in the development, for production, of components in radio and radar equipment. Capable of designing components under supervision of project engineer.

Well equipped laboratories in modern radio plant . . . Excellent opportunity . . . advancement on individual merit.

Baltimore Has Adequate Housing

Arrangements will be made to contact personally all applicants who submit satisfactory resumes. Send resume to Mr. John Siena:

BENDIX RADIO DIVISION
BENDIX AVIATION CORPORATION
Baltimore 4, Maryland

WANTED: CAREER MEN IN ELECTRONICS at RCA VICTOR

Camden, New Jersey

● Unlimited laboratory resources and facilities are waiting for top-flight men ready to assume responsibilities in handling and administering advanced projects in virtually every phase of electronics—infra-red, ultrasonic, audio and acoustic equipment; television receivers, antennas, radar, mobile communications; aviation communications and navigational aids in our Engineering Products and Home Instruments Departments at Camden, New Jersey.

These Openings Represent a permanent expansion in RCA Victor design and development activities at Camden, providing careers for men of high calibre with appropriate training and experience.

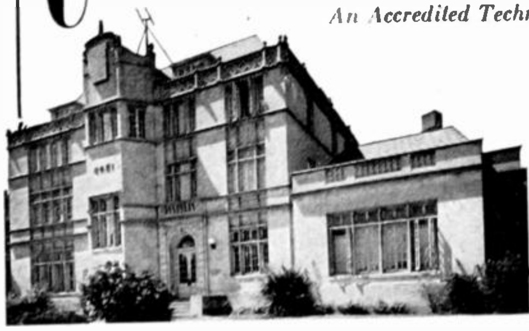
If You meet these Specifications, and if you are looking for a career which will open wide the door to the complete expression of your talents in the fields of electronics, write, giving full details, to:

ARNOLD K. WEBER,
Personnel Manager, Camden Plant
Box 133, RCA Victor Division
Radio Corporation of America
Camden, New Jersey

Pioneer in Radio Engineering Instruction Since 1927

CAPITOL RADIO ENGINEERING INSTITUTE

An Accredited Technical Institute



ADVANCED HOME STUDY AND RESIDENCE COURSES IN PRACTICAL RADIO-ELECTRONICS AND TELEVISION ENGINEERING

Request your free home study or residence school catalog by writing to:

REGISTRAR
16th and PARK ROAD N.W.
WASHINGTON 10, D.C.

Approved for Veteran Training

Radio and Radar Development and Design Engineers

Openings for experienced men at

HAZELTINE ELECTRONICS CORPORATION

Little Neck, L.I., N.Y.

Please furnish complete resume of experience with salary expected to:
Director of Engineering Personnel

(All inquiries treated confidentially)

POSITIONS OPEN

(Continued from page 51A)

specialist to become Chief Engineer. Background must be well balanced in all phases of aeronautical instrument business, including laboratory and shop practice from first hand experience. Interesting salary and bonus. Box 565.

LABORATORY TECHNICIAN

Excellent opportunity for experienced electronics laboratory technician having experience in aircraft instruments and allied fields. Location, Metropolitan New York. Experience and ability in all phases of laboratory testing and shop work desired. Salary commensurate with ability. Box 566.

COMMUNICATION ENGINEERS

Graduate Communication Engineers, Canadian citizens, interested in audio, radio and video frequency systems engineering. Salaries up to $350. per month depending on qualifications. Location, Montreal, Canada. Box 567.

TELEVISION ENGINEERS

Due to wide expansion in television program of long established radio and television manufacturing company, positions are now available for engineers with 3 or more years television experience. Excellent working conditions in modern, well-equipped laboratory. Company located in northwest portion of New York State. Housing no serious problem. Salaries commensurate with ability. Box 568.

SCIENTISTS AND ENGINEERS

Wanted for interesting and professionally challenging research and advanced development in the fields of microwaves, radar, gyroscopes, servomechanisms, instrumentation, computers and general electronics. Scientific or engineering degrees or extensive technical experience required. Salary commensurate with experience and ability. Direct inquiries to Mgr., Engineering Personnel, Bell Aircraft Corporation, P.O. Box 1, Buffalo 5, N.Y.

ENGINEERS

An unusual opportunity exists for qualified engineers to participate in the growth of the Industrial Television field. Experience in video, pulse, electronic display and circuit techniques with good educational background required. Openings range from Junior to Project Engineers. Write to: Leonard Mautner, Vice President, TELEVISION EQUIPMENT CORPORATION, 238 William Street, New York 7, N.Y.

NEEDED AT ONCE

DEVELOPMENT ENGINEER—at least 5 years experience in developing electronic equipment in the 100 to 1,000 megacycle range.

DESIGN ENGINEER—at least 5 years experience with electro-mechanical design problems where space and difficult service conditions are important.

Ideal laboratory conditions. 5 day week, paid vacation, paid medical and surgical policy, paid hospitalization policy and paid life insurance. Located in New Jersey. Box 570.

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.

RADIO ENGINEER

B.S.E.E., 1943, University of Illinois, Tau Beta Pi, Eta Kappa Nu. M.I.T. electronics, A.U.S. radar officer. Radio Engineer, P-5 Eighth Army, Japan, 1946-1948. Now in graduate study, Northwestern Technological Institute. Age 38. Box 228 W.

TECHNICAL WRITER-SALES ENGINEER

Technical writer, B.S.E.E., Columbia University. Experienced electronic and electrical equipment. 1 year as radio instructor, 3 years sales management experience plus business degree. Desires technical writing or sales engineering position in New York area, permanent connection. 29 years old, married, aggressive, best references. Call Lu-8-9164 mornings or write Box 230 W.

ENGINEER

B.S.E.E., June 1949, Kansas State College. Age 28. Married, no children. 3 years Army Signal Corps, Press Wireless School. 2 years installation of communications equipment. 5 years experience in broadcasting. 1st class telephone license. Eta Kappa Nu. Desires position in television station. Box 231 W.

TEACHING-PHYSICIST

Harvard Ph.D. expected June 1949 or a little later in theoretical physics. Some electronics background. Desires academic position with research possibilities. Box 234 W.

TELEVISION ENGINEER

Graduated American Television Institute of Technology October 1948 with B.S.T.E. Age. 31. Married. 1st class F.C.C. license. 4 years maintenance. Army radar equipment. Have been teaching advanced television for 15 months. Desires position as development or T.V. station engineer. Have a thorough understanding of R.C.A. and DuMont television equipment. Box 235 W.

INDUSTRIAL ELECTRONICS

B.S., B.E., M.E. June 1949, Yale. Age 24. Married. Desires job in industrial electronics. 2 years industrial engineering experience. Box 236 W.

(Continued on page 54A)

Positions Wanted

(Continued from page 53A)

COMMUNICATIONS ADMINISTRATOR

Communications and radio officer with experience in large size ships of the Fleet available in southwestern United States. 8 years experience Naval Communications Service, 12 years utility company, thoroughly conversant with communications problems.

Age 43. Married. Own home. College training. Hold FCC license. Would consider airline, railway, transportation and oil companies, also sales with established company. Brochure available upon request. Box 237 W.

ELECTRONIC ENGINEER

B.E.E. Age 35. 2 years project engineering. Army. Navy project experience, 2 years Navy electronics, 3 years general engineering experience. Box 238 W.

ELECTRICAL ENGINEER

Graduate electrical engineer. Industrial experience with magnetic and electronic control. Army experience, radar repair and maintenance. Desires servomechanisms research and design work. Member of Tau Beta Pi, Eta Kappa Nu. Box 239 W.

ASSISTANT PROFESSOR OR RESEARCH ENGINEER

E.E. degree. Lt. USNR, graduate work at Naval Academy. P.G. School Electronics Engineering. 2 years E.E. teaching all subjects. Desires position as assistant professor or research engineering. Middlewest or south area. All offers carefully considered. Box 240 W.

(Continued on page 55A)

Positions Wanted

(Continued from page 54A)

ELECTRONIC ENGINEER

B.S.E.E., University of Texas, August 1948, communications major. 2 years experience as Navy electronics technician, partly with guided missles; 2 years commercial radio servicing; 6 months experience in microwave design. Further details on request. Prefer position in southwest. Box 241 W.

ELECTRONIC ENGINEER

B.S. radio and electronic engineering, June 1949. Married, 3 children. Age 24. Air Force communications and radar maintenance officer. 1st class phone license. Prefer west coast. Box 242 W.

ENGINEER

University of Illinois graduate, B.S., M.S., in electrical engineering. 1 year Research Assistant at the University, and 6 months as a Research Engineer for a private company. Desires a position as a development and application engineer in the field of electronic controls. Box 249 W.

JUNIOR ENGINEER-STATISTICIAN

B.E.E. Summa cum Laude; 3 years experience as statistician, 1 year work research and development electronics field. Desires suitable position communications, electronics, or public utility. Age 30, married, no children. Prefer New York City but will consider position anywhere in United States. Box 250 W.

ENGINEER

B.S. in E.E., M.S. in physics. 6 years research and development experience in electronics and underwater sound. Age 27. Married. Desires research position. Prefer east or mid-west. Box 251 W.

TELEVISION ENGINEER

Available July, experienced television engineer, desires new station construction or small manufacturing plant. University of Pennsylvania graduate, good business man, hard worker. Prefer east. Box 252 W.

ENGINEER

B.S.T.E. Age 32. Married. Desires position in Chicago area, associated with engineer of experience in television, radio or U.H.F. Trained in all phases of television engineering. Experienced in servicing radio, television, record changers and F.M. Box 253 W.

TRANSFORMER ENGINEER

B.E.E. Age 27. Married. In Transformer field for 5 years. Last 3 years in design of power components. Desires position in same or allied fields. Box 254 W.

PROJECT ENGINEER

B.E.E. 1944 Cornell, M.E.E., 1948 Polytechnic Institute of Brooklyn. H.K.N., T.B.P., Sigma Xi at Cornell. Experience in pulse circuits, wide-band I.F. amplifiers, microwaves on microwave P.T.M. relay system, microwave signal generators. F.C.C. 1st class license. Aviation Radio-Radar officer U.S. Navy. Desires position in television or allied field. Box 255 W.

ENGINEER

B.E.E. George Washington University, June 1949. Communications major. 7 years experience long distance radio. 3 years hi-speed radio-telegraph operator, 4 years supervisory capacity. Married. Age 28. Energetic, adaptable, personable. Box 256 W.

(Continued on page 56A)

Positions Wanted

(Continued from page 55A)

SERVO ENGINEER

Age 31. Married. B.S. in E.E., 1946. 1 year audio experience. M.S. in E.E. 1949 with emphasis on servomechanisms. Desires position in industrial servo design. Box 257 W.

RADIO-ELECTRONICS ENGINEER

B.S. Engineering Sciences Harvard 1943. Age 28. Married, no children. 6 years broad experience in electronics including radar and airborne magnetometer. At present Government P-3 radio engineer. Desires job in France, wife to accompany. Speaks French and Spanish fluently and a little German and Chinese. Both commercial radio licenses. Teacher of High School Physics. Box 258 W.

ENGINEER

E.E. student, graduate June 1950. Desires position summer June 1949. Location immaterial. Have 3½ years experience as ETM in radar, sonar, loran, and other shipboard equipment. Some experience with particle accelerators, hi-vacuum gear. Also interested in ultrasonics. Box 259 W.

ELECTRICAL ENGINEER

M.S.E.E. June 1949, The Rice Institute, communications major, minor in math. and physics. Age 25, single. 1½ years experience various geophysical equipment. Desires research or development position in Los Angeles area. Work need not be related to geophysics. Member Tau Beta Pi and Sigma Xi. Box 260 W.

PATENT LAW CLERK

New York University June 1948 graduate with B.S.E.E. degree. Now completing 1st year evening Law School in preparation for LL.B. degree September 1950. Desires full time position with Patent Law firm in New York City metropolitan area only. Presently employed as a technical writer on electronic and electric equipment. Resume on request. Box 261 W.

ENGINEER

B.S.E.E., communications, May 1949, Oklahoma A. and M. Institute of Technology. Age 29. 4 years training and experience in Naval electronics, radar, maintenance and repairs, advanced to A.C.R.T. Desires position in electronic engineering, anywhere in U.S. References. Box 262 W.

ELECTRONIC ENGINEER

Currently engaged in production, design and development work, desires position in New York area. Have B.E.E., am now studying for M.E.E. evenings. 2 years experience board layout and testing. I year with transformer manufacturer and present employer. Age 26. Married. Box 263 W.

RADIO ENGINEER

Graduating in Radio Engineering June 1949. B.S. Age 25, married. 3½ years Army experience, fixed station installation and operation, antenna and transmission line construction. Also civilian broadcast experience. Prefer antenna or transmission line design and development. Box 264 W.

News—New Products

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

All-Triode Amplifier

An all-triode laboratory amplifier and dynamic noise suppressor, type 211-A, is now available from the designer, **Hermon Hosmer Scott, Inc.,** 385 Putnam Ave., Cambridge 39, Mass.

The 211-A has a frequency range of 20 to 20,000 cps, and a peak power output of 14 watts. This rather high output from a triode amplifier, the manufacturer claims, is attained by a special stabilized grid-bias circuit.

Distortion is less than ½% at 5 watts, and 2% at 12 watts. Hum is approximately 80 db below maximum power output at normal volume. Triodes are used in all amplifying stages, and low-mu triodes are featured in the high-level stages.

Four-Frequency FM Monitor

A new monitor, the FMM-1, which measures carrier frequency, carrier intensity, and modulation index at a considerable distance from a fixed or mobile station, is now in production by **West Coast Electronics Co.,** 1601 S. Burlington Ave., Los Angeles 6, Calif.

The specifications of the FMM-1 are: range, 30 to 20 Mc., any four frequencies within the band; rf input impedance, 52 ohms; a connector for RG-8/U coaxial cable is provided; rf sensitivity, 200 microvolts or better at all frequencies within the band; audio output, 500 milliwatts at 15 kc deviation; output impedance, 500 ohms; signal-to-noise ratio, 40 db at 200 microvolts input, 15 kc deviation at 400 cps; distortion is 2% or less; power supply, 105 to 125 volts, 50 to 60 cps.

Descriptive literature and price quotations available on request.

(Continued on page 58A)

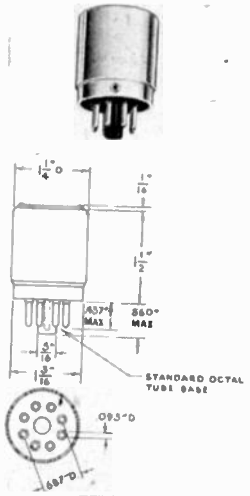
News—New Products

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

(Continued from page 57A)

Crystal-Temperature Stabilizer

A new crystal-temperature stabilizer, type TCO-1, which is a miniature crystal oven engineered for modern military and commercial communications equipment, by **Bliley Electric Co.**, 227 Union Station Bldg., Erie, Pa.



This oven is designed for use with Bliley type BH6 crystal units which mount in an internal socket. This combination will provide frequency stability down to $\pm 0.0001\%$ while crystal temperature is maintained within $\pm 2°C$.

The standard unit is supplied for operation at $75°C \pm 2°C$ and is equipped with a 6.3-volt heater rated at 5.5 watts. Supplied with type BH6 crystal units at any specified frequency in range 1 to 100 Mc, the TCO-1 oven introduces a new form factor in temperature-stabilized crystals for high precision.

Recent Catalogs

• • •Two bulletins: the first explaining the procedure of converting a four-terminal to a five- or six-terminal meter socket, the second, a 4-page catalog illustrating the basic line of No. 9000 electric meter sockets, with complete specifications, by **Anchor Mfg. Co.**, 30 Huntington Ave., Boston 16, Mass.

• • •A new brochure describing the complete line of products, especially the wide variety of oscilloscopes manufactured by **Furzehill Labs., Ltd.**, Boreham Wood, Herts., England. Contact should be made through **American British Technology, Inc.**, 57 Park Ave., New York 16, N. Y.

• • •A general catalog, the first to be released in some time, and therefore including a number of new items not previously mentioned, by **Meissner Mfg. Div., Maguire Industries, Inc.**, Mt. Carmel, Ill.

(Continued on page 60A)

News—New Products

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

(Continued from page 58A)

Hermetically Sealed DC Relay for High-Altitude Requirements

A hermetically sealed dc miniature relay, type CX3554, has been developed by **Struthers-Dunn, Inc., 150 N. 13 St., Philadelphia 7, Pa.**, to meet the extreme operating conditions of modern high-altitude and jet-propelled aircraft.

Special aviation design features include shock resistance up to 50 G's; vibration resistance better than 10 G's; high-speed opening and closing without contact bounce; and reliable operation over an ambient temperature range of −75°C to +200°C. A high degree of hermetic sealing makes the unit insensitive to humidity changes and capable of rated operation at altitudes as high as 70,000 feet.

The relay has SPDT contacts nominally rated at 2 amperes and capable of withstanding inrushes of 12 amperes at 26.5 volts dc. The coil is rated at 26.5 volts dc with an operating range of 18/32 volts dc. Of cylindrical shape with mounting "feet," the relay is approximately 1¾" high including terminals, with a flange diameter of 1.040", and 1⅜" between mounting hole centers. Complete details will gladly be sent upon request to the manufacturer

Recent Catalogs

• • • A bulletin on "Socket and Mounting Notes for Raytheon Flat Press Subminiature Tubes," which also explains other methods of connecting to the tube, shielding it, and potting it in plastic, by **Raytheon Mfg. Co., Inc.,** 55 Chapel St., Newton 58, Mass.

• • • A new television components folder No. P-1, illustrated with photographs and schematic diagrams, in which 19 television parts are analyzed according to function, general use considerations, ratings, and connections, by **Transvision, Inc.,** 460 North Ave., New Rochelle, N. Y.

MICROWAVE PLUMBING

10 CENTIMETER

WAVEGUIDE DIRECTIONAL COUPLER. 27 db. Navy type CABV-47AAN, with 4 in. slotted section as shown $42.50
SQ. FLANGE to rd choke adapter, 18 in. long OA 1½ in. x 3 in. guide, type "N" output and sampling probe $32.00
"S" BAND CRYSTAL MOUNT, gold plated, with 2 type "N" connectors $12.50
POWER SPLITTER: 726 Klystron input, dual "N" output $5.00
MAGNETRON TO WAVEGUIDE coupler with 721-A duplexer cavity, gold plated $27.50
10 CM WAVEGUIDE SWITCHING UNIT, switches 1 input to any of 3 outputs. Standard 1½" x 3" guide with square flanges. Complete with 115 vac or dc arranged switching motor. Mfg. Raytheon, CRP 24AAS. New and complete $150.00
10 CM END-FIRE ARRAY POLYRODS ....$1.75 ea.
"S" BAND Mixer Assembly, with crystal mount, pick up loop, tunable output $3.00
721-A TR CAVITY WITH TUBE. Complete with tuning plungers $12.50
10 CM. McNALLY CAVITY Type SG $3.50
WAVEGUIDE SECTION, MC 445A, rt. angle bend., 5¼" ft. OA. 8" slotted section $21.00
10 CM OSC. PICKUP LOOP, with male Homedell output $2.00
10 CM DIPOLE WITH REFLECTOR in lucite ball, with type "N" or Sperry fitting $4.50
10 CM FEEDBACK DIPOLE ANTENNA, in lucite ball, for use with parabola $8.00

⅞" RIGID COAX—⅝" I.C.

RIGHT ANGLE BEND, with flexible coax output pick-up loop $8.00
SHORT RIGHT ANGLE bend, with pressurizing nipple $3.00
RIGID COAX to flex coax connector $3.50
STUB-SUPPORTED RIGID COAX, gold plated 5' lengths. Per length $5.00
RT. ANGLES for above $2.50
RT. ANGLE BEND 15" L. OA $3.50
FLEXIBLE SECTION, 15" L. Male to female $4.25
MAGNETRON COUPLING to ⅞" rigid coax, with TR pickup loop, gold plated $7.50

⅞" RIGID COAX—¼" I.C.

⅞" RIGID COAX, Bead Supported ......$1.20 per ft.
SHORT RIGHT ANGLE BEND $2.50
Rotating joint, with deck mounting $15.00
RIGID COAX slotted section CU-60/AP $5.00

1.25 CENTIMETER

"K" BAND FEEDBACK TO PARABOLA HORN, with pressurized window $30.00
MITRED ELBOW cover to cover $4.00
TR/ATR SECTION choke to cover $4.00
FLEXIBLE SECTION 1" choke to choke $5.00
ADAPTER, rd. cover to sq. cover $5.00
MITRED ELBOW and S sections choke to cover..$4.50
PICKUP Loop, Type "N" output 10 cm $2.75
TR Box Pick Up Loop 10 cm. $1.25
WAVE GUIDE ½" x ¼" per ft. $1.00

****

3 CENTIMETER PLUMBING

(STD. 1" x ½" GUIDE, UNLESS SPECIFIED)

"X" BAND PREAMPLIFIER, consisting of 2-723 A B local oscillator-beacon feeding waveguide and TR/ATR Duplexer sect. incl. 60 mc IF amp $47.50
Random Lengths waveg. 6" to 18" Lg. ...$1.10 Ft.
WAVEGUIDE RUN 1⅛" x ½" guide, consisting of 4 ft. section with Rt. angle bend on one end 2" 45 deg. bend other end $8.00
WAVEGUIDE SECTION 1⅛" x ½" choke to choke 4 ft. long $10.00
Dummy Load. Ts 332/UP $22.50
"X" BAND PRESSURIZING gauge section w/15-lbs gauge Section w/15-lbs gauge & Pressurizing Nipple $18.50
45 DEG. TWIST 6" Long $10.00
12 SECTION 45 deg twist 90 deg. bend $6.00
11 STRAIGHT WAVEGUIDE section choke to cover $4.50
15 DEG BEND 10" choke to cover $4.50
special heavy construction Silver plated $14.50
5 FT SECTIONS choke to cover $17.50
18" FLEXIBLE SECTION $12.50
"F" and "H" PLANE BENDS $15.00
BULKHEAD FEED THRU $15.00
"X" BAND WAVEGUIDE 1¼" x ⅝" OD 1/16" wall Aluminum ....Per Foot $ .75
WAVEGUIDE 1" x ½" I.D. ......Per Foot $1.50
TR CAVITY For 724-A TR Tube $3.50
3" FLEX SECT. sq. flange to Circ Flang Adapt. $7.50
724 TR TUBE (41-TR-1) $2.50
TR/ATR DUPLEXER Sect. w/Iris flange $8.00
Twist 90 deg. 5" choke to Cover w/press nipple $6.50
Waveguide Sections 2½ ft. long silver plated with choke flange $5.75
WAVEGUIDE 90 deg. bend "E" Plane, 18" Lg $4.00
Rotary Joint choke to choke $17.50
Rotary Joint, choke to choke with deck mounting $17.50
S. CURVE WAVEGUIDE 8" Lg. cover to choke $3.50
DUPLEXER SECTIONS FOR 1B24 ....$10.00
CIRCULAR CHOKE FLANGES solid brass ......$.55
SQ. FLANGES FLAT BRASS ....ea. $.55
APS-10 TR/ATR DUPLEXER section with additional IRIS Flange $10.00
CU 105/APS 31 Directional coupler 25 lb. $15.00
CU 103/APS 33 Directional coupler 25 lb. $15.00
FLEXIBLE WAVEGUIDE ....$4.00/FT.

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MICROWAVE GENERATORS

AN/APS-15A "X" Band compl. RF head and modulator, incl. 725-A magnetron and magnet, two 723A/B klystrons (local osc. & beacon), 1B24 TR, rcvr-ampl, duplexer, HV supply, blower, pulse xtmr, Peak Pwr. Out: 45 KW apx. Input: 115, 400 cy. Modulator pulse duration .5 to 2 micro-sec. apx. 13 KV Pk Pulse. Compl. with all tubes incl. 715-B, 829B, RKR 73, two 72's. Compl. pkg., new $210.00
APS-15B. Complete pkg. as above, less modulator $150.00

"S" BAND AN/APS-2. Complete RF head and modulator, including magnetron and magnet, 417-A mixer, TR, receiver, duplexer, blower, etc., and complete pulser. With tubes, used, fair condition $75.00
10 CM, RF Package. Consists of: SO Xmtr.-receiver using 2127 magnetron oscillator, 250 KW peak input, 707-B receiver-mixer $150.00
Modulator-motor-alternator unit for above.....$ 75.00
Receiver-rectifier power unit for above........$ 25.00
Rotating Ant. with parabolic reflector for above. New ....$ 75.00

MAGNETRONS

TUBE	FREQ. RANGE	PK. PWR. OUT.	PRICE
2J31	2820-2880 mc.	285 KW.	$25.00
2J21-A	9345-9405 mc.	50 KW.	$25.00
2J22	3267-3333 mc.	265 KW.	$25.00
2J26	2992-3019 mc.	275 KW.	$25.00
2J27	2965-2992 mc.	275 KW.	$25.00
2J32	2780-2820 mc.	285 KW.	$25.00
2J37			$45.00
2J38 Pkg.	3249-3263 mc.	5 KW.	$35.00
2J39 Pkg.	3267-3333 mc.	87 KW.	$35.00
2J40	9305-9325 mc.	10 KW.	$65.00
2J49	9000-9180 mc.	58 KW.	$85.00
2J55 Pkg.	9345-9405 mc.	50 KW.	$35.00
2J61	3000-3100 mc.	35 KW.	$65.00
2J62	2914-3010 mc.	35 KW.	$65.00
3J31	24,000 mc.	50 KW.	$55.00
5J30			$39.50
714AY			$25.00
718DY			$25.00
720BY	2800 mc.	1000 KW.	$50.00
720CY			$50.00
725-A	9345-9405 mc.	50 KW.	$25.00
730-A	9345-9405 mc.	50 KW.	$25.00

Klystrons: 723A/B $12.50; 707B W/Cavity $20.00
417A $25.00 2 K 41 $65.00
707A-w/cavity $18.00

MAGNETRON MAGNETS

GAUSS	POLE DIAM.	SPACING	PRICE
4850	¾ in.	⅝ in.	$12.50
5200	21/32 in.	¾ in.	$17.50
1500	1⅝ in.	1½ in.	$12.50
D161392*	1⅝ in.	1 5/16 in.	$12.50

* Mfr's Number.

TUNABLE PKGD. "CW" MAGNETRONS

QK 61 2975-3200 mc.   QK 62 3150-3375 mc.
QK 60 2800-3025 mc.   QK 59 2675-2900 mc.
New, Guaranteed ....Each $65.00

MICROWAVE TEST EQUIPMENT

W. E. I 138. Signal generator, 2700 to 2900 Mc range. Lighthouse tube oscillator with attenuator & output meter. 115 VAC input, reg. Pwr. supply. With circuit diagram $50.00
3 cm. Wavemeter: 9200 to 11,000 m transmission type with square flanges $15.00
3 cm. stabilizer cavity, transmission type $20.00
3 cm. Wavemeter. Micrometer head mounted on X-Band guide. Freq. range approx. 7900 to 10,000 Mc. $75.00
THERMISTOR BRIDGE: Power meter 1-203-A. 10 cm. mfg. W.E. Complete with meter, interpolation chart, portable carrying case as shown $72.50
TS 12/AP VSWR METER. Consisting of Slotted Line, and all additional fittings and ampl. for 3CM operation. New and complete $450.00
Bell Labs. Dual Mount mixer-beacon assemblies, 2 complete mixer-beacon mounts on gold-plated waveguide section $50.00
Slotted Line, Bell Labs. 1⅛" x 5/16" guide, gold plated $150.00

THERMISTORS

D-167332 (tube)	$.95
D-170396 (bead)	$.95
D-167613 (button)	$.95
D-186228 (button)	$.95
D-164699 for MTO, in "X" band Guide	$2.50
D-167018 (tube)	$.95

COAX PLUGS

831SP	$.35
831AP	$.35
831HP	$.15
UG 21/U	$.85
UG 86 U	$.95
UG 254/U	$.75
UG 103/U	$.55

VARISTORS

D-171631	$.95
D-167176	$.95
D-168687	$.95
D-171812	$.95
D-171528	$.95
D 168549	$.95
D-162482	$3.00
D-166277	$2.50
D-99428	$2.00
D-16187A	$2.85
D-163075	$1.25
3A(12-43)	$1.50
D167020	$3.00

PULSE EQUIPMENT

MODULATOR UNIT BC 1203-B

Provides 200-4,000 PPS. Sweeptime: 100 to 2,500 microsec. in 4 steps, fixed mod. pulse, suppression pulse, sliding modulating pulse, blanking voltage, marker pulse voltage, calibration voltages, fil. voltages. Operates 115 vac. 50-60 cy. Provides various types of voltage pulse outputs for the modulation of a signal generator such as General Radio #804B or #804C used in depot bench testing of SCR 695, SCR 595, and SCR 535. New as shown $125.00
MIT. MOD. 3 HARD TUBE PULSER: Output Pulse Power: 144 KW (12 KV at 12 amp). Duty Ratio: .001 max. Pulse duration: 5, 1.0, 2.0 microsec. Input voltage: 115 v. 400 to 2400 cps. Uses 1-715-B, 1-829-B, 3-'72's, 1-'73. New $110.00
APQ-13 PULSE MODULATOR. Pulse Width .5 to 1.1 Micro Sec. Rep. rate 824 to 1348 Pps. Pk. pwr. out 35 KW. Energy 0.018 Joules $49.00
TPS-3 PULSE MODULATOR. Pk. power 50 amp. 24 KV (1200 KW pk); pulse rate 200 PPS, 1.5 microsec; pulse line impedance 50 ohms. Circuit—series charging version of DC Resonance type. Uses two 705-A's as rectifiers, 115 v. 400 cycle input. New with all tubes $49.50
APS-10 MODULATOR DECK. Complete, less tubes $75.00
APS-10 Low voltage power supply, less tubes ..$18.50

PULSE NETWORKS

G.E. #25E5-1-350-50P2T, 25 KV, 5 sections, "E" circuit, 1 microsecond pulse length, 350 PPS, 50 ohms impedance $45.00
G.E. #6E3-5-2000-50P2T, 6KV, "E" circuit, 3 sections, .5 microsecond, 2000 PPS, 50 ohms impedance ..$6.50
G.E. #3E (3-.84-810; 8-2.24-405) 50P4T: 3KV, "E" CKT Dual Unit: Unit 1, 3 Sections, .84 Microsec, 810 PPS, 50 ohms imp.; Unit 2, 8 Sections, 2.24 Microsec, 405 PPS, 50 ohms imp. $6.50
7.5E3-1-200-67P, 7.5 KV, "E" Circuit, 1 microsec, 200 PPS, 67 ohms impedance, 3 sections $7.50
7.5E4-16-60-67P, 7.5 KV, "E" circuit, 4 sections, 16 microsec, 60 PPS, 67 ohms impedance $15.00
7.5E3-3-200-6PT, 7.5 KV, "E" Circuit, 3 microsec, 200 PPS, 67 ohms imp., 3 sections $12.50

DELAY LINES

D-168184: .5 microsec. up to 2000 PPS, 1800 ohm term $4.00
D-170499: .25/.50/.75/. microsec. 8 KV, 50 ohms imp. $16.50
D-165997: 1¼ microsec. $7.50

PULSE TRANSFORMERS

G.E.K.-2745 ....$39.50
G.E.K.-2744-A $39.50
W.E. #D166173 Hi-Volt input transformer, W.E. impedance ratio 50 ohms to 900 ohms. Freq. range: 10 kc to 2 mc. 2 sections parallel connected, potted in oil $36.00
W.E. KS 9800 Input transformer. Winding ratio between terminals 3-5 and 1-2 is 1.1:1, and between terminals 6-7 and 1-2 is 2:1. Frequency range: 380-520 c.p.s. Permalloy core $6.00
G.E. #K2731 Repetition Rate: 835 PPS, Pri. Imp: 50 Ohms, Sec. Imp: 450 Ohms, Pulse Width: 1 Microsec, Pri. Input: 9.5 KV PK, Sec. Output: 28 KV PK, Peak Output: 800 KW, Bifilar 2.75 Amp. $64.50
W.E. #D169271 Hi Volt input pulse Transformer $27.50
G.E. K2450A Will receive 13KV, 4 micro-second pulse on pri., secondary delivers 14KV. Peak power out 100KW GE $45.00
G.E. #K2748A Pulse Input, line to magnetron ..$36.00
#9280 Utah Pulse or Blocking Oscillator XFMR Freq. limits 790-810 cy-3 windings turns ratio 1:1:1 Dimensions 1 13/16 x 1 1/8 x 19/32" $1.50

400 CYCLE TRANSFORMERS

352-7273: Pri: 115V. 400 cy. Sec: 6.3V. 2.5 Amp; 6.3V. .06 Amp; 6.3V. .9 Amp; 5V. 6 Amp; 700 VCT. 2-5U4's. For **APS-15. T201** $4.75
352-7176: Pri: 115V. 400 cy. Sec: 8.3V, 20 Amp; 6.3V. .5 Amp; 8.3V. .5 Amp; 320V (2-6x5's). For **APS-15. T202** $5.25
352-7278: Pri: 115V. 400 cy. Sec: 2.5V. 1.75 Amp. 3500V (2x2). For **APS-15. T203** (Anode #2 5FP7) $5.85
352-7070: Pri: 118V, 440 cy. Sec. 2.5V, 2.5 Amp; 2.5V, 2.5 Amp; (2000V, Ins.); 6.3V, 2.25 Amp; 1200V, Tpd. at 1000 and 750V. P/o **AN/APS 15** $4.95
#7469105: Pri: 115V. 400 cy. Sec: Tpd. to give 742.5V. 50 MA; 709V, .0477 A, 671V, .045 A. $2.95
M-7474319: Pri: 115V, 400 cy. Sec: 6.3V, 2.7 Amp; 6.3V, 66 Amp; 6.3V, .21 Amp. $2.95
32332: Pri: 115V, 400-2400 cy. Sec: 400 Vct, 35 MA; 8.4V, 2.5 Amp; 6.4V, .15 Amp. $2.25
332-7138M: Pri: 115V, 400-2400 cy. Sec: 640V, 5 MA; 2.5V, 1.75 Amp. $3.85
352-7179: Pri: 115V, 400-2400 cy. Sec: 6.5V, 12 Amp, Ct. 250V, 100 MA; 5V, 2 Amp. $3.50
#9069: Pri: 115/80V, 400-2600 cy. Sec: 850 Vct. 50 MA; 6.3 Vct. 2 Amp; 5 Vct. 2 Amp. $2.15
352-7096: Pri: 115/80V, 400-2400 cy. Sec: 2.5V, 1.75 Amp. 3 Kv. Ins.; 5V, 3 Amp; 6.5V, 6.5 Amp; 6.5V, 1.2 Amp. $3.95
KS 9607: Pri: 115V, 400-2400 cy. Sec: 734 Vct. 177 MA. 1710 Vct. 177 MA. $5.95
D-166333: Pri: 115V, 400-2400 cy. Sec: 6.3V, 0.9 Amp, 7.7V, 0.365 Amp. $2.79
GE #7471957: Pri: 100/110/120/130V, 400-2400 cy. Sec: 2.5V, 20 Amp, HV Ins. $4.85
D-163254: Pri: 115V, 400 cy. Sec: 6.3V, 12 Amp; 6.3V, 2A; 6.3V, 1A, **P/O AN/APQ5** $5.85
KS-9685: Pri: 115V, 400-2400 cy. Sec: 6.4Vct, 7.5 Amp; 6.4V, 3.8 Amp; 6.4V, 2.5 Amp. $4.35
using 2J27 magnetron oscillator, 250 KW peak input, 707-B receiver-mixer $150.00
Modulator-motor-alternator unit for above ....$ 75.00
Receiver-rectifier power unit for above ....$ 25.00

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