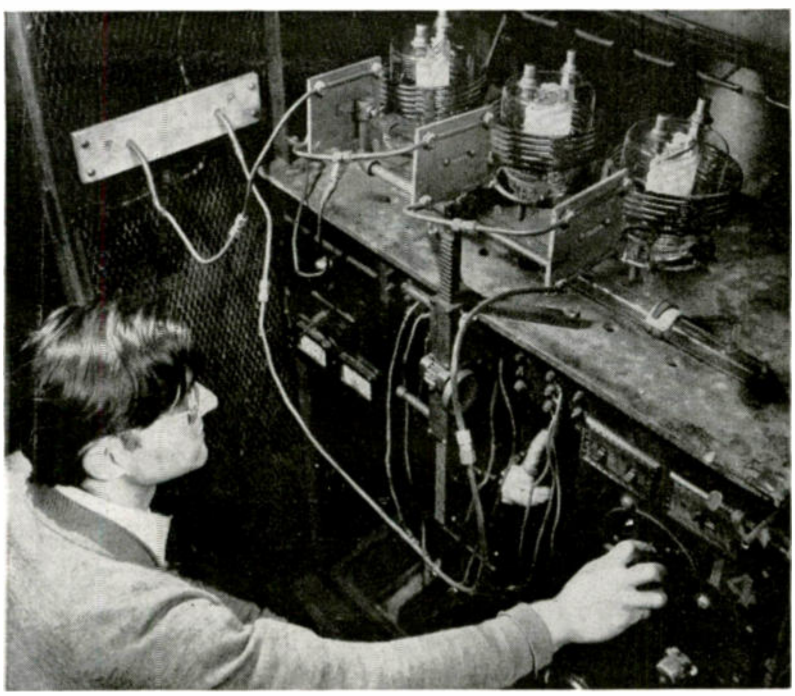
Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

(Including the WAVES AND ELECTRONS Section)



Amperex Electronic Corporation

RADIO-TUBE MANUFACTURE USES RADIO TECHNIQUES

After evacuation by an oil-diffusion pump, the internal metal parts in the tubes on the glass manifold are degassed by radio-frequency induction heating.

Convention Program and Summaries of Technical Papers in this issue.

March, 1948

Volume 36 Number 3

PROCEEDINGS OF THE I.R.E.

Waves and Electrons Section

Table of Contents Follows Page 32A

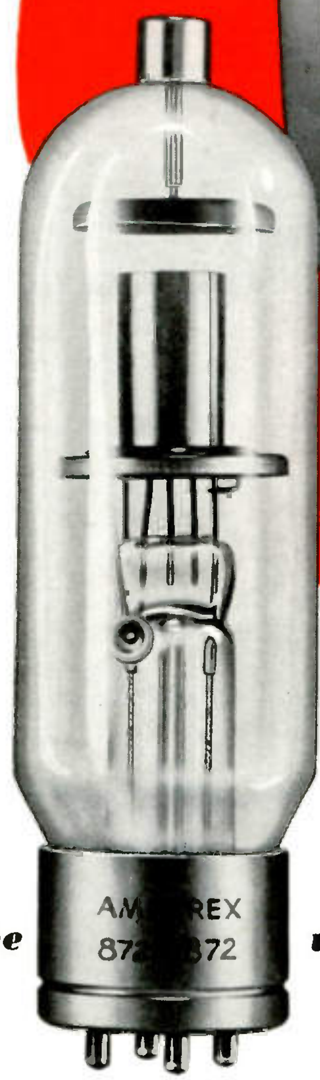
The Institute of Radio Engineers

"I like Amperex tubes because"...

It takes a lot more than 25 words to sum up all the reasons engineers prefer and specify Amperex tubes. For example: the engineers of Induction Heating Corporation specify Amperex 833-A power tubes and Amperex 872-A rectifiers for their Model 43 induction heater because they find that Amperex

The Model 43 Ther-Monic Induction Heater manufactured by Induction Heating Corporation is factory equipped with ***Amperex*** *833-A power tubes and* ***Amperex*** *872-A rectifier tubes.*

tubes have longer life, give a minimum of trouble and help produce satisfied customers. Too, they like that extra engineering that goes with the Amperex name; those little differences that make a big difference . . . and they also like the application engineering of the Amperex staff which is theirs, and yours, to command.



re-tube with Amperex

AMPEREX ELECTRONIC CORPORATION

25 WASHINGTON STREET, BROOKLYN 1, N. Y.
In Canada and Newfoundland: Rogers Majestic Limited
11-19 Brentcliffe Road, Leaside, Toronto, Ontario, Canada

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Table of contents will be found following page 32A

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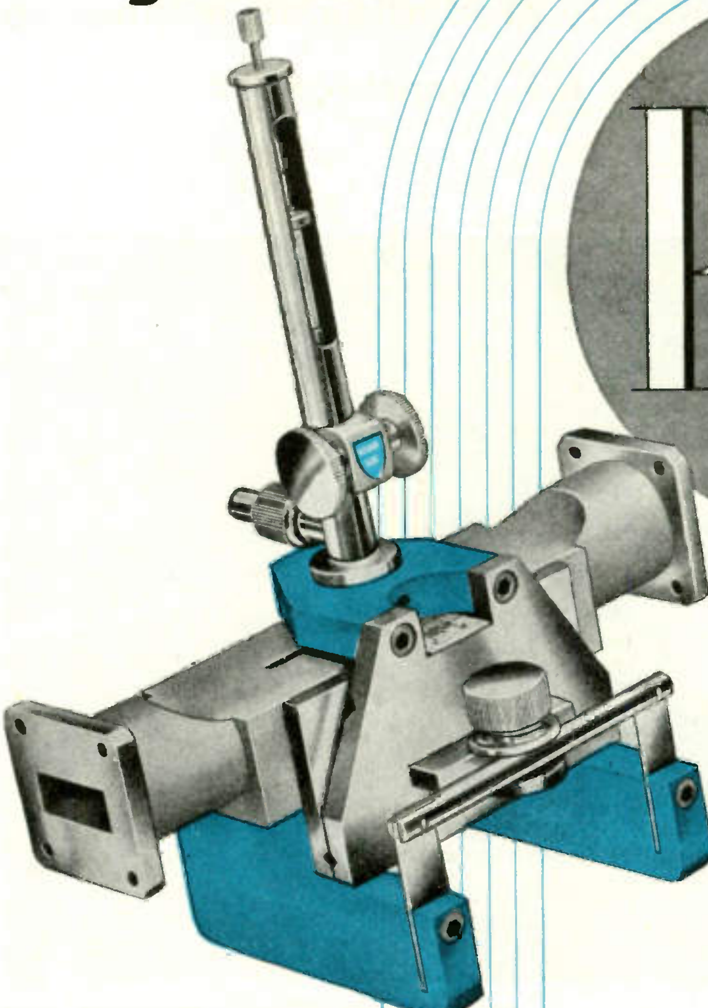
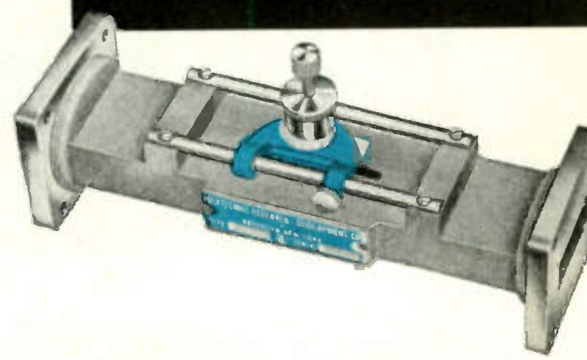


MICROWAVE MEASUREMENT COMPONENTS



Our engineering staff stands ready to assist you with your microwave problems. Adequate shop facilities are available for the fabrication of special systems or measurement components to meet your specific requirements.

Our Booth at the 1948 IRE SHOW is #276

PRD Microwave Measurement Instruments and Components represent the scientific mastery and creative design of a staff of over 20 engineers and physicists, many of whom have worked in this field since its first application in practice. Ingeniously and precisely fabricated to meet the exacting requirements of the art, these instruments have become the accepted standards in the foremost government, industrial and educational laboratories.

The PRD Microwave line now includes instruments in all wave-guide and coaxial line sizes covering the frequency range from 2600 to 26,000 megacycles per second. Available equipment includes such items as Resistive Pads and Attenuators, Slotted Sections and Probes, Frequency Meters, and impedance matching devices. An illustrated catalog may be obtained by writing Dept. R3 on company letterhead.

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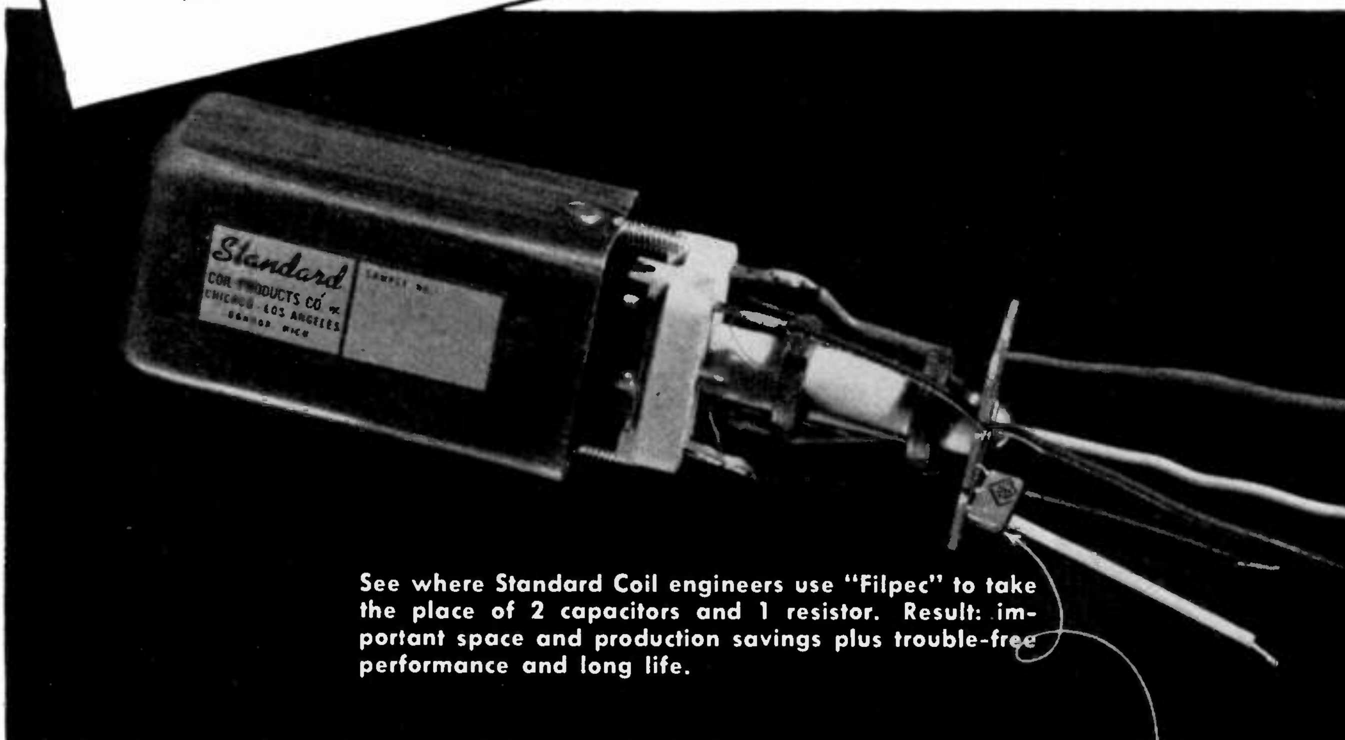
Polytechnic RESEARCH & DEVELOPMENT COMPANY, Inc.

PROGRESS REPORT ON P.E.C.*

How Standard Coil uses Centralab "Filpec" to simplify production, save space on I. F. Transformers

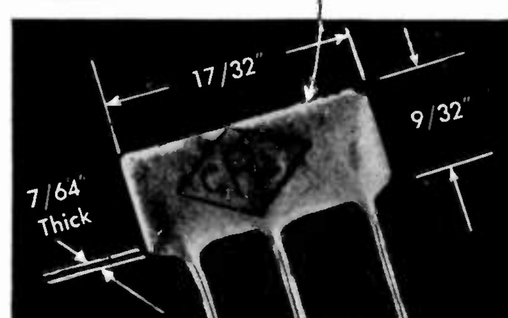


I. F. Transformer courtesy of Standard Coil Products Co.

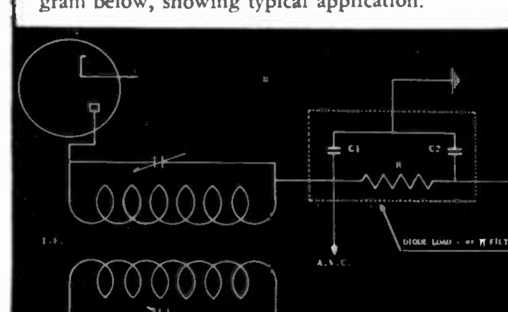
*Centralab's revolutionary Printed Electronic Circuit — Industry's newest method for stepping-up manufacturing efficiency!

FOR SMALL SIZE, light weight and long life, there's nothing like Centralab's new *printed electronic circuit filter!* That's why the Standard Coil Products Co. uses "*Filpec*" in its new I. F. Transformers. And that's why you'll want to see how it gives you higher circuit efficiency, more dependable performance as well as a reduction of line operations in set and equipment manufacturing.

Filpec combines two capacitors and one resistor into one tiny balanced diode load filter unit, saves space, cuts inventory, is highly adaptable to a variety of circuits. Capacitor values from 50 to 200 mmf. Resistor values from 5 ohms to 5 megohms. Resistance rating: 1/5 watt. 100 WVDC. Flash test: 200 VDC. For complete information about *Filpec* performance, see your Centralab representative, or write for Bulletin 976.



"Filpec" gives you integral construction! Made with high dielectric Ceramic-X, CRL's *Filpec* assures long life, low internal inductance, resistance to humidity and vibration. Note schematic diagram below, showing typical application.



LOOK TO Centralab IN 1948!

Division of GLOBE-UNION INC., Milwaukee

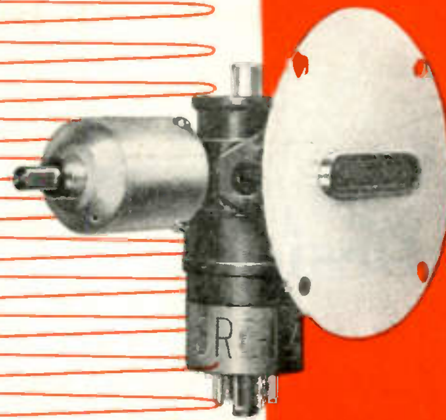
new klystron tubes

NOW AVAILABLE FOR MICROWAVE RELAYS

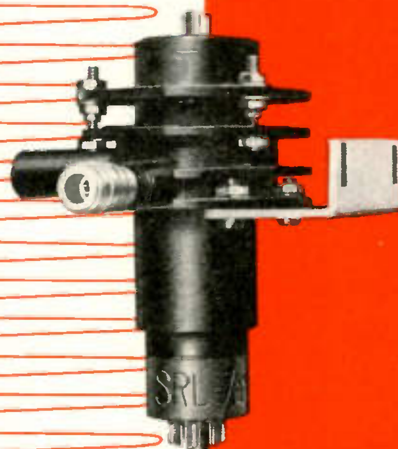
■ Four Sperry Reflex Klystron oscillators for microwave relay systems are now available for commercial use. These Klystrons can be used either as transmitting types or local oscillators. They can also be used in the laboratory as bench oscillators in the development of microwave relay systems.

■ With these new Klystron tubes, relay techniques are simplified and the mechanical problems associated with lower frequency relay links are overcome.

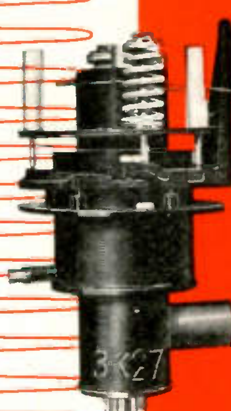
■ Other Sperry Klystrons are available in the frequency range from 500 to 12,000 megacycles. Our Industrial Department will gladly supply further information.



TYPE SRC-12,-20,-21
FREQUENCY 4400-5000 mc
POWER OUTPUT 5 WATTS MAX.



TYPE SRL-7a
FREQUENCY 1825-2100 mc
POWER OUTPUT 5 WATTS MAX.



TYPE 3K27
FREQUENCY 750-960 mc
POWER OUTPUT 1.5 WATTS MAX.

TYPE SRC-8 SERIES
FREQUENCY 5500-7800 mc*
POWER OUTPUT 4.5 WATTS MAX.

*The SRC-8 tubes are available in 100 megacycle steps except for 3 models, SRC-8A, SRC-8B, SRC-8C which are bench oscillators in 400 megacycle steps from 5850 to 7050

Sperry Gyroscope Company

DIVISION OF THE SPERRY CORPORATION · GREAT NECK, NEW YORK

NEW YORK · CLEVELAND · NEW ORLEANS · LOS ANGELES · SAN FRANCISCO · SEATTLE

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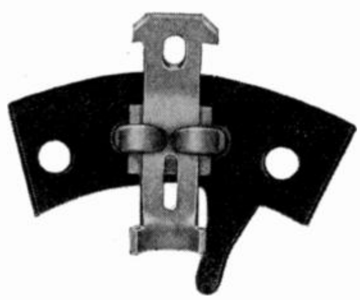
Compact Design... Unlimited Circuits...

...plus terminals that really stay put!

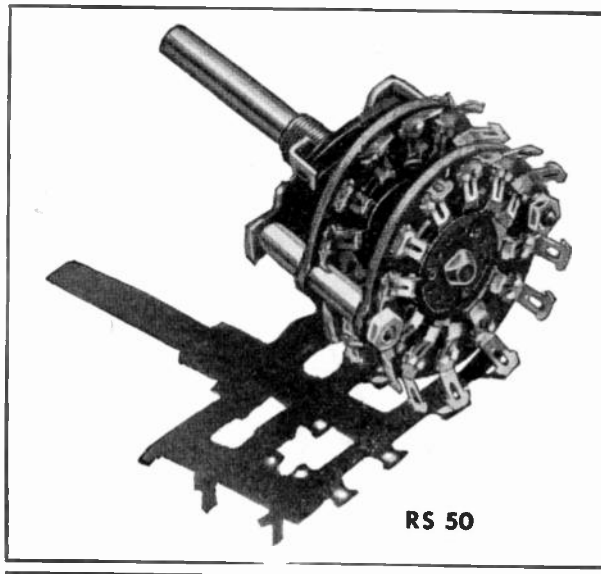


THEY'RE small, they're flexible, they're ruggedly designed. That's the story of the RS 50 and RS 60—two Mallory switches especially designed for radio receiver applications where low torque indexing is required.

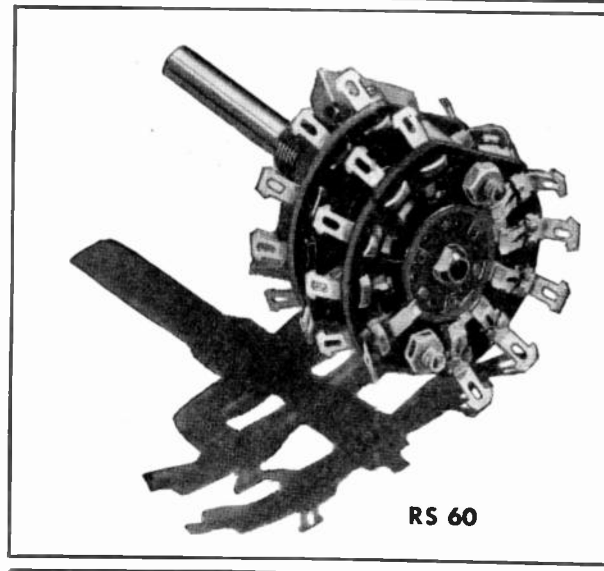
An outstanding feature of these switches is the two-point stapling which assures that terminals won't work loose. The terminals themselves are made of heavy spring brass for strength, silver plated, formed for flexibility, insuring low contact resistance.

Many other features are notable too: the improved low-loss phenolic in stator and rotor . . . the star wheel ball indexing with 30° between positions . . . silver-to-silver double wiping contacts . . . where desired the exclusive Mallory silver-indium treatment may be applied to rotor segments permitting higher contact pressure with lower, smooth operating torque and a minimum of contact resistance with extremely low noise level and long life.

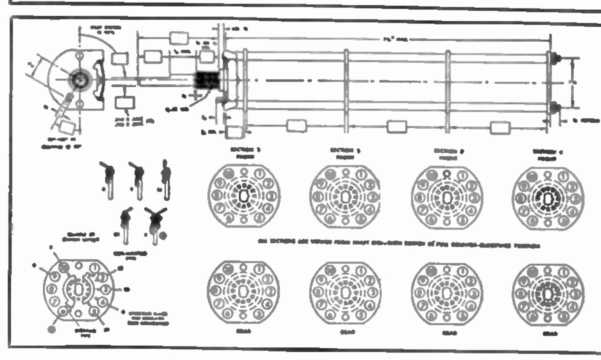
The RS 50 is made with from 2 to 10 positions—the RS 60 with from 2 to 5. For more details, write for engineering data folder.



RS 50



RS 60



Ask for RS Specification Sheets

Printed on thin paper to permit blueprinting, these sectional drawings indicate standard and optional dimensions—make it easy for you to specify Mallory RS switches built to meet your circuit requirements. Ask your nearest Mallory Field Representative or write direct for a supply.

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(ELECTRONIC, INDUSTRIAL and APPLIANCE)

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MEASURE TOTAL DISTORTION
Between 20 cps and 20 kc

-hp- 330B DISTORTION ANALYZER

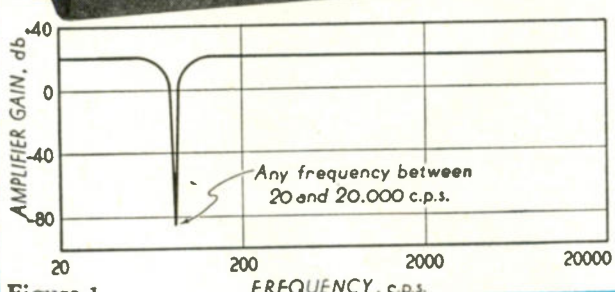

Figure 1

CHECK THESE SEVEN IMPORTANT FUNCTIONS:

1. **Measures total audio distortion.**
2. **Checks distortion of modulated r-f carrier.**
3. **Determines voltage level, power output.**
4. **Measures amplifier gain and response.**
5. **Directly measures audio noise and hum.**
6. **Determines unknown audio frequencies.**
7. **Serves as high-gain, wide-band stabilized amplifier.**

This fast, versatile *-hp-* 330B Analyzer measures distortion at *any* frequency from 20 cps to 20 kc. Measurements are made by eliminating the fundamental and comparing the ratio of the original wave with the total of remaining harmonic components. This comparison is made with a built-in vacuum tube voltmeter.

The unique *-hp-* resistance-tuned circuit used in this instrument is adapted from the famous *-hp-* 200 series oscillators. It provides almost infinite attenuation at one chosen frequency. All other frequencies are passed at the normal 20 db gain of the amplifier. Figure 1 shows how attenuation of approximately 80 db is achieved at any pre-selected point between 20 cps and 20 kc. Rejection is so sharp that second and higher harmonics are attenuated less than 10%.

Full-Fledged Voltmeter

As a high-impedance, wide-range, high-sensitivity vacuum tube voltmeter, this *-hp-* 330B gives precision response flat at any frequency from 10 cps to 100 kc. Nine full-scale ranges are provided: .03, .1, .3, 1.0, 3.0, 10, 30, 100 and 300. Calibration from +2 to −12 db is provided, and ranges are related in 10 db steps.

The amplifier of the instrument can be used in cascade with the vacuum tube voltmeter to increase its sensitivity 100 times for noise and hum measurements.

Accuracy throughout is approximately ±3% and is unaffected by changing of tubes or line voltage variations. Output of the voltmeter has terminals for connection to an oscilloscope, to permit visual presentation of wave under measurement.

Measures Direct From R-F Carrier

The *-hp-* 330B incorporates a linear r-f detector to rectify the transmitted carrier, and input circuits are continuously variable from 500 kc to 60 mc in 6 bands.

Ease of operation, universal applicability, great stability and light weight of this unique *-hp-* 330B Analyzer make it ideal for almost any audio measurement in laboratory, broadcast or production line work. Full details are immediately available. Write or wire for them—today Hewlett-Packard Company, 1437D Page Mill Road, Palo Alto, Calif.

hp laboratory instruments
FOR SPEED AND ACCURACY

Noise and Distortion Analyzers — Wave Analyzers — Frequency Meters
Audio Frequency Oscillators — Audio Signal Generators — Vacuum Tube Voltmeters
Amplifiers — Power Supplies — UHF Signal Generators — Attenuators
Square Wave Generators — Frequency Standards — Electronic Tachometers

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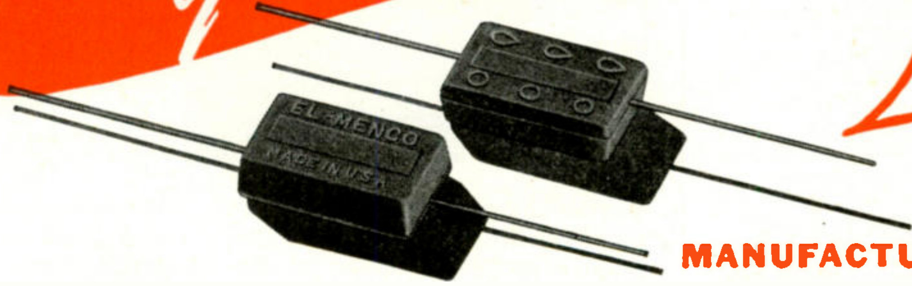

Facts are stubborn things . . . El-Menco Capacitors are backed by impressive facts . . . proven performance, dependable quality . . . earned as components in the world's finest radio and electronic equipment.

You can't discount the reputation of leadership El-Menco has built with the most renowned manufacturers. If you want your product to win preference through perfection . . . use El-Menco Capacitors . . . improve its performance.

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Willimantic, Connecticut

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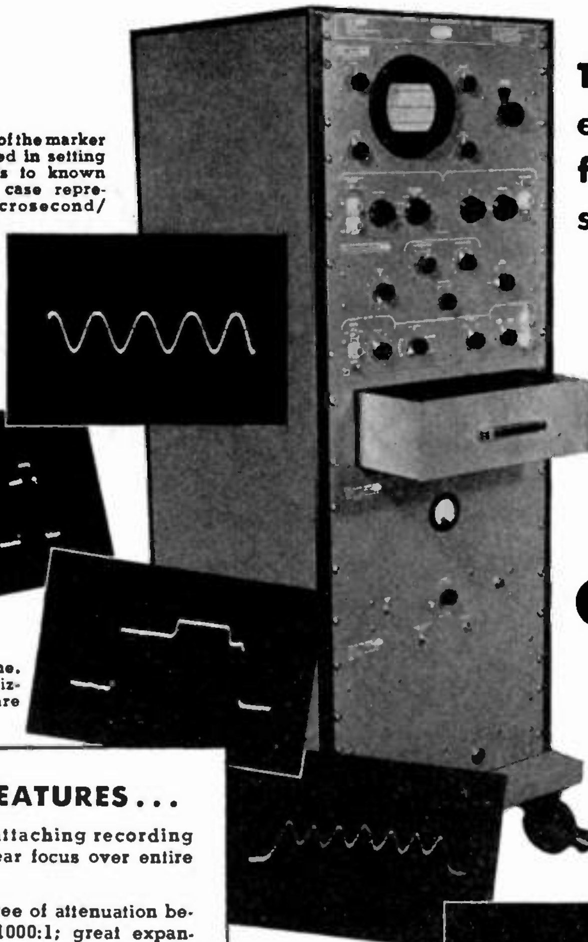
CAPACITORS



Television waveforms selected even to the scanning line and fraction of that line, for critical study or recording, with the <u>new</u>

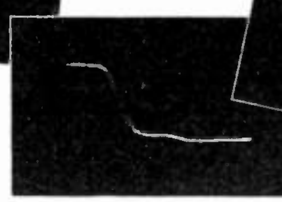
DU MONT
Type 280
Cathode-ray
OSCILLOGRAPH

Output of one of the marker oscillators used in setting sweep speeds to known values. This case represents 0.2 microsecond/inch.

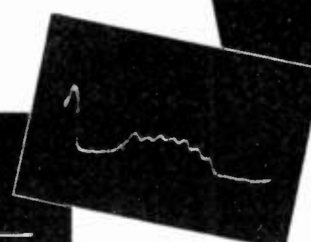
1.2 lines of television signal. Horizontal synchronizing and blanking pulses at each end. Video modulation in center.

Fractional part of a line. Horizontal synchronizing and blanking are shown.

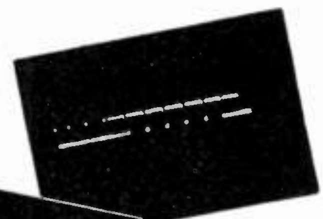
Fractional part of line near center of line. Video modulation produced by wedge, is shown.



Trailing edge of horizontal synchronizing pulse.



Fractional part of line near center of a test pattern where wedge elements are more closely spaced. Note loss in amplitude of modulation.



Vertical synchronizing and equalizer pulses as seen with 60-cycle-sweep repetition rate; used for checking interlace.

OTHER FEATURES . . .

Provisions for attaching recording camera. Fine, clear focus over entire length of trace.

Y-axis: Any degree of attenuation between 1:1 and 1000:1; great expansion of negative polarity signal; undistorted deflection of at least 2"; frequency response within 3 db. from 10 cps. to 10 mc.

X-axis: Time-base duration variable from 1 to 15,000 microseconds. Horizontal deflection of at least 4".
5RP-A Cathode ray Tube. 12,000 volt accelerating potential.

Time-base can correspond with any horizontal line in either or both interlaced fields. Calibrating generator for calibration of sweep-writing speeds by signals of 10, 1, and 0.2 microsecond/cycle.

Wide range of sweep-writing speeds; continuous variation between 0.25 and 3000 microseconds/in.

Delay ranges of 100 or 1000 microseconds selectable for linear time base.

Indication as to exact occurrence of time-base with respect to overall television picture.

Interval of 0.25 microsecond may be measured to plus/minus 0.01 microsecond.

DU MONT proudly announces the new Type 280 Cathode-Ray Oscillograph especially designed for television studio and transmitter installations. Here at last is a means for accurately determining the duration and shape of the waveform contained in the composite television signal, as well as the character of the picture-signal video in conjunction with transmitter operation, according to FCC standards and practices.

Excellent for research on all television equipment. Also for study of wide-band amplifiers. Well suited for industrial use wherever high-speed single transients are studied.

Consists of four units mounted on standard relay-rack type panels and chasses, and installed on mobile rack. Removable side and rear panels. Grouped controls for easy operation.

By virtue of its great range of applications, Type 280 becomes a "must" for television studio and research laboratory.

Further Details on Request!

DUMONT *Precision Electronics & Television*

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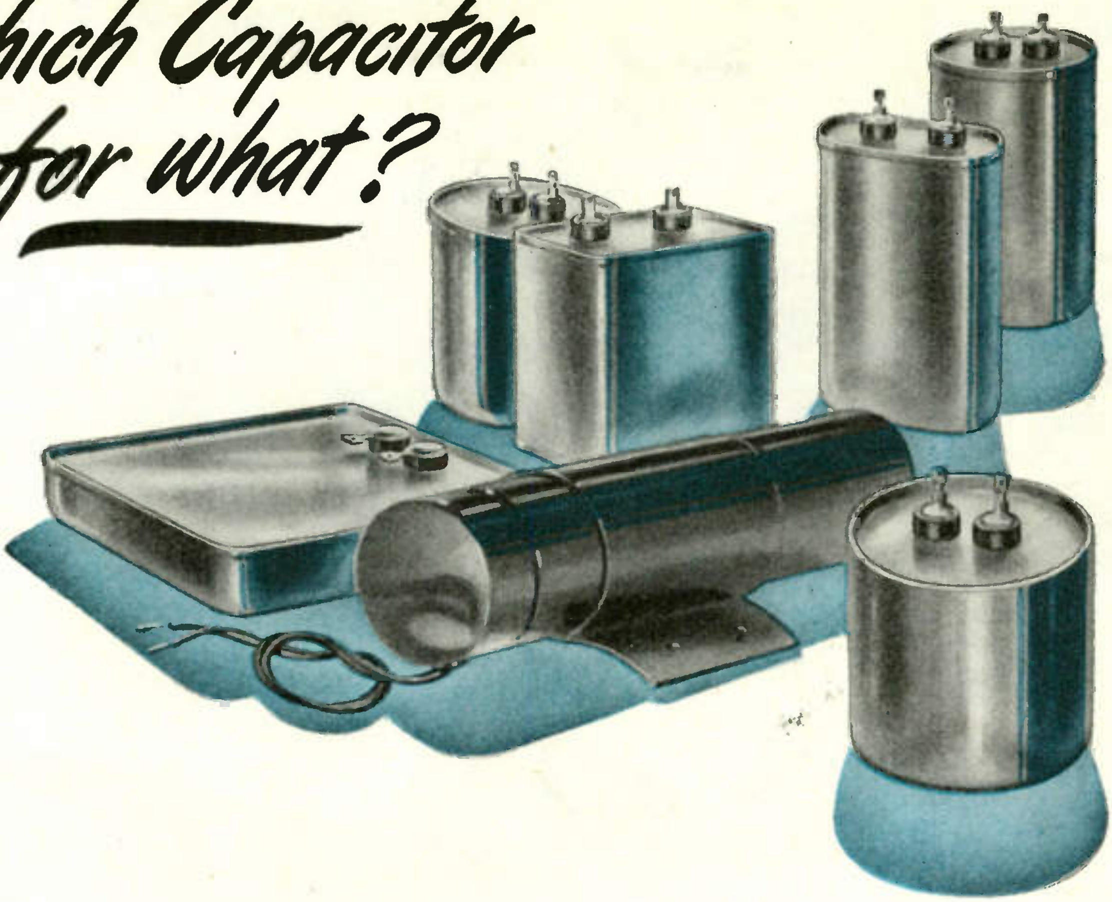
which Capacitor for what?



These capacitors are identical, electrically. The different case styles were, most of them, developed for specific applications. However, since the capacitors *are* electrically the same, it is perfectly practical to use them interchangeably—to use a ballast capacitor on a motor, or a motor capacitor with a sign transformer.

We have made just such proposals at times—and have frequently been able to help manufacturers solve an unusual mounting or space problem, and cut their capacitor costs by recommending a unit not normally thought of for the application.

The capacitor that *you* should use of course depends on your own problem. For assistance in any specific case, get in touch with the nearest G-E Apparatus Office, or write *General Electric Company, Pittsfield, Massachusetts.*

GENERAL ELECTRIC

407-151

Specialty Capacitors **FOR**

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Luminous-tube transformers
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Industrial control
Radio Filters
Radar
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Communication systems
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Dust precipitators
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AND MANY OTHER APPLICATIONS

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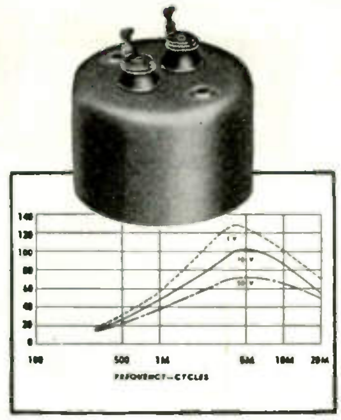
There are many applications in the audio, carrier, and supersonic fields requiring inductors of high Q and great stability. The HQ series of units developed for these applications have remarkable characteristics, as illustrated below. HQA coils have high Q (100 at 5000 cycles) and are available in inductances from 5 MHY to 15 henrys. HQB coils have very high Q (200 at 4000 cycles) and are available in inductances from 10 MHY to 25 henrys.

HUM PICKUP is low due to the toroidal winding structure, 70 and 140 microvolts per gauss respectively for the HQA and HQB at 60 cycles.

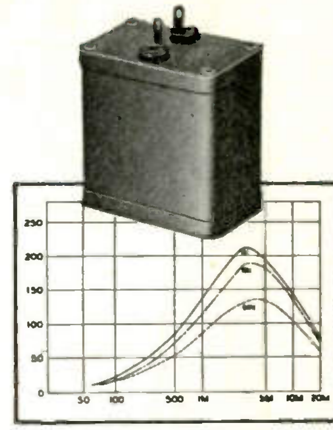
Stability is excellent. For the HQA-7 coil illustrated, inductance change is less than 1% for applied voltages from .1 to 25 volts 1000 cycles. For the HQB-5 coil illustrated, the inductance change is less than 1% for applied voltages from .1 to 50 volts 1000 cycles. Change in inductance due to DC current is approximately 1% per 10 MA linearly for the HQA unit illustrated and 3/4% for the HQB. All cased units are hermetically sealed. Standard inductance tolerance is 1%.



TYPE HQA

DIMENSIONS—1 1/8" Dia., x 1 3/16" H.—Wt. 5 ozs.

Inductance Value	Type No.	Net Price
5 mhy.	HQA-1	$7.00
12.5 mhy.	HQA-2	7.00
20 mhy.	HQA-3	7.50
30 mhy.	HQA-4	7.50
50 mhy.	HQA-5	8.00
80 mhy.	HQA-6	8.00
125 mhy.	HQA-7	9.00
200 mhy.	HQA-8	9.00
300 mhy.	HQA-9	10.00
.5 hy.	HQA-10	10.00
.75 hy.	HQA-11	10.00
1.25 hy.	HQA-12	11.00
2. hy.	HQA-13	11.00
3. hy.	HQA-14	13.00
5. hy.	HQA-15	14.00
7.5 hy.	HQA-16	15.00
10. hy.	HQA-17	16.00
15. hy.	HQA-18	17.00



TYPE HQB

DIMENSIONS—2 5/8" L. x 1 5/8" W. x 2 1/2" H.—Wt. 14 ozs.

Inductance Value	Type No.	Net Price
10 mhy.	HQB-1	$20.00
30 mhy.	HQB-2	20.00
70 mhy.	HQB-3	20.00
120 mhy.	HQB-4	20.00
.5 hy.	HQB-5	20.00
1. hy.	HQB-6	22.00
2. hy.	HQB-7	24.00
3.5 hy.	HQB-8	25.00
7.5 hy.	HQB-9	26.00
12. hy.	HQB-10	27.00
18. hy.	HQB-11	28.00
25. hy.	HQB-12	29.00

UNCASED HIGH Q TOROIDS

We can supply any of the Toroids listed without case. Deduct $1.50. Specify type and inductance value when ordering.

SPECIAL TOROIDS

Sizes other than those shown in our stock list can be supplied on special order at price of next highest value. Type HQC and HQD coils, having maximum Q at 50 kc and 100 kc respectively, are also available.

NEW ITEMS

CG-50 DYNAMIC NOISE SUPPRESSION INDUCTOR

Incorporates two accurately tuned high Q inductors of .8 hy. and 2.4 hy., respectively, for use in dynamic noise suppressor circuits. Write for Circular No. CG-50 for additional details .... List Price $16.00

CGE-1 UNIVERSAL INTERSTAGE EQUALIZER

This new UTC unit is the ideal device for any application requiring frequency response correction. Designed to be connected between two triode audio stages or will match a high impedance (5000 to 30000 ohms) source to grid. The CGE-1 equalizer is not a simple R-C tone control, but employs resonant circuits to permit low or high end equalization without affecting mid-frequencies.

Write for completely detailed manual.

CGE-1 Panel Dim. 2 3/8 x 4" .... List Price $25.00

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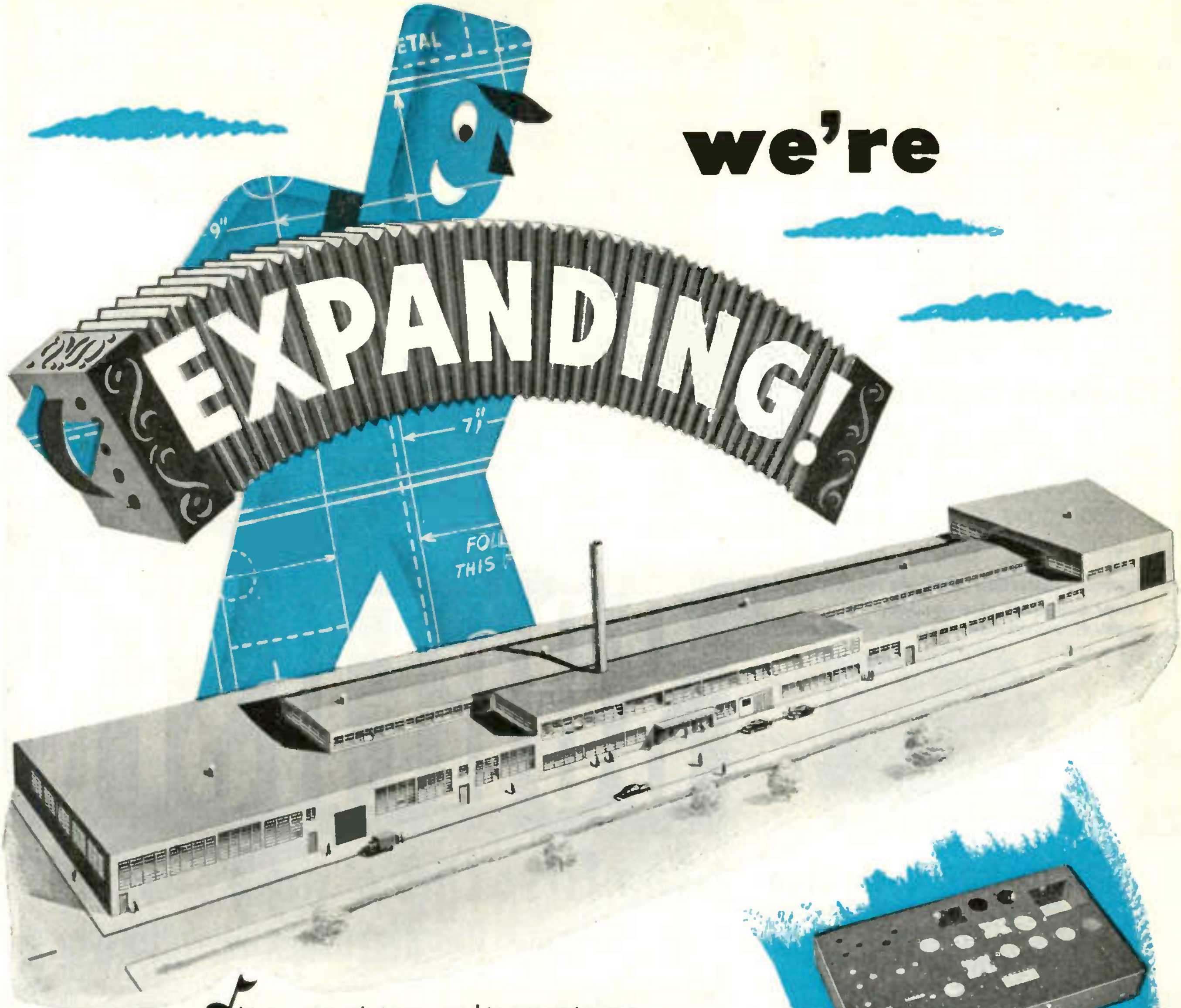

It's sweet music to us...and to our customers, we think, to know that Karp Metal Products Co., Inc. soon will move into a brand new streamlined building of 70,000 square feet of space, with a 600 foot frontage.

Our new plant will be the last word in modern manufacturing quarters, equipped with the newest and most efficient machinery and facilities, including the most complete and up-to-date paint and finishing department, scientifically air conditioned and dustproof. These advancements will enable us to extend the scope of the precision service we render the leaders of the radio and electronics industry.

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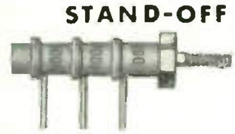
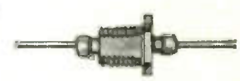
GENERAL PURPOSE CAPACITORS

HI-Q General Purpose Capacitors can be supplied in any size from .200 x .375 to .340 x 1.875, with capacity range from 5 to 33,000 MMF. They are insulated with a clear non-hydroscopic styrene coating. I. R. of 10,000 Megohms, working voltages up to 500 volts D.C. and power factor well under 3 per cent can be assured.

CHOKE COILS

STAND-OFF CONDENSERS





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HI-Q components are uniformly superior because of rigid quality control throughout all stages of manufacture. Final individual inspection insures their conformance to electrical and physical specifications. When you specify **HI-Q** components, you can be sure they meet your most stringent requirements for precision, dependability, compactness and uniformity. Write for complete information and engineering data.

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PRECISION Tested step by step from raw material to finished product. Accuracy guaranteed to your specified tolerance.

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MINIATURIZATION The smallest BIG VALUE components in the business make possible space saving factors which reduce your production costs . . . increase your profits.

HI-Q *Electrical Reactance Corp.*

FRANKLINVILLE, N. Y.

Plants: FRANKLINVILLE, N. Y.—JESSUP, PA.

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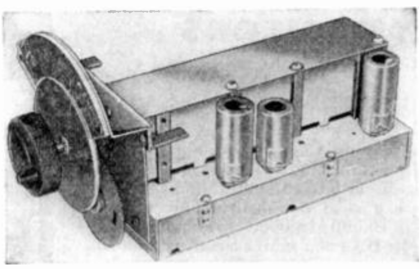
NEWS and NEW PRODUCTS

March, 1948

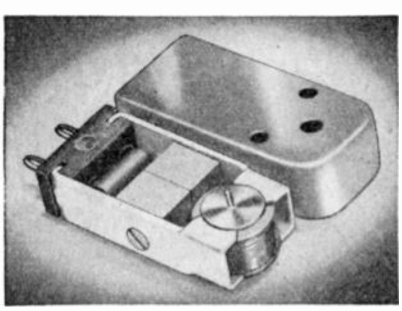
Improved Inputuner

A new model of the Inputuner with several refinements over previous models is announced by **Allen B. DuMont Laboratories, Inc.,** 2 Main Ave., Passaic, N. J. This packaged r.f. head is available to television custom-built and line-production set manufacturers alike, eliminating costly problems and establishing in advance the major items of cost.



The Inputuner is a compact, rugged assembly as easy to install as a speaker. It requires no aligning, adjusting, or calibrating. Built around the Mallory-Ware Inductuner and including all necessary components for the complete r.f. head, it provides for continuous tuning in the 44–216-Mc. range. This means the coverage of all 13 television channels plus the f.m., amateur, aviation, telephone, and commercial services in that range without a break. Only one tuning knob is required for both coarse and fine adjustments.

Magnetic Pickup



The **Clarkstan Corp.,** 11927 W. Pico Blvd., Los Angeles 34, Calif., has recently announced a new variable-reluctance pickup which is claimed to be a high-fidelity, wide-range device of extreme simplicity and ruggedness. The stylus can be instantly removed and replaced by the fingers without the use of tools. This pickup has a flat frequency response beyond f.m. requirements. The needle, which weighs 31 mg., is the armature and is the only moving part. A high-impedance winding is standard, but the unit can be had in impedances of 5, 50, 250, and 500 ohms.

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

Model 10-D Amplifier

The newest addition to the line of Brook high-quality audio amplifiers is the Model 10-D now being manufactured by **Brook Electronics, Inc.,** 34 DeHart Place, Elizabeth, N. J.

This amplifier is a 30-watt rack-mounting unit with 75-db gain, equipped with volume control and on-off switch on the front panel. As in all other amplifiers produced by this company, Model 10-D uses triodes throughout.

Designed essentially for broadcasting stations, recording studios, and high-quality public-address installations, this audio amplifier provides frequency response from 20 to 20,000 cycles within 0.2 db. At 5 watts output, harmonic distortion is only 0.6% and inter-modulation distortion is only 0.2%. Total distortion is claimed to be under $2\frac{1}{2}\%$ at full 30-watt output. The power supply is self-contained. Noise level is 70 db below full output. Power available for external tuner or preamplifiers is 250 volts at 90 ma. and 6.6 volts at 5 amperes.

NOTICE

Information for our News and New Products section is warmly welcomed. News releases should be addressed to **Mrs. Harriet P. Watkins, Industry Research Division, Proceedings of I.R.E., Room 707, 303 West 42nd St., New York 18, N. Y.** Photographs, and electrotypes if not over 2" wide, are helpful. Stories should pertain to products of interest specifically to radio engineers.

Sine Wave Clipper

A new Sine Wave Clipper has recently been announced by **Barker & Williamson,** Upper Darby, Pa., which will be welcomed by many engineers interested in audio-frequency circuits.

This new instrument provides a test signal particularly useful in examining the frequency response and transients of audio circuits. Designed to be driven by an audio oscillator, the clipper provides a clipped sine wave—hence the name *Sine Wave Clipper.*

By feeding the output of the clipper into audio equipment under test and in turn introducing the equipment's output into an oscilloscope, the experimenter or engineer may quickly view and analyze distortion introduced by the amplifier.

A sine-wave analysis after every change in a component becomes time-consuming and tedious. By means of the clipper, however, the effect of making changes in a circuit may be seen instantly and thus guide the course of development in the proper direction. The routine use of the clipped sine wave, in addition to sine-wave measurements, makes for a more complete check on the stability of equipment in regular operation.

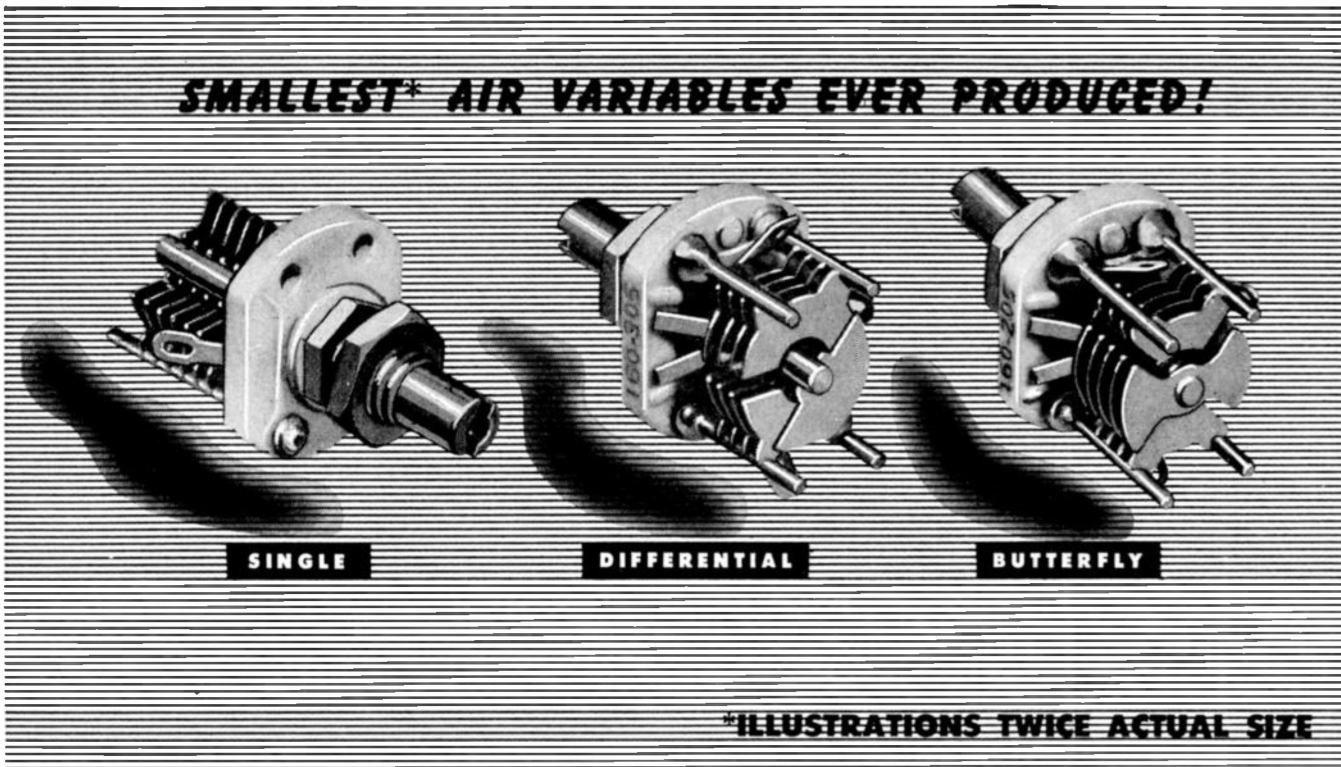
An illustrated instruction book accompanies each Sine Wave Clipper. Complete information on this new device is available from the manufacturer.

New Broadcast-Station Light

A new indicator light, Type "Q," designed especially for radio broadcasting stations, film studios, and applications where controlled warning lights are required has recently been put on the market by **Cannon Electric Development Co.,** 3209 Humboldt St., Los Angeles 31, Calif.

The new light is available in 24 volts, 15 c.p. or 115 volts, 10 watts. It may be wired in multiple for several locations so that all will operate simultaneously. The left, or green, light is illuminated for 10 seconds and then goes out, when the red light (at right) glows and the panel is illuminated with "On The Air" or other brief wording adaptable to the use.

(Continued on page 26A)


SMALLEST* AIR VARIABLES EVER PRODUCED!
SINGLE
DIFFERENTIAL
BUTTERFLY
*ILLUSTRATIONS TWICE ACTUAL SIZE

Where Collins Serves the Airlines

The chart above indicates main routes of airlines using Collins radio communication equipment in the air, on the ground, or both. This tremendous acceptance had its beginning in the middle thirties, and is the result of early and never-ending Collins research and development in the field of aviation communications.

The airlines whose routes are shown include Air France, All American Aviation, American, American Overseas, Braniff, British Overseas, Chicago & Southern, Colonial Airlines, Eastern Air Lines, FAMA (Argentine Republic), Hawaiian, Northwest, Panagra (Pan American-Grace), Pan American World Airways System (Latin American division, Atlantic division, Pacific-Alaska division), Pennsylvania-Central, Peruvian International, Quantas (Australia), Royal Dutch, Sabena (Belgium), SILA — SAS — ABA (Scandinavian Air Carriers), South African, TACA Airways Agency Aerlinte Eireann (Irish), Trans-Australia, Transcontinental and Western, United, and Western.

Our own planes are in constant use, testing equipment of advanced Collins design for Government and commercial aviation. A recent and notable example of accomplishment is the Collins 51R VHF airborne receiver and attendant instrumentation, which equip an airplane for navigational use of the new omnidirectional range system. This equipment was designed and thoroughly tested in 1946, and was demonstrated to the airlines throughout 1947. As a result, Collins has been awarded the majority of the contracts which have been let to the time this announcement is written.

IN RADIO COMMUNICATIONS, IT'S . . .

COLLINS

COLLINS RADIO COMPANY, Cedar Rapids, Iowa

11 West 42nd Street, New York 18, N. Y. **458 South Spring Street, Los Angeles 13, California**

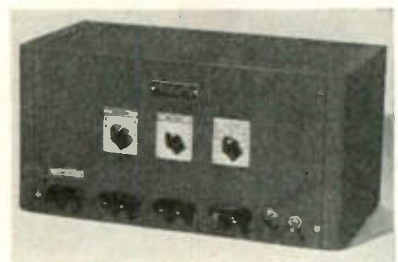
News—New Products

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

(Continued from page 22A)

Multiple Power Supply

The 103 multiple power supply was developed by **Kepco Laboratories, Inc.,** 142–45 Roosevelt Ave., Flushing, N. Y., to meet the need for a source of power that would supply four commonly used voltages from a single compact unit. This multiple power supply eliminates the cumbersome use of three, possibly four, power units to supply heater, plate, and grid voltages. The power supply is particularly designed to be used in the study of the characteristics of vacuum and gas-filled tubes as well as the characteristics of electronic circuits employing these tubes.



The power supply contains two continuously variable B supplies delivering from 0 to 300 volts at currents up to 120 ma., one variable C supply delivering from −50 to +50 volts at 5 ma., and one heater supply delivering 6.3 volts at 5 amperes.

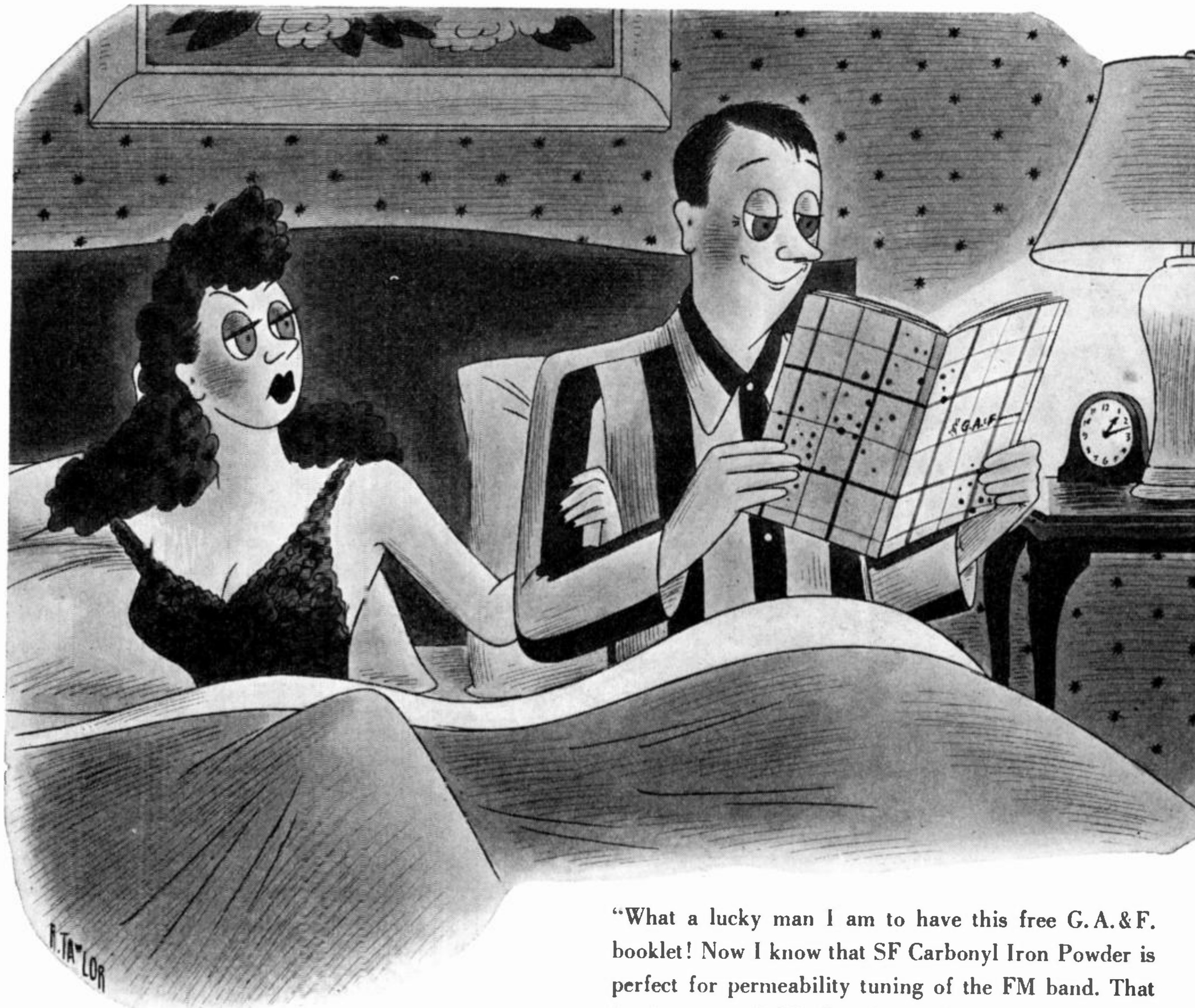
The two B supplies originate from a common power transformer and are controlled by a special circuit containing two 6Y6 control tubes. Each supply will deliver from 0 to 300 volts at 60 ma., or 120 ma. together. The ripple voltage is less than 5 millivolts throughout the entire range of the operating voltage. The supplies are isolated from the chassis to allow grounding of the positive terminal if necessary without affecting the C supply. The voltages are controlled from the front panel.

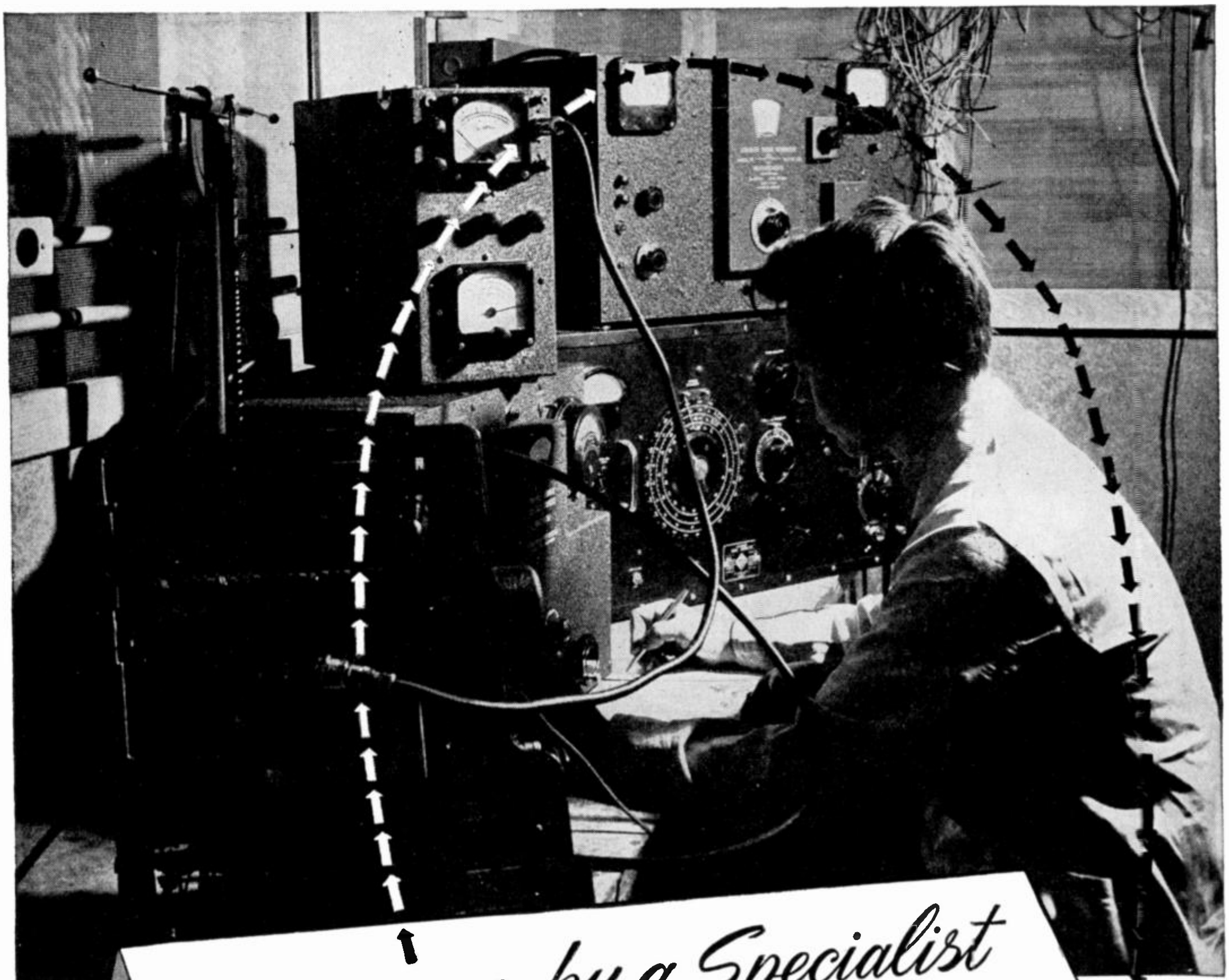
The C supply originates from an entirely separate power transformer and rectifying circuit. A special resistor network allows a continuously variable voltage from −50 volts to +50 volts at 5 ma. The multiple supply is available, upon request, with the C voltage variable from −150 volts to +150 volts at no extra cost. The ripple voltage is less than 1 millivolt throughout the operating range. The C voltage is controlled from the front panel.

Recent Catalogs

• • • On amplifiers, systems, phonographs and accessories, Catalog P9-47A by **David Bogen Co., Inc.,** 663 Broadway, New York 12, N. Y.

(Continued on page 46A)





DIAGNOSIS *by a Specialist*

THE CURE OF RADIO NOISE is a highly specialized task that involves much more than simply "hooking a condenser across the line". It requires exact knowledge of the proper size and type of capacitor to use . . . of the correct place to add it to the noise-making circuit . . . of the necessary length or positioning of connecting leads . . . and of many other seemingly trivial, but actually vital, bits of information that cannot rightfully be expected of the electrical design engineer.

This exact knowledge is available to you when you must provide radio silence for electrical apparatus. Just send us the offending equipment and we will measure its radio noise output according to standard specifications, will design the most efficient Filterette to cure the noise, will specify the proper means of installing it, and, upon your adoption of our recommendations, will authorize your use of the FILTERIZED label that tells buyers your apparatus will not interfere with radio reception. This service is free to users of Tobe Filterettes . . . write for details.

SUPPRESS MAN-MADE RADIO STATIC

TOBE

FILTERIZED

TOBE DEUTSCHMANN CORPORATION NORWOOD, MASSACHUSETTS

ORIGINATORS OF FILTERETTES . . . THE ACCEPTED CURE FOR RADIO NOISE

GL-866-A

There's a type and capacity to meet every broadcast need

FROM mikes to tower, the chain of broadcast equipment must have strong links if "off-the-air" periods are to be avoided with success. General Electric offers you a line of rectifier tubes that will shoulder a full load reliably . . . husky tubes built for around-the-clock performance and plenty of it.

If a designer of transmitters, you may choose from more than a dozen G-E rectifier tubes that run the gamut of sizes. Five are shown here. Mercury-vapor content gives these tubes the ability to pass high peak currents—also keeps the internal voltage drop low. All the tubes are proved veterans of exacting broadcast and industrial service.

If a station operator . . . do you want fast service on rectifier-tube replacements, plus THE BEST in quality? See your nearby G-E tube distributor or dealer. He has the tubes—can get them to you by speedy local delivery; and should his inventory of any type happen to be low, G-E coast-to-coast branch stocks mean overnight replenishment.

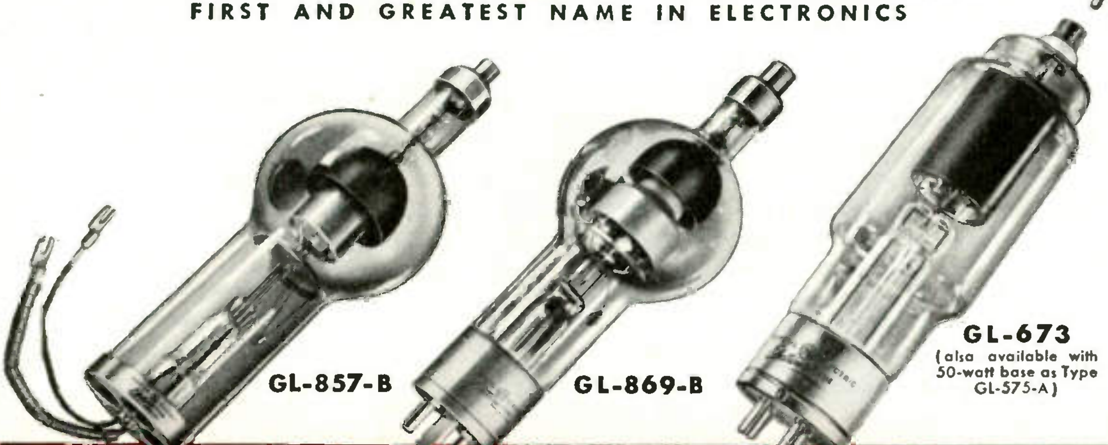
There's pocketbook protection for you, too, in G.E.'s ironclad tube warranty. Specify G-E rectifier tubes in original equipment for efficiency, reliability, and value; replace with G-E tubes to gain the same advantages, plus fast delivery to your door! *Electronics Department, General Electric Company, Schenectady 5, N. Y.*

GL-8008
(also available with 50-watt base as Type GL-872-A/872)

GENERAL ELECTRIC

161-G3-8850

FIRST AND GREATEST NAME IN ELECTRONICS



GL-857-B

GL-869-B

GL-673
(also available with 50-watt base as Type GL-575-A)

Type	Cathode voltage	Cathode current	Anode peak voltage	Anode peak current	Anode avg current
GL-866-A	2.5 v	5 amp	10,000 v	1 amp	0.25 amp
GL-8008	5 v	7.5 amp	10,000 v	5 amp	1.25 amp
GL-673	5 v	10 amp	15,000 v	6 amp	1.5 amp
GL-869-B	5 v	18 amp	20,000 v (*15,000 v)	10 amp	2.5 amp (*5 amp)
GL-857-B	5 v	30 amp	22,000 v	20 amp (*40 amp)	5 amp (*10 amp)

*Quadrature operation)


SYMBOL OF QUALITY
ERIE RESISTOR
Temperature Compensating Molded Insulated Ceramicons
0.5 MMF—550 MMF
Temperature Compensating Dipped Insulated Ceramicons
0.5 MMF—15,000 MMF
Temperature Compensating Non-Insulated Ceramicons
0.5 MMF—1,770 MMF
Types 504B, ½ Watt—518B, 1 Watt Resistors
10 ohms—22 megohms
Erie "GP" Molded Insulated Ceramicons
10 MMF—5,000 MMF
Erie "GP" Dipped Insulated Ceramicons
0.5 MMF—15,000 MMF
Erie "GP" Non-Insulated Ceramicons
10 MMF—10,000 MMF
Custom Injection Molded Plastic Knobs, Dials, Bezels, Name Plates, Coilforms, etc.
Feed-Thru Ceramicons
3 MMF—1,000 MMF
3 MMF—1,500 MMF
Types L-4, L-7, S-5 Suppressors for Spark Plugs and Distributors
Button Mica Condensers
15 MMF—6,000 MMF
High Voltage Double Cup and plate Condensers
10,000 VOLTS WORKING
Cinch-Erie Plexicon Tube Sockets with 1,000 MMF built in by-pass condensers
Type 554 Ceramicon Trimmer
3-12 MMF
5-25 MMF
5-30 MMF
8-50 MMF
Type 557 Ceramicon Trimmer
Type TS2A Ceramicon Trimmer
1.5-7 MMF 3-13 MMF 4-30 MMF
3-12 MMF 5-20 MMF 7-45 MMF
Type 720A
Types 323 and 324 Insulated
Type 2322
Type 2336
Erie Stand-Off Ceramicons
MAKERS OF QUALITY
Electronic Components
ERIE RESISTOR has developed and manufactured a complete line of Ceramic Condensers for receiver and transmitter applications; Silver-Mica and Foil-Mica Button Condensers; Carbon Resistors and Suppressors; Custom Injection Molded Plastic Knobs, Dials, Bezels, Nameplates and Coil Forms. Complete technical information will be sent on request.
Electronics Division
ERIE RESISTOR CORP., ERIE, PA.
LONDON, ENGLAND · TORONTO, CANADA

PROCEEDINGS OF THE I.R.E.

(Including the WAVES AND ELECTRONS Section)

Published Monthly by

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VOLUME 36 — *March, 1948* — NUMBER 3

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.

James E. Shepherd

Board of Directors, 1948–1950

James E. Shepherd was born on May 29, 1910, in Houston, Texas. He studied electrical engineering at the University of Missouri from 1928 to 1933, receiving the degrees of A.B. in 1932, and M.A. in 1933 with a thesis on filter networks. He was active in a variety of student organizations, both in the engineering school and in the University at large, including the Engineering School Council and the University Student Council. He was president of his social fraternity in 1931. He was elected to membership in the honor societies of Tau Beta Pi, Eta Kappa Nu, Phi Beta Kappa, Sigma Xi, QEBH, and Blue Key, and was named a Magna Cum Laude Knight of St. Pat in the engineering school.

In 1934, Dr. Shepherd received a Gordon McKay Scholarship to the graduate school of engineering at Harvard University, from which he received the degrees of M.S. in communication engineering in 1935 and D.S.C. in communication engineering in 1940. His thesis was concerned with the properties of power triodes operating as frequency multipliers. He served as instructor in communication engineering and physics at the Cruft Laboratory of Harvard from 1936 to 1941. During this period, he developed the "wide-range, linear, unambiguous, direct-reading, electronic phase-meter."

In June, 1941, Dr. Shepherd became a project engineer with the Sperry Gyroscope Company, concerned with the development of airborne electronic devices and early radar systems. Since 1943, he has been a research engineer and head of the Armament Radar Department of the Sperry Gyroscope Company, a group of over fifty engineers and technicians engaged in the development of new radar and electronic equipment, responsible for all phases of the engineering of these equipments from the customer contact and early conception stages through the research, development, design, manufacture, and test stages. He holds a number of patents on electronic circuits.

Dr. Shepherd is a member of the American Institute of Electrical Engineers, the Radio Club of America, the Acoustical Society of America, and the Harvard Engineering Society. He joined The Institute of Radio Engineers as an Associate in 1936, transferred to the grade of Senior Member in 1944, and was recently named a Fellow of the Institute. He has been a member of the Executive Committee of the New York Section in various capacities since 1943, and is now the Chairman of the New York Section of the Institute. He is one of two I.R.E. delegates to the Technical Societies Council of New York, in which he served on the original Constitution Committee and at present is Chairman of the Admissions and Membership Committee. He served on the Admissions Committee of the Institute in 1944, the Membership Committee in 1945, the Tellers Committee in 1946 and 1947, and is currently serving on the Sections Committee of the Institute. He served as Vice-Chairman of the 1945 and the 1946 I.R.E. National Winter Technical Meetings, and was Chairman of the 1947 I.R.E. National Convention in New York. He is a member of the General Committee for the 1948 I.R.E. National Convention, and has served as Chairman of the Institute's Convention Policy Committee since 1946.

Dr. Shepherd was recently elected to the Board of Directors of the I.R.E. as a Director-at-large for a three-year term.

The more closely the words of a man approach the truth, the more sturdily they endure, without erosion or erasure, the rude buffetting of time and criticism. It follows that only the most thoughtful and accurate statements of men survive.

The Institute of Radio Engineers was fortunate in numbering among its Presidents the late John Stone Stone, who served effectively in that capacity through 1915. Even in that early day, while the Institute was yet young, Dr. Stone was recognized as a great scientist, an inspiring and brilliant teacher, and one of the leaders of engineering thought in the field of communications.

At the session of the International Electrical Congress, Section G, held at St. Louis, Missouri, on September 12–17, 1904, Dr. stone presented a paper characteristic of his mastery of scientific methods and mathematical problems. This paper was later published in the October 15, 1904, issue of the journal *Electrical Review*. Since this paper is in itself one of the appropriate monuments to the memory of Dr. Stone, and since its contents should be of timely interest to the engineers of a later generation, it has been deemed appropriate to reprint it in the PROCEEDINGS OF THE I.R.E. forty-four years after its original presentation! It has been necessary to re-draw the original illustrations and to make a few minor clerical changes of no scientific significance. Otherwise the paper stands as it left the pen of one of the great builders of the I.R.E. Re-publication of the paper was made possible through the co-operative permission of the McGraw-Hill Publishing Company, which acquired ownership of the *Electrical Review*.

It is hoped that the readers of the PROCEEDINGS will thus become increasingly aware of the great tradition of engineering progress in the communications and electronic field which has been established and maintained by The Institute of Radio Engineers through the decades. The following paper, which was kindly submitted through the helpfulness of Frederick A. Kolster (himself a radio pioneer of high standing), is an admirable illustration of the nature of the work of the Institute and of the nature of the personalities who have guided its activities.—*The Editor*

The Theory of Wireless Telegraphy

JOHN STONE STONE

JOHN STONE STONE

FOREWORD

FREDERICK A. KOLSTER

The remarkable scientific contributions of the late John Stone Stone in the advancement of the radio art, from its earliest conception, have, unfortunately, too often passed unnoticed by the later generation of radio scientists and engineers.

With this thought in mind, and with the sanction of the Editor of the Institute, the above-titled paper, which appeared in the *Electrical Review* of October 15, 1904, is reproduced in its entirety, not only because of its historical interest but also because the information therein contained is completely pertinent to present-day radio technique, especially as it concerns antenna theory and design, a subject which has become of increasing importance with the introduction in practice of ultra-high frequencies.

I believe everyone will agree that this historical paper deserves to become a permanent record in the PROCEEDINGS OF THE I.R.E. in tribute to the memory of one of its most distinguished Past Presidents, and a great teacher who profoundly inspired those whose good fortune it was to have known him and worked with him.

The Theory of Wireless Telegraphy*

JOHN STONE STONE

THE theory of modern wireless telegraphy may be treated in at least two widely different ways depending upon whether it be the object to produce a simple mental picture of the phenomena involved, or whether it be the object to lay the foundations for engineering calculations and quantitative research. The first mode of treatment leads to what may be termed the popular theory, and the latter to what may be termed the working or engineering theory.

In this paper only that form of wireless telegraphy shall be considered in which electrical vibrations are set up in electrical oscillators whose axes are normal to the earth's surface, and which are connected to the earth's surface at their lower extremities.

Part I—Popular Theory

If the equations for the moving field produced by Hertz's dumbbell oscillator be examined, they will be found to show that, in the equatorial plane of the oscillator, the potential is everywhere zero, that there is no component of magnetic force normal to that plane, and that there is no component of electric force parallel to that plane. From this it would follow that if a perfectly conducting sheet, which is initially at zero potential, be passed through the equatorial plane of the oscillator, no currents will be induced in it by the field of the oscillator. In other words, the presence of the conducting sheet should not distort or otherwise affect the field of force produced by the oscillator.

On each surface of the conducting sheet will exist currents which, in their reaction upon the electric and magnetic field on the corresponding side of the sheet, will be the exact equivalent and take the place of the field of force on the other side of the sheet. These currents will extend radially from the point of intersection of the axis of the oscillator with the conducting sheet, and at any point in the sheet will be equal in amplitude, but opposite in direction or phase on the two surfaces of the sheet. For a radial distance, measured along the sheet from the point of intersection with the axis of the oscillator, approximately equal to one quarter of the length of the wave radiated by the oscillator, the energy of the currents will travel out from and a portion of it back to the oscillator in the time of each oscillation, whereas for points beyond this radius, the energy will all flow away from the oscillator, never to return to it, provided only the conducting sheet be infinitely extended in all directions.

Since the infinitely conducting sheet is a complete barrier between the two regions it separates, it is easy to see that each half of the Hertz oscillator, with its appropriate infinitely conducting and infinitely extensive surface is a complete oscillating system entirely independent of anything which may take place on the other side of the conducting sheet, and that the field of force at or above the conducting sheet is the same as that which would be found at or above the equatorial plane of the complete Hertz oscillator, were the conducting sheet absent. These considerations lead to a very simple and popular theory or means of explaining the manner in which the electromagnetic waves of wireless telegraphy are developed and propagated.[1]

This theory regards the vertical transmitting oscillator of wireless telegraphy as one-half of a Hertz oscillator normal to the earth's surface, which must be regarded as practically infinitely conductive in the immediate neighborhood of the oscillator, or for about a quarter of a wavelength from the point at which the oscillator is connected to the surface of the earth. By this theory, therefore, the waves of wireless telegraphy are developed in exactly the same manner as if the vertical oscillator and its electrical image below the surface of the earth together formed the real oscillator of which the surface of the earth is the equatorial plane.

A graphical representation of this theory is given in Figs. 1 and 2.

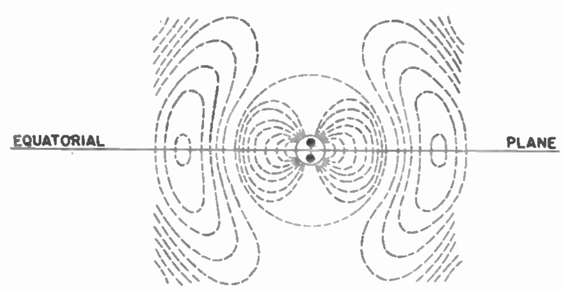

Fig. 1

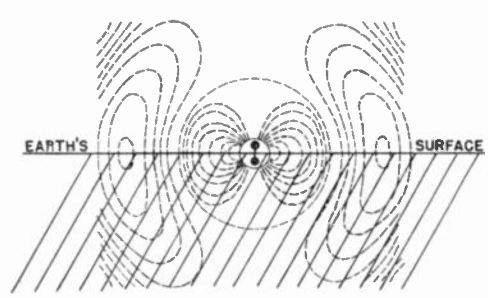

Fig. 2

* Decimal classification: R100. A republication of a paper presented by John Stone Stone before the International Electrical Congress, Section G, at St. Louis, Mo., September 12–17, 1904.

[1] André Blondel, Association Français pour l'Avancement des Sciences, Congress of Nantes, 1898.

This theory, which for convenience may be termed the electrical image theory, bears a close resemblance to that mode of treating a single-wire or grounded telegraph or telephone circuit as one-half of a two-wire or metallic circuit which was first suggested by Oliver Heaviside.[2] He conceives a metallic circuit such as that shown in Fig. 3, cut in half longitudinally by an

Fig. 3

infinitely conducting plane at zero potential as shown in Fig. 4. Since the points on the metallic circuit cut by the plane would normally be at zero potential, no change in the distribution of currents results from the connection with the infinitely conductive plane. A little consideration will also show that the electrostatic capacitance and inductance of the circuit will moreover remain unchanged. The surface of the earth is not infi-

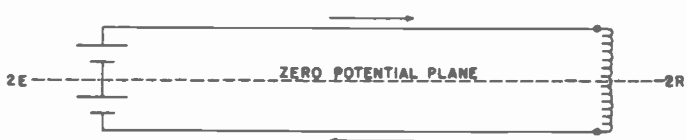

Fig. 4

nitely conductive, however, and therefore neither the assumptions made in the electrical image theory of the transmitting oscillator of wireless telegraphy or the electrical image theory of the grounded telephone line are completely justified, though the conditions of the theory may be more nearly approximated in the case of wireless telegraphy, as will become apparent later.

Before proceeding to a consideration of a more comprehensive theory, some of the more obvious conclusions to be drawn from this theory may well be stated. These are:

1. The waves which emanate from the vertical oscillator are horizontally polarized electromagnetic waves.

2. The energy of these waves will diminish as the square of the distance from the oscillator, if the surface of the earth be assumed to be flat.

3. The energy of the waves is greatest at the earth's surface and diminishes gradually as the point of observation is raised above the earth's surface.

4. The waves do not induce currents in the earth's surface, except when the surface deviates from the equatorial plane of the system formed by the vertical oscillator and its electrical image.

5. At points where the earth's surface is at an angle to the equatorial plane of the system formed by the oscillator and its electrical image, the currents which will be induced in the earth's surface tend to bend the wave front at the earth's surface into a position normal to that surface.

6. In consequence of the tendency of the wave front at the earth's surface to maintain itself normal to that surface, the waves will not necessarily travel in straight lines, but will tend to follow the earth's surface, whatever be its contour.

7. Owing to the fact that when the waves meet irregularities in the earth's surface, currents are developed in that surface which dissipate a portion of the energy of the waves, the energy of the waves will, in general, be better conserved when the transmission takes place over the surface of the sea than when it takes place over land, and more particularly when the land is mountainous or heavily wooded.

The first four consequences of the electrical image theory, above cited, follow directly from the ordinary theory of the Hertz oscillator, while the sixth and seventh consequences cited above are self explanatory. It therefore remains to consider the fifth consequence. For this purpose it will be sufficient to consider what happens to the wave front when a plane-polarized electromagnetic wave falls upon a conducting surface inclined at a definite angle to the plane of the electric force and at a definite angle to the plane of the magnetic force. Under those conditions, only that component of the electric force which is parallel to the conducting surface is effective in producing a current in the surface, and the energy of this component of the electric force is therefore dissipated or redistributed, partly in the form of heat in the surface and partly in a reflected wave which travels off in a direction normal to the surface.

The remainder of the electric force of the primary wave at the conducting surface is therefore normal to that surface.

That component of the magnetic field at the conducting surface which is normal to that surface likewise tends to develop a current in the surface, and its energy is likewise redistributed in the form of heat and in the production of a reflected wave. The remaining magnetic force of the primary wave at the conducting surface is therefore parallel to that surface. The direction of motion of the primary wave must be normal both to the magnetic force and to the electric force, and will therefore be parallel to the conducting surface. It follows, therefore, that the electromagnetic waves of wireless telegraphy emanating from a vertical oscillator grounded at its lower extremity will pass over and around hills and other irregularities in the surface of the earth, and that they will also follow the general curvature of the earth.

The electrical image theory lends itself to the explanation of most of the phenomena of wireless telegraphy in a gross and qualitative way, for it is not, in general, a very difficult task to make the surface of the earth in the immediate neighborhood of the oscillator highly conductive, and at greater distance from the oscillator the current density in the surface of the earth is so

[2] Oliver Heaviside, "Heaviside's Electrical Papers," vol. 2, p. 326.

slight that the conductivity need be but slight in order to guide the waves without great loss of energy. This theory is, however, ill-adapted to give quantitative results, and particularly the class of quantitative results most desired by the wireless telegraph engineer, for he is as much, if not more, interested in the currents and potential in the vertical oscillator as he is in the field surrounding the oscillator. Moreover, the vertical oscillators best adapted for wireless telegraph purposes are quite different from the Hertz dumbbell oscillator, and the field produced by the electrical oscillations of a system formed of one of these oscillators and its electrical image would in many instances be difficult to predetermine.

Some roughly quantitative results which may be predicted by this theory are:

The rate of radiation of energy is caeteris paribus proportional to the square of the length of the oscillator, the square of the quantity of electricity set in motion in the oscillator, and the fourth power of the frequency of the oscillations.

If we assume that the receiving vertical oscillator is exactly similar to the transmitting oscillator, and is as good an absorber as it is a radiator, then the energy received should be directly proportional to the fourth power of the lengths of the oscillators and inversely proportional to the square of the distance separating them, and we should therefore expect that with a receiver of a given sensitiveness, i.e., requiring a given amount of energy to operate it, the distance to which transmission could be carried on between these two stations would caeteris paribus be proportional to the square of the lengths of the oscillators at the two stations.[3]

Part II—Working Theory

When the effects of radiation may be neglected, it is in general not excessively difficult to predetermine the electrical vibrations in simple electrical systems. The problem is then much the same as that of determining the mechanical vibration of mechanical systems, and the modes of attacking such problems have been exhaustively treated and are to be found in the literature.[2,4,5]

In wireless telegraphy, however, the damping of the vibrations in the vertical oscillator is almost wholly due to the radiation of energy from the oscillator, and the effect of this radiation cannot be neglected, whether the oscillator considered be a transmitting or a receiving oscillator. It is often possible, however, to use the same mathematical methods in treating those cases which involve radiation as are applicable in the study of cases with no radiation, and in order to illustrate this point, a very simple system may first be considered.

Let a source of electromotive force be connected in a straight uniform wire at a point distant a from the end of the wire, which end shall be assumed to be insulated, and let the wire extend to infinity on the other side of the source. Such a system is illustrated diagramatically in Fig. 5.

Fig. 5

In order to exclude the possibility of radiation from this wire, it may be assumed to lie in the axis of a perfectly conducting cylindrical shell. The conductor will then have uniformly distributed resistance, inductance, leakage, and permittance as in the case of a single wire cable. If now the electromotive force of the source vary abruptly by changing from one constant value to another, two waves of potential and current will be developed in the wire. This, of course, means two waves of electric and magnetic force about the wire. One of these waves will travel off from the source to infinity along the wire, carrying with it a portion of the energy developed by the source, while the other wave travels from the source to the insulated terminal of the wire, is there reflected, and returns along the wire past the source and on to infinity along the wire, taking with it the remainder of the energy developed by the source, with the exception of that which has been converted into heat in the wire.

The distance apart of the two waves as they travel off to infinity will be four times the distance from the source to the insulated end of the wire, or if the distance between the two waves' fronts be designated by λ, then

$$\lambda = 4a.$$

It will be readily seen that the infinite wire to the right of the source shown in the system illustrated in Fig. 5, draws off the energy from the source and the rest of the system in much the same way as that in which the conducting surface of the earth is supposed to draw off the energy from the vertical oscillator in the electrical image theory considered in Part I of this paper. The wire to the left of the source may therefore be likened to the vertical oscillator, and the infinite wire to the right may be likened in its function to the infinite conducting plane of that theory.

The operational solution of the problem just considered in the case of pure diffusion has been given by Heaviside[5] who also shows how such operational solutions may be readily converted into the ordinary algebraic form, both in the case in which the impressed electromotive force varies as a simple harmonic function of the time, and in the case in which it abruptly changes from one constant value to another.

Let the impressed electromotive force be e. Let the resistance, inductance, leakage conductance and per-

[3] This relation between the lengths of the vertical oscillators and the distance to which transmission may be successfully carried has been empirically determined by Mr. Marconi and is termed "Marconi's Law," by Professor Fleming.

[4] Lord Rayleigh, "The Theory of Sound."

[5] Oliver Heaviside "Electromagnetic Theory," vol. 2, p. 69.

mittance per unit of length of the wire be respectively R, L, K, and S.

Then if distances along the wire be measured from the insulated end of the wire and be designated by x, the potential for points to the right of the source will be:

$$V_1 = \frac{e}{2}\left(\epsilon^{-q(x-a)} - \epsilon^{-q(a+x)}\right)$$

and for points to the left of the source the potential will be:

$$V_2 = -\frac{e}{2}\left(\epsilon^{-q(a-x)} + \epsilon^{-q(a+x)}\right)$$

where

$$q = \left\{(K+Sp)(R+Lp)\right\}^{1/2}.$$

The corresponding currents are

$$C_1 = \tfrac{1}{2}e\sqrt{\frac{K+Sp}{R+Lp}}\left(\epsilon^{-q(x-a)} - \epsilon^{-q(a+x)}\right)$$

to the right, and

$$C_2 = \tfrac{1}{2}e\sqrt{\frac{K+Sp}{R+Lp}}\left(\epsilon^{-q(a-x)} - \epsilon^{-q(a+x)}\right)$$

to the left.

At the source the current is:

$$C_0 = \tfrac{1}{2}e\sqrt{\frac{K+Sp}{R+Lp}}\left(1 - \epsilon^{-2qa}\right).$$

At the source, the potential on the right and left of the source is

$$V_{01} = \tfrac{1}{2}e(1 - \epsilon^{-2qa})$$

to the right, and

$$V_{02} = \tfrac{1}{2}e(1 + \epsilon^{-2qa})$$

to the left.

The resistance operator of the wire measured from the source to the right is:

$$Z_1 = \frac{V_{01}}{C_0} = \sqrt{\frac{R+Lp}{K+Sp}},$$

while the resistance operator measured to the left from the source is

$$Z_2 = \frac{V_{02}}{C_0} = -\sqrt{\frac{R+Lp}{K+Sp}}\,\frac{1+\epsilon^{-2qa}}{1-\epsilon^{-2qa}}.$$

If e be a simple harmonic function of the time and of frequency $n/2\pi$, it is sufficient to substitute ni for p in the above expressions in order to algebraize them.

In this case, therefore,

$$Z_1 = \left(\frac{RK+LSn^2}{K^2+S^2n^2} + in\,\frac{KL+RS}{K^2+S^2n^2}\right)^{1/2}$$

which shows that, so far as the currents and potential in the rest of the system are concerned, the infinite length of wire to the right of the source may be replaced by any device having dissipative resistance

$$\left\{\frac{1}{2(K^2+S^2n^2)}\left(\sqrt{(R^2+L^2n^2)(K^2+S^2n^2)}+RK+Ln\right)\right\}^{1/2}$$

and reactance:—

$$\left\{\frac{1}{2(K^2+S^2n^2)}\left(\sqrt{(R^2+L^2n^2)(K^2+S^2n^2)}-RK-Ln\right)\right\}^{1/2}$$

such device being grounded as shown in Fig. 6. Another arrangement which is the exact equivalent of the systems shown in Figs. 5 and 6, is shown in Fig. 7.

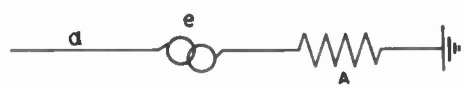

Fig. 6

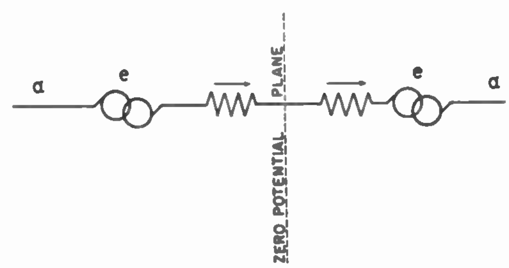

Fig. 7

If the wire be of copper and the frequency of e be sufficiently great, a condition always present in the vertical oscillators of wireless telegraphy Z_1 reduces to $\left(\frac{L}{S}\right)^{1/2}$ or by the relation $S=\frac{1}{Lv^2}$ it further reduces to Lv where v is the velocity of light.

Under these conditions, the device A of Figs. 6 and 7, which takes the place of the infinite wire to the right of the source in Fig. 5, becomes a simple resistance of value Lv.

This resistance is such as completely to absorb the energy of the waves which emanate directly from the source, and of those which are reflected from the insulated end of the wire to the left of the source. It corresponds exactly, therefore, in its reaction on the rest of the system, to the reaction produced by the infinite extension of the wire to the right of the source in drawing away the energy from the rest of the system. It may be likened to the reaction produced on the system by the complete radiation of its energy in each half period.

To illustrate the application of the foregoing considerations to an oscillator of known form, they may be employed to determine the relation between the impressed force and current in the Hertz dumbbell oscillator.

In the case of this oscillator the energy radiated per second is $\Phi^2n^4/3v^3$, where Φ is the maximum electrical moment of the oscillator expressed in absolute electrostatic units. The amplitude of the current is $\Phi n/2a$ in the same units, $2a$ being the length of the oscillator; therefore, the value of the resistance which must be conceived to be placed in the oscillator in order to simulate the effect of radiation from the oscillator is $8/3\ a^2n^2/v$ in absolute electromagnetic units, or $8a^2n^2/(9\times10^{19})$ ohms. The oscillator may now be treated as if it were a circuit from which there is no radiation, but having resistance

$$2Ra + R' = 2Ra + \frac{8}{3}\frac{a^2v^2}{v},$$

inductance

$$L' = 4a\left(\log_e \frac{4a}{\rho} - \frac{3}{2}\right)$$

and permittance

$$\delta' = \frac{r}{2v^2},$$

where ρ is the radius of the wire connecting the two spheres of the oscillator and r is the common radius of the spheres. If then e_0 be the amplitude of the impressed simple harmonic force which maintains the oscillations of periodicity $n = 2\pi/T$ and c_0 be the amplitude of the resulting current:

$$c_0\left\{(2Ra + R')^2 + \left(L'n - \frac{1}{S'n}\right)^2\right\}^{1/2} c_0$$

which suggests the more general expression

$$e = \left(2Ra + L'p + \frac{1}{S'p} - \frac{8}{3}\frac{a^2}{v}p^{-1}\right)^c.$$

Where, as before, p stands for the operation of differentiation with respect to time, p^{-1} for the inverse operation of integration with respect to the time, and where R is the true dissipative resistance per unit of length of the wire connecting the spheres of the oscillator. It should be carefully noted, however, that the mathematical solutions so far obtained for the field of force about a Hertz oscillator are only applicable when the length of the oscillator is a small fraction of one-half of the length of the wave radiated by it into space. When this condition is fulfilled, the oscillator may be regarded as a straight current element of length $2a$, the current at every point of which is $\Phi n/2a$. The expressions for the field at great distances from the oscillator are then applicable, as are therefore also the expressions for the energy radiated.

Since a straight linear oscillator is the equivalent of an infinite number of such current elements varying in lengths from zero to the full length of the oscillator, the field at a distance from such an oscillator may be determined as the vector sum of the fields produced by the separate uniform current elements.

By considering the straight linear oscillator as composed of a limited or finite number of uniform current elements the field at a distance from the oscillator and the energy radiated may be determined to any desired degree of precision for any given or assumed distribution of current along the oscillator. The value of R', or what may be termed the resistance equivalent of the radiation, may then be determined, and the relation of impressed electromotive force to the currents and potentials along the oscillator may thereafter be treated as if there were no radiation from the oscillator, as in the case of the Hertz oscillator considered above.

The exact predetermination of the distribution of current and potential in a linear oscillator consisting of a straight wire of length $2a$, alone in space, or of a straight wire of length a normal to the earth's surface and connected to the earth at its lower extremity, presents grave difficulties which as yet have not, as far as I am aware, been completely overcome. Fortunately, however, a great variety of cases in modern wireless telegraphy may be readily treated with sufficient precision for engineering purposes upon the assumption that the waves of potential and current travel along the conductor of the vertical oscillator with a constant velocity v.

The distribution of current and potential in a straight wire grounded at its lower extremity through a source of electromotive force e and through a system A whose resistance operator is Z_0 as illustrated in Fig. 6, may next be considered under the above-mentioned assumption. In this instance, it will be convenient to regard distances as measured from the earthed terminal of the oscillator.

The circuital equations for the wire are then:

$$-\frac{dV}{dx} = LpC \quad\text{and}\quad -\frac{dC}{dx} = SpV$$

from which flow

$$\frac{d^2V}{dx^2} = \frac{p^2}{v^2}V \quad\text{and}\quad \frac{d^2C}{dx^2} = \frac{p^2}{v^2}C.$$

The most general solution of these equations is

$$V = A\cosh\frac{p}{v}x + B\sinh\frac{p}{v}x$$

$$C = -\frac{1}{Lv}\left(B\cosh\frac{p}{v}x + A\sinh\frac{p}{v}x\right).$$

At $x = a$, $C = 0$,

$$\therefore\quad B \doteq -A\tanh\frac{p}{v}a.$$

At $x = 0$, $V_0 = A$ and

$$C_0 = \frac{A}{Lv}\tanh\frac{p}{v}a$$

$$\frac{V_0}{C_0} = Lv\cosh\frac{p}{v}a.$$

This is the resistance operator measured from the source in the direction of the insulated end of the wire and shall be designated by Z.

It follows that

$$C_0 = \frac{e}{Z + Z_0}$$

$$\therefore \quad A = \frac{eZ}{Z_0 + Z} \quad \text{and} \quad B = -\frac{eLv}{Z_0 + Z}$$

$$V = \frac{eZ}{Z_0 + Z}\left(Z \cosh \frac{p}{v} x - Lv \sinh \frac{p}{v} x\right)$$

$$C_0 = \frac{e}{Lv(Z_0 + Z)} \left(Lv \cosh \frac{p}{v} x - Z \sinh \frac{p}{v} x\right).$$

In the simple harmonic regimen, $p = in$ and the hyperbolic functions are converted into the corresponding circular functions.

The chief interest to the engineer lies in the functons Z and Z_0, and more particularly in the former which becomes

$$-Lv \cot \frac{n}{v} a \quad \text{or} \quad -\frac{1}{Sv} \cot \frac{n}{v} a.$$

We see that Z vanishes when $n = m(\pi v/2a)$, where m is any integer. This corresponds to the case of $m\lambda = 4a$ where λ is the length of the waves on the wire. For the fundamental or gravest mode of vibration of the oscillator, $m = 1$ and $\lambda = 4a$.

It appears, therefore, that for oscillations graver than the fundamental of the oscillator formed by the wire per se and its electrical image, the reactance Z is negative or a capacitance or permittance reactance, whereas for periodicities higher than that of such fundamental the reactance of the oscillator becomes positive, or an inductance reactance. In other words, the reactance of the wire measured at the source or driving point of the system may be the equivalent of a capacitor of capacitance.

$$S' = \frac{1}{Lvn} \tan \frac{n}{v} a = \frac{Sv}{n} \tan \frac{n}{v} a$$

or of an inductance.

$$L' = \frac{Lv}{n} \cot \frac{n}{v} a = \frac{1}{Svn} \cot \frac{n}{v} a$$

depending upon whether cot $(n/v)a$ positive or negative, respectively.

Curve 1, Fig. 8, shows the variation of the reactance Z, i.e., the reactance of the wire a of Fig. 6 per se for different periodicities n of the impressed force.

Curves 2 and 3 of Fig. 8 show the equivalent capacitance and equivalent inductance of the same wire for different values of the periodicity n of the impressed force, the equivalent capacity being shown by curve 3 and the equivalent inductance being shown by curve 2.

With regard to the resistance operator of the system A of Fig. 6, if this be a simple dissipative resistance R_0

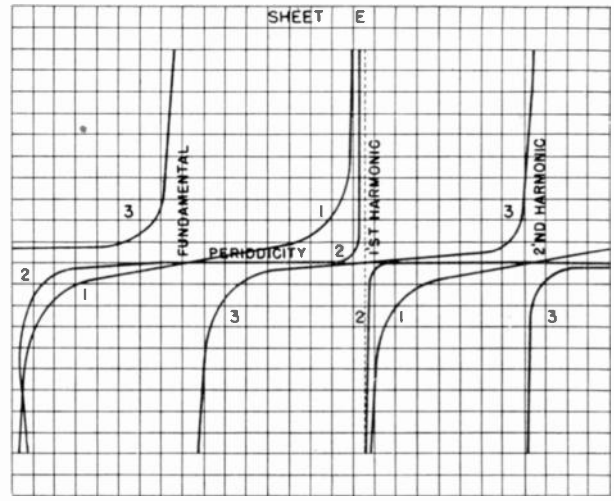

Fig. 8

then $Z = R_0 + R'$. If it be coil of resistance R_0 and inductance L_0, $Z_0 = R_0 + L_0 p + R'$. If there be a condenser of permittance S_0 in sequence with the coil, then

$$Z_0 = R_0 + L_0 p + \frac{1}{S_0 p} + R',$$

and if the condenser be in parallel with the coil,

$$Z_0 = \frac{1 + R_0 S_0 p + L_0 S_0 p^2}{R_0 + L_0 p} + R'.$$

In every case the resistance equivalent of radiation must be added to the resistance operator of the system A. For the high values of the time rate of change of current employed in wireless telegraphy,

$$Z_0 = R' + R_0,$$

$$Z_0 = R' + L_0 p,$$

$$Z_0 = R' + L_0 p + \frac{1}{S_0 p},$$

or

$$Z_0 = R' + R_0 \frac{S_0}{L_0} - S_0 p,$$

for the four cases considered above.

For more complex systems the resistance operator may be readily determined by the simple operational method devised by Heaviside. The algebraizing in the case of a simple harmonic regimen is also easily accomplished by the substitution ni for p.

The foregoing treatment applies more specifically to a transmitting linear oscillator. In the case where the oscillator is employed for receiving, the circuital equations become:

$$E - \frac{dV}{dx} = LpC \text{ and } -\frac{dC}{dx} = Spv$$

in which E is the induced electromotive force per unit of length of the wire.

From these equations result

$$\frac{d^2V}{dx^2} = \frac{p^2}{v^2}V \quad \text{and} \quad \frac{d^2C}{dx^2} = \frac{p^2}{v^2}C - Esp.$$

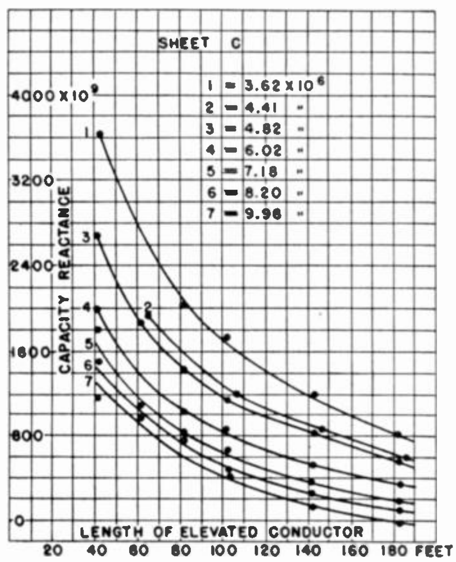

Fig. 9

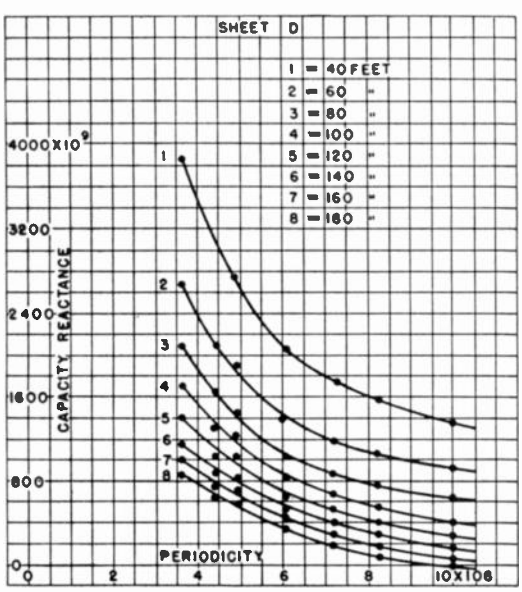

Fig. 10

The general solution is:

$$V = A \cosh \frac{p}{v} x + B \sinh \frac{p}{v} x$$

$$C = \frac{1}{Lp}$$

$$\left\{E - \frac{p}{v}\left(B \cosh \frac{p}{v} x + A \sinh \frac{p}{v} x\right)\right\}.$$

At $x = a$, $C = 0 \therefore B =$

$$\frac{E - A\frac{p}{v}\sinh\frac{p}{v}a}{\frac{p}{v}\cosh\frac{p}{v}a}.$$

At $x=0$, $V_0 = A = -Z_0C_0$

$$\therefore \quad C_0 = E\frac{\cosh\frac{p}{v}a - 1}{Lp\cosh\frac{p}{v}a + Z_0\sinh\frac{p}{v}a}.$$

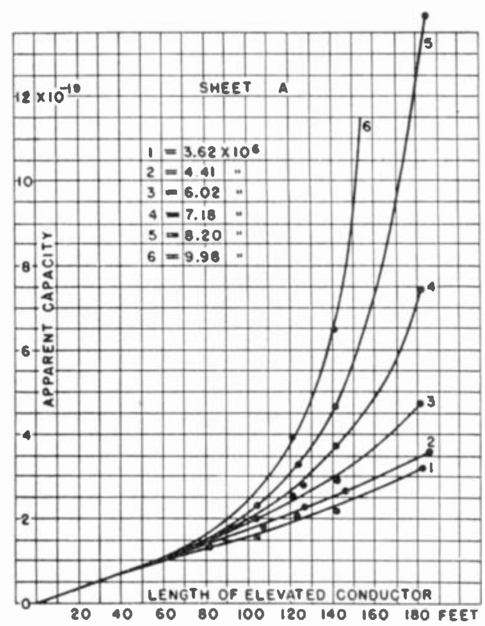

Fig. 11

In the foregoing the explicit assumption has been made that the inductance and capacitance are uniformly distributed along the oscillator and that the velocity of propagation of the waves along the oscillator is equal to that of light. This was done in order to simplify the mathematical analysis, and to present the theory in a concrete and easily understood form; but these conditions do not completely limit the applications of the formulas deduced, for it is capable of demonstration that even when L and S are functions of x provided only that the ratio of L/S be independent of x, then though the velocity of the waves will vary from point to point along the oscillator, yet there will be no reflection of the waves except at the ends of the wire, and the most important function, namely Z, the resistance operator of the oscillator does not change its form. It is sufficient, under these circumstances, to substitute a' for a in the expressions for Z, and C_0 where $a/v = a'/v'$, v' being the average velocity of the waves along the oscillator.

Another important case which may occur is that in which L and S are both functions of x, but in which the product LS is constant. Under these conditions, the quantity $1/\sqrt{LS}$ which is of the nature of a velocity, is constant along the oscillator, but reflection takes place at every point, giving rise to a variable wave velocity.

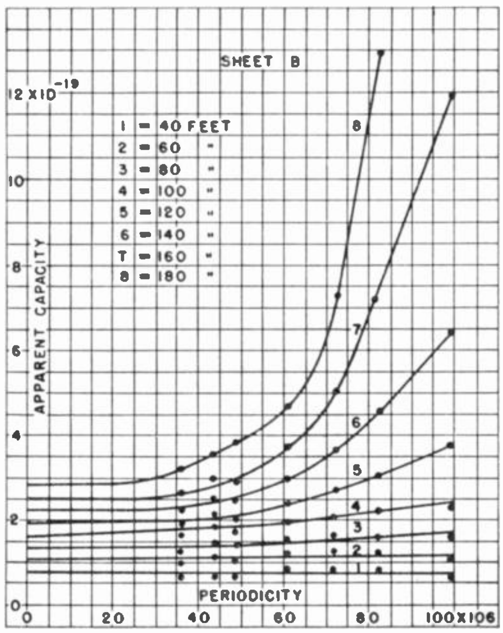

Fig. 12

The solution in this case is no longer of the same form as that considered above, but may be readily obtained in the form of cylindrical harmonics, provided L and S are respectively proportioned to x^m and x^{-m} where m is any quantity integral or fractional; positive or negative.

Some writers have regarded the vertical oscillator as a simple capacitance area. This is obviously inadmissible.

The first approximation to a more complete theory is to regard the vertical oscillator as a capacitance area connected to the earth through an inductance. This mode of treatment corresponds to the first approximation to the theory of the transverse vibration of a stretched string in which the mass of the string is assumed to be collected at its center.

The theory here outlined corresponds to the second approximation to the complete theory of the transverse vibrations of a stretched string in which the mass is assumed to be uniformly distributed along the length of the string.

It is not to be expected that the results of experiments should verify in all details the conclusions to be drawn from the theory which has been presented, but all the most important characteristics of the behavior of a vertical oscillator as indicated by this theory are found to be confirmed by certain experiments, the results of which are presented to you in the form of curves in Figs. 9, 10, 11 and 12.

These curves need no explanation, the title of each showing sufficiently clearly its purport.

Figs. 11 and 12 are the most instructive, showing as they do very clearly the increase of the apparent capacitance of the oscillator as the frequency of the oscillations is gradually increased and the tendency of this apparent capacitance to become infinite as the frequency of the oscillations approaches the frequency of the fundamental of the oscillator per se.

Mr. Stone expressed his thanks to the United States Naval authorities at Washington, and very particularly to Captain E. K. Moore, for the courtesy he had received in being permitted to use the 180-foot wireless telegraph mast of the Boston Navy Yard for the prosecution of the experiments the results of which he had just presented to the congress.

A Proposed Loudness-Efficiency Rating for Loudspeakers and the Determination of System Power Requirements for Enclosures*

H. F. HOPKINS† AND N. R. STRYKER†

Summary—**Experimental and computed data relating to the loudness contribution of various ranges of the frequency spectra of speech and music are correlated with the corresponding energy distribution. A relatively simple measurement of sound pressure and a knowledge of certain acoustic radiation phenomena are applied to this correlation to form the basis of a method for predicting the loudness established by loudspeakers in enclosures. A loudness-efficiency rating for loudspeakers is suggested, and its application to sound-system engineering problems is described.**

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Introduction

For some time those associated with the industrial application of acoustics have expressed the need for a loudspeaker rating directly related to the loudness that the instrument can produce under specified acoustic conditions. Assuming that the suitability of a loudspeaker for its intended use has been determined on the basis of a full appraisal of its various attributes, this paper is confined to the problem of defining its loudness, discussing in detail a study of factors involved in establishing a practical loudness rating, and presenting a method for applying it.

Certain factors based on the theory of loudness[1,2] are used in developing the relationships involved in this study. Loudness is a subjective function, and requires considerable experimental data before it can be quantitatively expressed. Such data have been compiled, although not all are available in published form. Sound intensity, on the other hand, is very generally understood, and is commonly obtained from a measurement of sound pressure. The relationship of loudness and intensity is complex, but is readily derived using physical factors which have been determined experimentally. Since the energy spectrum of the reproduced sound must be known in determining the loudness contribution of specified frequency bands, the present discussion will be limited to speech and music, for which such data are available. These are, of course, the most commonly reproduced sound spectra.

The conclusion is that a relatively simple measurement of sound pressure can be used to determine the loudness efficiency of a loudspeaker. This measurement must be related to the total acoustic output of the instrument, and, therefore, the directivity must be determined. The loudness-efficiency factor thus obtained can be used to determine the loudness per available electrical watt in any enclosure for which the acoustic constants are known. Sound levels, necessary for adequate reproduction of speech and music, are established, and the power requirements for any specified enclosure are readily determined. Certain simplifications have been introduced in the interests of practicability.

Determination of Loudness-Efficiency Rating

Theory

The intensity-versus-frequency distribution in average speech for men and women has been published by French and Steinberg.[3] The data are shown in curve A, Fig. 1, in the form of the intensity per cycle throughout the frequency range for a maximum r.m.s. intensity level of 78 db over 0.25-second intervals. This is a representative level[4] existing at a distance of 2.5 feet from the lips of a person talking conversationally. Curves B and C indicate the intensity-versus-frequency distribution for music played by a 15- to 18-piece and by 75-piece orchestras at intensity levels of 96 and 106 db, respectively.[5,6] These levels[7] are representative and exist at a distance of 30 feet from the source. As shown later, the sound meter and volume indicator, which integrate over 0.25-second intervals, will indicate levels 10 db below those given above, when used to indicate

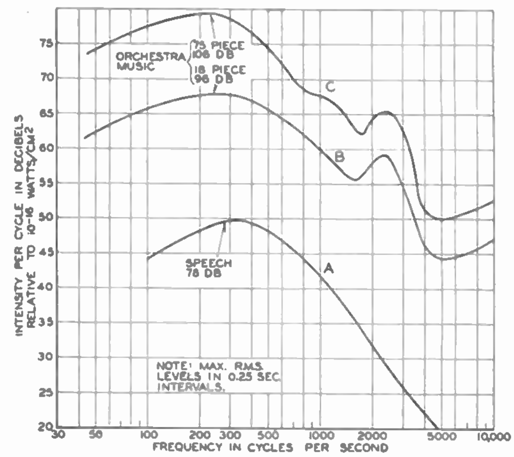

Fig. 1—Intensity-versus-frequency distribution for speech and music.

long r.m.s. intensity levels over intervals much greater than 0.25 second. On Fig. 2 the data for speech and

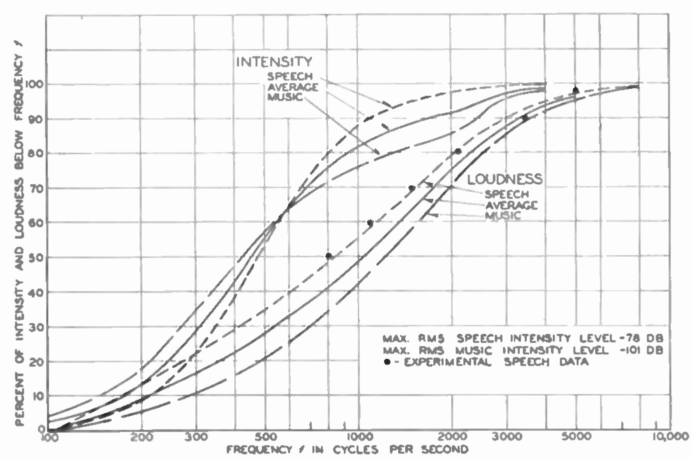

Fig. 2—Percentage of intensity and loudness below any frequency in the spectrum for speech or music.

music are replotted in terms of the percentage of intensity in the frequency band below each frequency of the abscissa. An average curve which is assumed to be sufficiently representative for either speech or music is also shown. The data for speech are more comprehensive than those for music, but, because of the similarity of the two spectra, it is believed that the average should provide a good compromise on which to base an over-all loudness rating.

Experimental speech data from unpublished work of W. A. Munson and given very briefly by Fletcher[6] are shown by the dots on Fig. 2 in terms of the percentage of loudness in the frequency band below each frequency

[1] H. Fletcher and W. A. Munson, "Relation between loudness and masking," *Jour. Acous. Soc. Amer.*, vol. 9, pp. 1–10; July, 1937.

[2] H. Fletcher and W. A. Munson, "Loudness, its definition, measurement, and calculation," *Jour. Acous. Soc. Amer.*, vol. 5, pp. 82–108; October, 1933.

[3] N. R. French and J. C. Steinberg, "Factors governing the intelligibility of speech," presented May, 1945. *Jour. Acous. Soc. Amer.*, vol. 19, pp. 90–120; January, 1947.

[4] 76 db at a distance of 1 meter.

[5] L. J. Sivian, H. K. Dunn, and S. D. White, "Absolute amplitudes and spectra of certain musical instruments and orchestras," *Jour. Acous. Soc. Amer.*, vol. 2, pp. 330–371; January, 1931.

[6] H. Fletcher, "Hearing, the determining factor for high-fidelity mission," Proc. I.R.E., vol. 30, pp. 266–277; June, 1942.

[7] 104 and 94 db at a distance of 10 meters.

of the abscissa. Employing curve A of Fig. 1 and the methods outlined by Fletcher and Munson,[1] loudness computations for speech were made. The results are plotted on Fig. 2. The computed and experimental data for speech agree quite closely. Therefore, it may be assumed with reasonable confidence that the same method of calculating loudness may be applied to other sound spectra such as those indicated for music. Applying this method to curves B and C of Fig. 1, an average curve for music is shown on Fig. 2. A representative average of the curves for speech and music is also shown. It is observed that the average curve for intensity differs greatly from that for loudness. The loudness-versus-frequency distribution will vary somewhat with intensity. Between intensity levels of 90 to 110 db for music and 70 to 90 db for speech, however, the maximum variation is only 2 per cent. The intensity levels employed in this analysis are 78 db for speech and 101 db for music.

If the relationship between loudness and intensity for these typical high-level sound spectra can be established, a simple acoustic measurement involving the intensity may be made indicative of the loudness. The intensity of any part of a sound spectrum is equal to the product of the frequency bandwidth and the average intensity per cycle within that band. Thus, for a flat sound spectrum, equal frequency increments would contribute equal proportions of the total intensity. The loudness contributions for the various frequency increments, as indicated on Fig. 2, however, differ materially from the intensity contributions in the same frequency increments. Therefore, an intensity measurement will be proportional to loudness only if the intensity contributions of the frequency bands are properly weighted.

Complicated relationships between sound intensity and loudness exist for complex, intermittent sounds such as those involved in speech and music. Since we are here concerned with a limited range of sound levels and loudspeakers having relatively uniform response, some simplification can be attained by neglecting certain factors in the general loudness theory. This leads to the conclusion that, within certain limits, a sufficient approximation of loudness may be determined from a loudness-versus-intensity relationship involving only frequency weighting. This relationship may be derived from the curves of Fig. 2, establishing a frequency-weighting factor which, when applied to a sweep-frequency band, reduces equal sweep-time intervals to equal proportions of the total loudness. Applying this sweep-frequency band to a loudspeaker, a single measurement of the resulting sound pressure can be used as a measure of loudness for the intensity-level ranges specified above. Experimental verification of this procedure is presented later.

A frequency band having a range of 100 to 6000 cycles includes 96 per cent of the loudness spectrum. Ten frequency bands in this range which contribute equal loudness increments may be selected, as shown on the abscissa of Fig. 3, with midfrequencies as indicated on the curve. The data from this curve may be utilized to establish the time rate of frequency change for a weighted sweep-frequency band, since the frequency sweep in each of these ten loudness increments should occur during an equal time interval. This relationship is shown by the curve of Fig. 4, the slope of which indicates the rate of frequency change in a sweep-frequency band which, when applied to a loudspeaker, will permit a pressure measurement to be made that is representative of loudness. This measurement of pressure must be related to the total acoustic output of the loudspeaker.

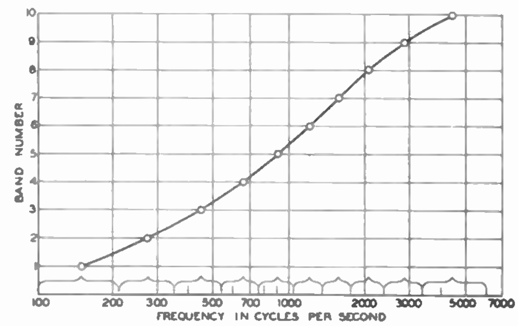

Fig. 3—Midfrequencies of ten equal loudness bands.

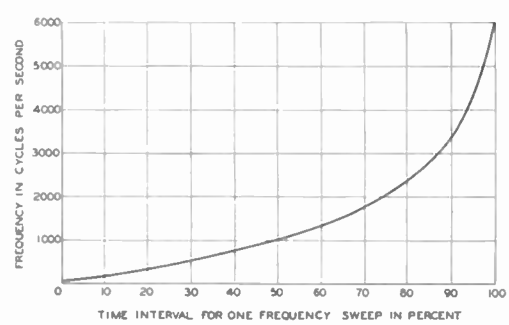

Fig. 4—Frequency-versus-time modulation for loudness weighting.

The total acoustic power radiated from a loudspeaker is a function of the size and shape of the radiating area as well as of the frequency. If the radiating area is a point source the total power is easily derived, because the spatial energy distribution is uniform throughout a solid angle of 4π steradians. As the size of the radiating area of the practical loudspeaker is increased, it becomes more directional. Other investigators[8–11] have derived

[8] Lord Rayleigh, "Theory of Sound," Macmillan Publishing Co., New York, N. Y., vol. II, 1896.

[9] H. O. Stenzel, *Elec. Nach. Tech.*, vol. 6, pp. 165–181; August, 1927.

[10] H. O. Stenzel, *Elec. Nach. Tech.*, vol. 4, pp. 239–253; June, 1927.

[11] I. Wolff and L. Malter, "Directional radiation of sound," *Jour. Acous. Soc. Amer.*, vol. 2, pp. 201–241; October, 1930.

the means for the determination of the total power radiation from a line or a rigid disk radiator located in an infinite baffle. Since all types of loudspeakers are not located in an infinite baffle, the total power radiation must be obtained for other boundary conformations. For a given electrical input, the axial pressure is a function of the efficiency of the loudspeaker as well as of its boundary conformation. Consequently, pressure measurements on the axis of a loudspeaker are indicative of the total acoustic power radiated only if a proper correction factor for the directivity of the device can be determined.

Considering these facts, let us assume first that the loudspeaker is a point source of sound located in free space. An electrical power W_e is supplied over the loudness-weighted sweep-frequency range of 100 to 6000 cycles. The rate of frequency change is assumed to be in accordance with the slope of the characteristic of Fig. 4. The axial sound pressure p_{ax}, in dynes per square centimeter, as indicated by a thermal meter, is determined at a distance of 30 feet from the source. The reason for choosing a test distance of 30 feet will be made evident later.

The sound intensity in watts/cm.², for an electrical power of W_e watts, is

$$I_{ax} = \frac{p_{ax}^2}{\rho c} \times 10^{-7}. \tag{1}$$

The intensity level in db relative to reference intensity (10^{-16} watts/cm.²) is

$$L_{I_{ax}} = 10 \log I_{ax} + 160 \tag{2}$$

or the pressure level in decibels

$$L_{p_{ax}} = 20 \log p_{ax} + 74. \tag{3}$$

Then the total loudness-weighted acoustic power radiated, in watts, is

$$W_L = S_e \times I_{ax} \tag{4}$$

or, if the pressure p_{ax} is used,

$$W_L = \frac{S_e p_{ax}^2}{\rho c} \times 10^{-7} \text{ watts} \tag{5}$$

in which

S_e = surface of a sphere in cm.² having a radius of 30 feet.
= 10.5×10^6 cm.² or 70.2 db relative to 1 cm.²

ρc = characteristic plane-wave impedance of air in mechanical ohms/cm.². If a reference pressure p_c of 0.0002 dynes per square centimeter is assumed at a reference intensity I_0 of 10^{-16} watts/cm.², a value of 40 mechanical ohms per square centimeter follows for ρc. ρc does not actually attain this value for typical atmospheric conditions, but the error due to this assumption is only a few tenths of a decibel.

Defining W_{ec} as the power capacity[12] of the loudspeaker in watts, the total loudness-weighted acoustic power per available electrical watt input is

$$W_{Le} = \frac{S_e I_{ax}}{W_{ec}}. \tag{6}$$

Then L_e, the loudness-weighted acoustic power level in db relative to 1 acoustic watt per available electrical watt, is

$$\begin{aligned} L_e = 10 \log_{10} W_{Le} &= 10 \log_{10} \frac{S_e I_{ax}}{W_{ec}} \\ &= L_{I_{ax}} - 160 + 70.2 - k \\ &= L_{I_{ax}} - 89.8 - k \end{aligned} \tag{7}$$

where $k = 10 \log_{10} W_{ec}$.

In terms of the pressure p_{ax},

$$L_e = 20 \log_{10} p_{ax} - 15.8 - k. \tag{8}$$

Equations (7) and (8) apply to a point source, which is a convenient reference because maximum power is radiated throughout the entire frequency range for a given axial intensity. The acoustic power radiated from a loudspeaker, which is a source of finite size, can be expressed in terms of its ratio to that radiated from a point source. This ratio, K_1 expressed in db, may be termed the loudness-directivity index, and can be applied as a correction factor in (7) or (8). Since loudness-weighted sweep-frequency power has been assumed in determining the effective pressure, either of these equations may be used to derive a loudness rating for loudspeakers. Thus L_e, the intensity level in db relative to 1 acoustic watt per available electrical watt, is

$$L_e = L_{ax} - 89.8 - k - K_1, \tag{9}$$

[12] Considerable thought has been given to an appropriate rating for electrical power input to a loudspeaker. Recently the available-power method has been gaining wide acceptance because of certain simplifications in measurement which result from its use. By this method, the power is defined as that delivered to a resistance R equal to the rating impedance of the loudspeaker from a source of constant voltage E in series with a resistance also equal to the rating impedance. The power available is then $E^2/4R$, and when power to the loudspeaker is referred to, this quantity is meant.

The power capacity of a loudspeaker is then the maximum available power at which satisfactory operation of the instrument may be obtained. Depending upon the type of loudspeaker, the power capacity may be limited, due to distortion or mechanical breakage. Tolerable distortion may be determined by listening tests or measurements, and the value will depend on the requirements involved in the specific type of application. There appears to be no standardized procedure for determining the safe operating point from the standpoint of mechanical failure. At Bell Telephone Laboratories, we have been testing in a manner which appears to insure mechanical stability but which may result in a conservative rating as compared to other methods. For direct-radiator devices, a uniform sweep-frequency band from 50 to 1000 cycles is applied to the loudspeaker set up in the recommended operating condition. The power capacity of the loudspeaker is then considered to be the maximum available power at which no failures occur in a continuous testing period of 100 hours. No ambient temperature is specified except where special applications are involved. For horn-driver units, a sweep frequency 2000 cycles wide whose lowest frequency is 100 cycles below the lowest resonant frequency of the loudspeaker is used. This assumes that the unit is equipped with a recommended horn. Since no standardized method for determining the power capacity of loudspeakers exists at the present time, W_{ec} may be considered to be the manufacturer's rating of his product.

or the total loudness-weighted acoustic power per available electrical watt is

$$W_{L_e} = 10^{\frac{L_e}{10}} . \tag{10}$$

The term loudness-efficiency factor LR has been selected as a suitable expression for rating the loudness of a loudspeaker. Thus,

$$LR = 100 W_{L_e} \text{ in per cent.} \tag{11}$$

This factor provides an expression for the loudness efficiency of a loudspeaker which may be obtained from a simple measurement of the axial sound pressure in a free field for a weighted sweep-frequency power supply, and is suitable for determining the amplifier power and the number of loudspeakers required for a specified sound-system installation.

The correction factor K_1 used in (9) has been termed the loudness-directivity index, and may be defined as the ratio, expressed in decibels, of the total loudness-weighted power radiated by a loudspeaker to that radiated by a point source producing the same axial pressure. It is possible to compute the total acoustic power from most radiating devices at any specific frequency. The ratio of this power to that radiated by a point source producing the same axial pressure, expressed in decibels, may be defined as the directivity index. Since the directivity index of a loudspeaker is a function of the shape of the radiating area and its boundary conditions, as well as of the frequency, these factors must be taken into account in determining the loudness-directivity index. If the directivity index of a given radiating device be computed for each of the ten midfrequencies of the equal loudness bands shown on Fig. 3, the loudness-directivity index may be computed as shown by (53) in the Appendix.

Although loudspeakers exist in a wide variety of shapes, the radiating areas, in general, are simple geometric forms, either baffled or unbaffled, most of which lend themselves to theoretical analysis. Various types of loudspeakers used in practice are described in the Appendix, and derivations of their loudness-directivity indexes are given.

Determination of Test Sweep-Frequency Power

The test-frequency range of 100 to 6000 cycles, which includes 96 per cent of the loudness range, was used for accuracy in computing the loudness-directivity index K_1. In order to attain simplicity in the measuring equipment, a modification in the width of the test sweep-frequency band can be made without materially affecting the results. Fig. 2 indicates that 75 per cent of the loudness as well as the intensity of speech and music occurs between 300 and 3300 cycles (only 1.3 db less than the total). A sweep-frequency band of this width would appear to provide a range adequate for a pressure measurement indicating loudness. The frequency-versus-time relation providing a loudness-weighted sweep band is indicated by the "NORMAL" curve of Fig. 5. The

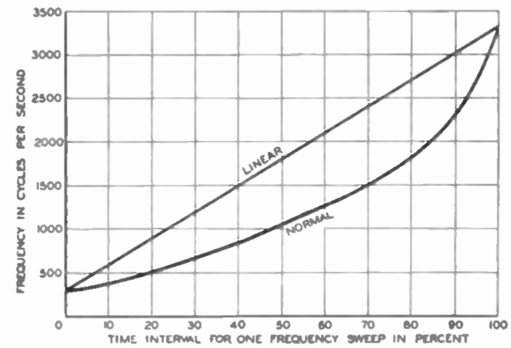

Fig. 5—Frequency-versus-time relations of 300- to 3300-cycle sweep band.

rate of frequency change is not linear and would require a specially shaped capacitor plate in a frequency-modulated generator. It appears desirable from a practical standpoint to make this frequency variation linear throughout the range, as indicated by the "LINEAR" curve of Fig. 5. A linear frequency sweep can be made to produce the same pressure or intensity level as the "NORMAL" frequency sweep if the proper corrective electrical network is inserted in the output circuit of the generator. Equalization for this purpose was computed using the average curve of loudness for speech and music shown on Fig. 2. The computed curve as well as the frequency characteristic of a suitable equalizer providing a close approximation are shown on Fig. 6.

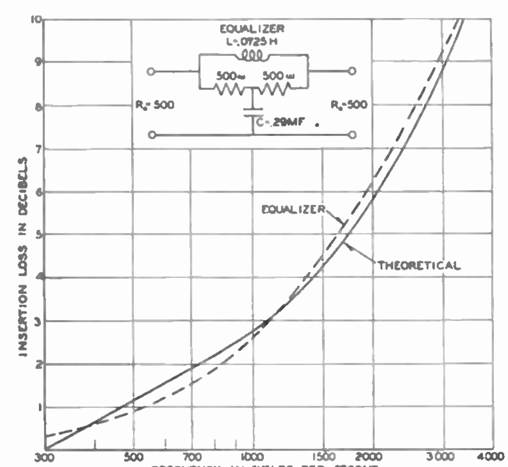

Fig. 6—Loudness-weighting equalization for a linear frequency-versus-time sweep band.

A schematic of this equalizer is also shown on the figure. A possible alternative source of power might be a flat noise spectrum equalized in this manner.

A sweep-frequency band is a frequency-modulated signal in which discrete frequency components result throughout the entire bandwidth, as pointed out by other investigators.[13–15] The amplitude and the number of components are dependent on the modulation index, which is a function of bandwidth and the rate at which the carrier is modulated. The form of the envelope of the components is of great importance in this problem. The most uniform amplitude envelope and the maximum number of components occur for a linear frequency-versus-time relation having a unidirectional frequency sweep, a sawtooth envelope, and a high modulation index. A reciprocating frequency sweep such as that shown on Fig. 7(a), with a sweep rate of 6 per second, produces 500 components having an envelope of the form shown on Fig. 7(b). With the equalizer in circuit, the envelope is modified as shown by Fig. 7(c).

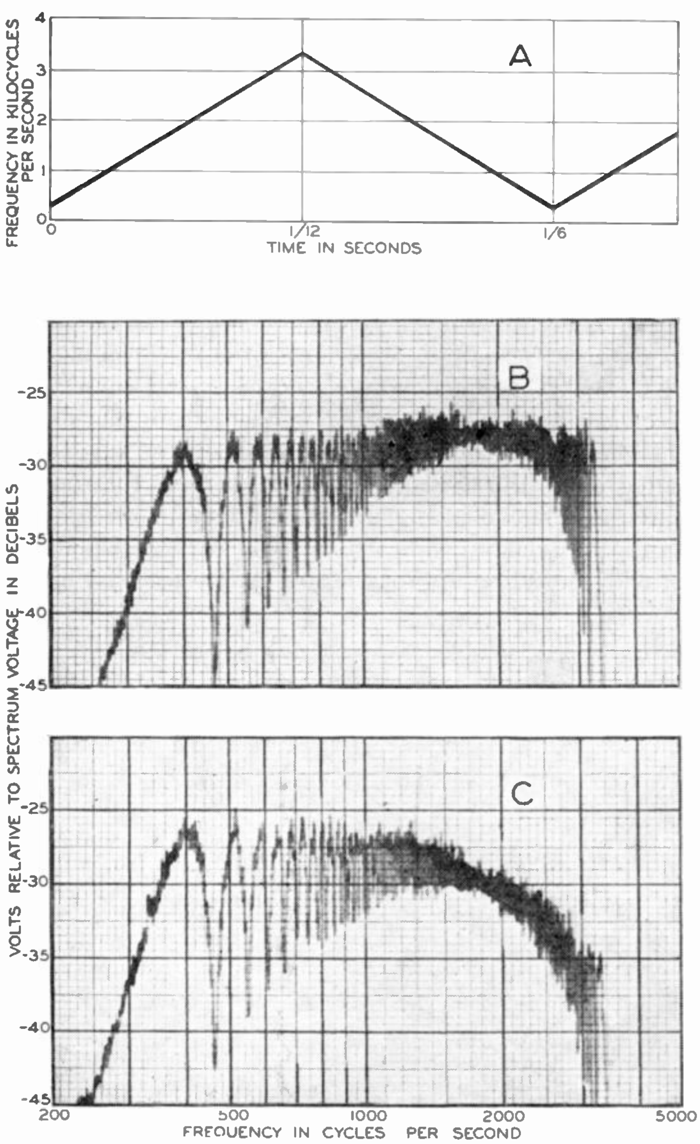

Fig. 7—(a) Frequency-versus-time relation for a linear reciprocating sweep, (b) Envelope of the components of the sweep band of (a). (c) Envelope of the components of the sweep band of (a) with loudness-weighting equalization.

Experimental

In order to justify the validity of this method, the loudness rating of a series of loudspeakers representing a wide range of response-versus-frequency characteristics was determined experimentally. The loudness of these loudspeakers relative to that of a reference condition as judged by a number of observers was also determined for comparison with the measured loudness ratings.

A Western Electric 728-B loudspeaker was used as a reference instrument. The loudspeaker system shown on Fig. 8, in which various test conditions could be obtained by the use of networks, was used for these tests.

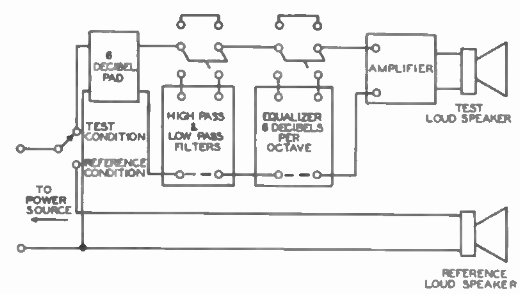

Fig. 8—Test circuit for loudness observations.

The networks employed consisted of low- and high-pass filters and a network producing a 6-db-per-octave rise in the response of the test loudspeaker. This loudspeaker system consisted of a 6-db pad, the networks, a variable-gain amplifier, and a 728-B loudspeaker having practically the same response-versus-frequency characteristic as that of the reference unit. The system was considered as an individual loudspeaker for each circuit condition. The gain settings of the amplifier were different for each condition, to provide a range of efficiencies. A D-173181 Western Electric loudspeaker, developed to provide high intelligibility under noisy conditions, was included as an additional test unit.

Response-versus-frequency characteristics for the reference condition and for each of the test conditions were made in a dead room at a distance of 3 feet from and on the speaker axis, and are shown on Fig. 9. The power supply and the sensitivity of the measuring circuit were held constant in each case. These data permit a determination of the efficiency in the passed band for each condition relative to that for the reference condition.

Using a source of sweep-frequency power having the characteristics shown on Fig. 7(b) and 7(c), the axial pressure p_{ax}, 3 feet from each speaker, was measured in the dead room for a range of input power, all measurements being made with thermal meters. Over the range of power employed, a linear pressure-versus-power relation existed and, therefore, the effective pressure for 1

[13] J. R. Carson, "Notes on the theory of modulation," Proc. I.R.E., vol. 10, pp. 57–66; February, 1922.
[14] Balth van der Pol, "Frequency modulation," Proc. I.R.E., vol. 18, pp. 1194–1205; July, 1930.
[15] W. R. Bennett, unpublished memorandum, Bell Telephone Laboratories, Inc.

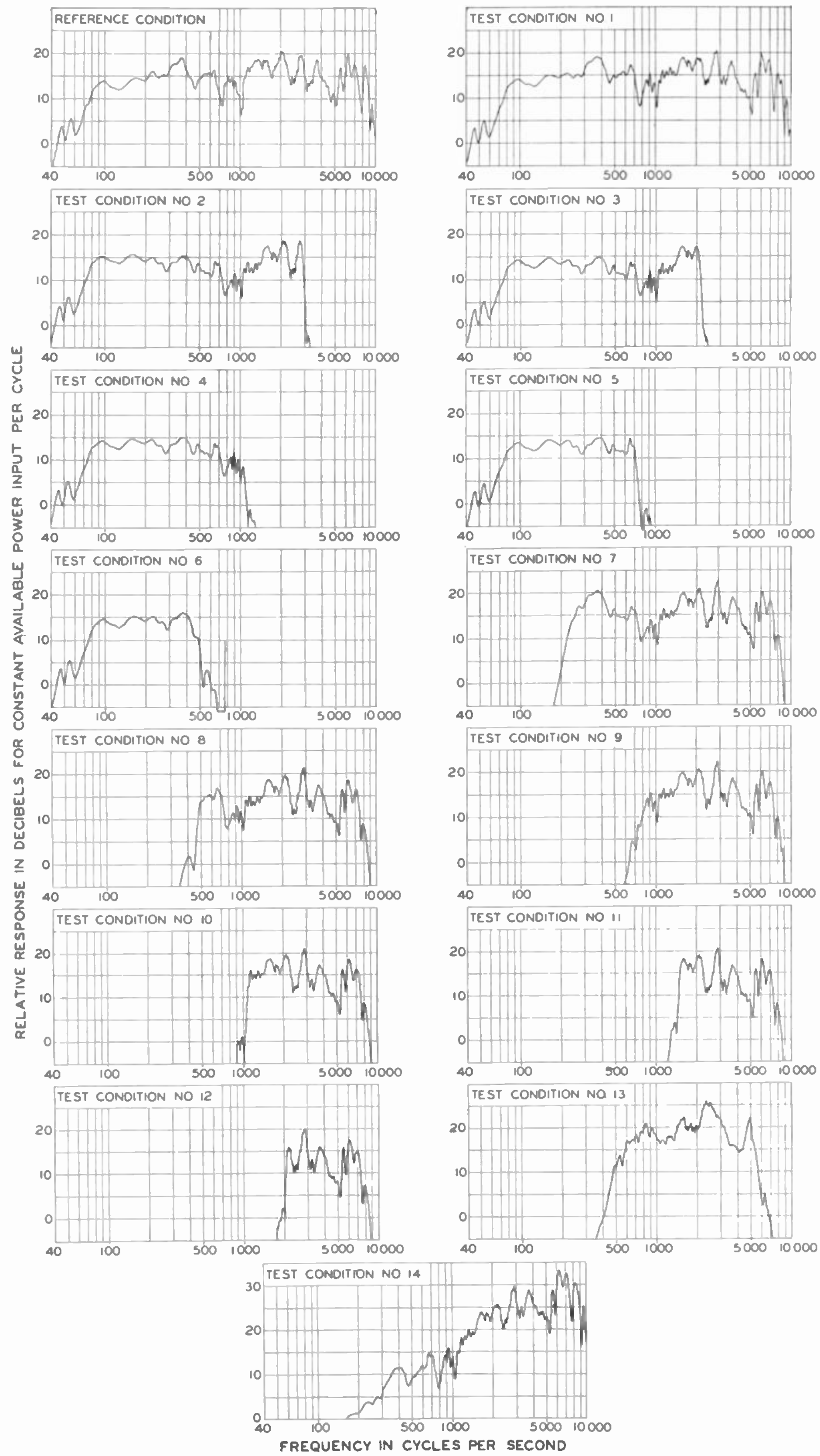

Fig. 9—Response-versus-frequency characteristics of loudspeaker used for the reference and test conditions.

watt input was readily obtained for each test condition. These data and the loudness-efficiency factors, computed from (9), (10) and (11), are shown in Table I. It will be observed that the effective pressures for the unweighted power supply deviate materially from those for the weighted conditions as the frequency range is decreased and the response is made less uniform. This deviation is negative for the low-pass and positive for the high-pass filter conditions. The loudness-efficiency factor for each of the test conditions relative to that for reference condition is expressed in decibels in the last column of the table. If this method of rating loudspeakers is valid, the relative loudness of the loudspeaker conditions represented when judged by an observer listening to speech and music should confirm the data in this column.

To provide the desired correlation of the measured data with aural observations, listening tests were conducted in a room 26×18×12 feet. The observer was located 15 feet away from and in front of the reference and test loudspeakers, which were placed as closely as possible to one another at one end of the room. A switch was provided to permit a quick change to be made from the reference to the test condition. The source material consisted of selected speech and orchestral records. For all tests, the intensity level supplied by the reference speaker at the observer's position was maintained at 68 db for speech and 91 db for music, as indicated by a sound meter. The power in the test loudspeaker was adjusted until the observer judged that reproduction from the reference and test loudspeakers was equally loud. Six observers made judgments for all conditions, while one judged only a few conditions. The resultant data are shown in Table II. The variation in loudness judgment among the observers is surprisingly small except in the cases where highly distorted systems are involved. The loudness for speech reproduction is observed to be approximately the same as that for music for all low-pass filter conditions, but an increasing departure from equality is shown to exist as the cutoff frequency of the high-pass filter condition is raised.

In order to express the aural data in a form which permits comparison with the measured data, the observations for speech and music were averaged together for each condition. These data and the comparable measured data from Table I are shown in the last two columns of Table II. It is observed that very good agreement exists for all except high-pass filter conditions having cutoff frequencies above 1100 cycles per second. It is doubtful if any loudspeaker having distortion as great as this would ever be used in any normal sound reproducing system. The reason for the discrepancy that appears for the high-pass filter conditions will be made evident in later discussion. This experimental work indicates that, from a practical standpoint, a satisfactory measure of the relative loudness for a wide range of loudspeaker conditions may be obtained by the proposed method.

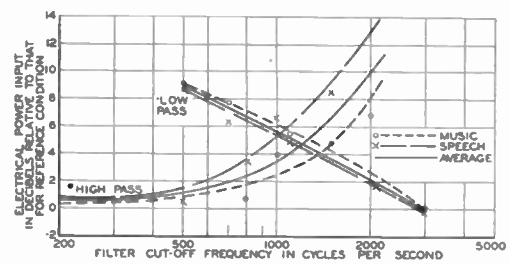

Fig. 10—Power input versus spectrum range of speech and music for equal loudness in a room.

As indicated by the response curves of Fig. 9, the relative pass-band efficiencies are different for the vari-

TABLE I

LOUDNESS MEASUREMENTS OF REFERENCE AND TEST LOUDSPEAKERS

Loudspeaker condition	Effective Pressure p_{az} (dyne/cm.²) at 3 feet for 1 watt of sweep-frequency power — Weighted[16]	Unweighted	Loudness directivity index K_1	Intensity level at 30 feet $L_{I_{az}}$ in db relative to 10^{-16} watts/cm.² — Weighted	Loudness efficiency factor LR—%	Loudness efficiency factor of test condition relative to that of the reference condition in db.
Reference—728-B #1	19	18.6	7.0	79.6	1.89	0
1—728-B #2	17	16.2	7.0	78.6	1.51	− 1.0
2—Condition 1+3000∼LP	14	12.9	7.0	76.9	1.02	− 2.7
3—Condition 1+2100∼LP	10	9.0	7.0	74.0	0.525	− 5.6
4—Condition 1+1100∼LP	6.0	4.2	7.0	69.9	0.203	−10.0
5—Condition 1+ 700∼LP	4.9	3.5	7.0	67.8	0.126	−11.8
6—Condition 1+ 500∼LP	4.0	2.6	7.0	66.0	0.0825	−13.5
7—Condition 1+ 230∼HP	18	17.6	7.0	79.1	1.69	− 0.5
8—Condition 1+ 500∼HP	13.7	14.7	7.0	76.7	0.966	− 2.8
9—Condition 1+ 800∼HP	14.7	15.7	7.0	77.4	1.14	− 2.2
10—Condition 1+1100∼HP	12.0	14.4	7.0	75.6	0.75	− 4.0
11—Condition 1+1500∼HP	9.6	11.3	7.0	73.6	0.475	− 6.0
12—Condition 1+2100∼HP	6.2	7.1	7.0	69.8	0.205	− 9.7
13—D-173181	24	27	4.8	81.6	4.98	+ 4.2
14—Condition 1+6-db-per-octave network	9.2	11.0	7.0	73.3	0.477	− 6.3

[16] Weighted in accordance with equalization of Fig. 6. These values were used in determining loudness-efficiency factors.

ous test conditions. The system was purposedly adjusted to provide these differences so that a range of loudness efficiencies would be encompassed by the tests. However, since the loudspeaker response is relatively flat, a fundamental loudness relationship of considerable interest may be shown by correcting the test conditions for equal passed-band efficiencies. This is the condition that would exist if perfect filters were inserted in the transmission system. The aural data corrected in this manner were plotted as shown on Fig. 10 to indicate this relationship. The ordinates of the curves represent the power required for a loudspeaker having a limited frequency range relative to that for a loudspeaker of the same relative efficiency per cycle reproducing the complete spectrum of speech and music at equal loudness. Since the response of the system is flat, the ordinates are also a measure of the relative intensity levels in the room. For the low-pass filter conditions, the curves are similar for speech and music. For the high-pass filter conditions, however, the relative in-

TABLE II

Loudness Judgment Tests

Test condition versus reference condition	Ratio in Decibels of Electrical Power in Reference Loudspeaker Relative to that in Test Loudspeaker for Equal Loudness										Loudness efficiency factor of test condition relative to that of reference condition in decibels from Table I
	Observer 1	Observer 2	Observer 3	Observer 4	Observer 5	Observer 6	Observer 7	Average for all observers	Range of observations in decibels	Average for speech and music	
1. 728-B #2											
Speech	0	0	0	−1	−1	−1	−1	−0.6	1	−0.4	−1.0
Music	0	0	0	−1	−1	0	0	−0.3	1		
2. Condition 1+3000∼LP											
Speech	−2	−2.5	−4	−3	−3	−3	−3	−2.9	2	−2.8	−2.7
Music		−4	−4	−3	−3	−1	−2	−2.7	3		
3. Condition 1+2100∼LP											
Speech	−7	−4	−5	−5.5	−6.5	−6	−6	−5.6	3	−5.9	−5.6
Music		−4	−4	−8	−8	−8	−8	−6.2	4		
4. Condition 1+1100∼LP											
Speech	−11	−10	−10	−9	−9	−9	−9	−9.5	2	−9.7	−10.0
Music	−11	−9	−9	−10	−10	−10	−10	−9.8	2		
5. Condition 1+ 700∼LP											
Speech	−12	−12	−13	−11	−12	−8	−8	−10.4	5	−11.1	−11.8
Music		−12	−11	−14	−14	−10	−11	−11.75	4		
6. Condition 1+ 500∼LP											
Speech		−14	−15	−15	−14	−13	−14	−14.1	2	−14.0	−13.5
Music		−13	−13	−16	−16	−14	−13	−14.0	3		
7. Condition 1+ 230∼HP											
Speech	−2	−1	−1	−2	−1	0	−1	−1	2	−0.9	−0.5
Music		−2.5	−1	0	0	−1	−1	0.85	2.5		
8. Condition 1+ 480∼HP											
Speech				−1	−1	−3	−3	−1.9	2	−2.2	−2.8
Music				−2	−2	−3	−3	−2.5	1		
9. Condition 1+ 800∼HP											
Speech				−3	−3	−6	−6	−4.2	3	−2.5	−2.2
Music				−1	−1	−2	−1	−1.2	1		
10. Condition 1+1100∼HP											
Speech	−12	−10	−9	−5	−5	−11	−12	−7.8	7	−6.4	−4.0
Music		−7	−6	−4	−4	−7	−5	−5.3	3		
11. Condition 1+1500∼HP											
Speech		−10	−11	−9	−9	−15	−14	−10.8	6	−8.6	−6.0
Music		−10	−8	−10	−11	−4	−4	−7.2	7		
12. Condition 1+2000∼HP											
Speech		−15	−16	−13	−13	−20	−18	−15.1	7	−11.3	−9.7
Music		−14	−8	−11	−10	−9	−7	−9.3	7		
13. D-173181											
Speech	+2	+1.5	+2	+3	+3	+2	+2	+2.2	1.5	+3.1	+4.2
Music		+4	+4	+4	+4	+4.0	+5	+4.2	1.0		
14. Condition 1+6-db-per-octave network											
Speech	−10	−10	−10	−8	−7	−10	−9	−9	3	−7	−6.3
Music		−7	−7	−5	−8	−4	−4	−5.5	4		

tensities for speech differ materially from those for music for cutoff frequencies above 1000 c.p.s. due to the differences in their energy spectra. The point at which the high-pass and low-pass curves intersect represents the frequency at which the loudness is equally divided. This intersection point for speech occurs at a frequency of 1000 c.p.s. The reduction in intensity at this point is about 5 db, whereas Munson's earlier data, obtained with headphone receivers, indicated a value of 10 db. The only apparent explanation for this difference is that the acoustic environment existing when listening to a loudspeaker in a room is quite different from that existing when headphone receivers are used.

The average curve from Fig. 10 has been replotted on Fig. 11 for comparison with similarly treated measured

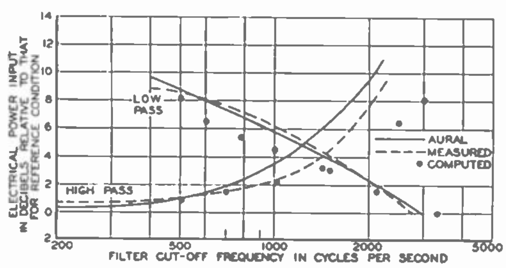

Fig. 11—Averaged power input versus spectrum range of speech or music for equal loudness in a room, as determined by aural, measured, and computed methods.

data and computed values. The computations were made using the applicable response-versus-frequency characteristics of Fig. 9 and the loudness-weighting characteristic of Fig. 6. The difference between the computed and measured data is due to the nonuniformity of the energy distribution with frequency in the sweep-frequency power applied as indicated in Fig. 7. From the curves, it may be concluded that the choice of this type of sweep-frequency power results in measured values of relative loudness which are in close agreement with the aural data.

Acoustic Power Requirements in Enclosures

When a loudspeaker projects sound into an enclosure, its acoustic performance as determined under open-air conditions is modified by the acoustic properties of the space, but the total power radiated is essentially unchanged. If the enclosure has very high absorption, the direct energy predominates and the characteristics of the loudspeaker will be similar to those for open air. The more live the room becomes, the more the reflected energy will predominate and, therefore, the greater will be the effect upon the radiation from the loudspeaker.

When a source of sound is started in a room, the energy spreads from the source and then strikes the various wall surfaces, where it is partially absorbed and partially reflected to other surfaces, where again it is partially absorbed and partially reflected. This process continues until the energy in the room builds up to a steady-state value, when the rate of absorption at the various surfaces and in the air is equal to the emission of energy from the source. At any point in the room, then, the energy density may be conveniently considered to be made up of two parts. One portion is contributed by direct radiation from the source and is equal to that which would be established at the point if the walls of the room were removed and the source were radiating into free space. The second portion is made up of energy which has been reflected one or more times from the various surfaces of the room. The first will be called the direct and the second the reverberant energy. The direct energy is distributed according to the inverse square law, while the reverberant may be considered as random in direction and uniform in distribution throughout the volume of the room.

The total energy in a room in the steady state is taken as[17]

$$\rho_{av}V = \frac{4EV}{\alpha S_R c} \tag{12}$$

in which

V = volume of the room
ρ_{av} = average energy density
E = rate at which the source emits energy
α = average absorption coefficient for the surfaces of the room
S_R = the total surface area of the room
c = velocity of sound.

This energy is made up of the total reverberant energy, assumed uniform in distribution, and the total direct energy. The reverberant energy is obtained by subtracting from (12) the total direct energy which depends upon the directional characteristics of the source and the position of the source in the room. Thus, if the source radiates uniformly in all directions (a point source), the direct energy will be a maximum when the source is in the center of the room; while if the radiation is concentrated in a relatively small solid angle, the direct energy will be a maximum when the source is at the side of the room and the energy radiated toward the center. Let us assume that the direct energy is that contained in a sphere having the source at its center and a radius equal to the mean free path between reflections in the room.[18–20] Such a sphere will have a volume very nearly equal to that of the room for all rooms of reasonable proportions. The two volumes will be equal if the mean free path is taken as $0.63\sqrt[3]{V}$, while accepted values of the mean-free-path range from $0.63\sqrt[3]{V}$ to $4V/S_R$

[17] V. O. Knudsen, "Architectural Acoustics," John Wiley and Sons, New York, N. Y., 1932, p. 127.
[18] E. R. Eyring, "Reverberation time in 'dead' rooms," *Jour. Acous. Soc. Amer.*, vol. 1, pt. I, pp. 217–241; January, 1930.
[19] See p. 137 of footnote reference 17.
[20] E. H. Bedell, unpublished work.

If we use the value $4V/S_R$, the total direct energy contained in such a sphere is

$$\rho_d V = \frac{4EV}{S_R c} \tag{13}$$

where ρ_d is the average direct energy density. The total reverberant energy in the room is then

$$\rho_r V = \frac{4EV}{\alpha S_R c} - \frac{4EV}{S_R c} \tag{14}$$

where the average reverberant energy density[21] ρ_r is

$$\rho_r = \frac{4E}{S_R c}\left(\frac{1-\alpha}{\alpha}\right). \tag{15}$$

The direct energy density ρ_d due to radiation from a point source at a point distant r from the source is

$$\rho_d = \frac{E}{4\pi r^2 c}\,. \tag{16}$$

If the source does not radiate uniformly,

$$\rho_d = \frac{EQ}{4\pi r^2 c} \tag{17}$$

in which $Q = 4\pi/\Omega$, Ω being the solid angle of radiation which is related to the directivity.

The average energy density ρ_{av} at any point within the enclosure is the sum of the direct and reverberant energy density ρ_d and ρ_r.

$$\rho_{av} = \frac{E\cdot Q}{4\pi r^2 c} + \frac{4E(1-\alpha)}{S_R c\alpha} = \frac{E}{4\pi c}\left[\frac{Q}{r^2} + \frac{16\pi}{R}\right] \tag{18}$$

in which $\alpha S_R/1-\alpha$ is defined as the room coefficient R because of its flexible use in practical problems.

From (18) it is possible to determine the manner in which the average energy density varies with the distance from the source in enclosures having various room coefficients. It is also useful to determine this variation relative to an arbitrary open-air condition ($R=\infty$), using $r=1$ as a reference since it is intended to refer an axial-pressure measurement of a loudspeaker made under open-air conditions to that which would exist in an enclosure. Thus, the ratio of the average energy density in an enclosure relative to the reference open-air condition, expressed in decibels, is

$$\delta = 10\log_{10}\frac{\rho_{av_{enc}}}{\rho_{av_{\text{open air at } r=1}}} = 10\log_{10}\left[\frac{Q}{r^2} + \frac{16\pi}{R}\right]. \tag{19}$$

[21] H. F. Olsen and F. Massa, "Applied Acoustics," P. Blakiston's Son and Co., Philadelphia, Pa., second edition, 1939.

This relationship for point-source radiation ($Q=1$) is plotted on Fig. 12 for various values of room coefficient. These results are independent of the power radiated by the source. These curves show that it is possible to obtain a substantial increase in energy density in a room as compared to that for open air. Since point-source radiation was assumed in obtaining the curves of Fig. 12,

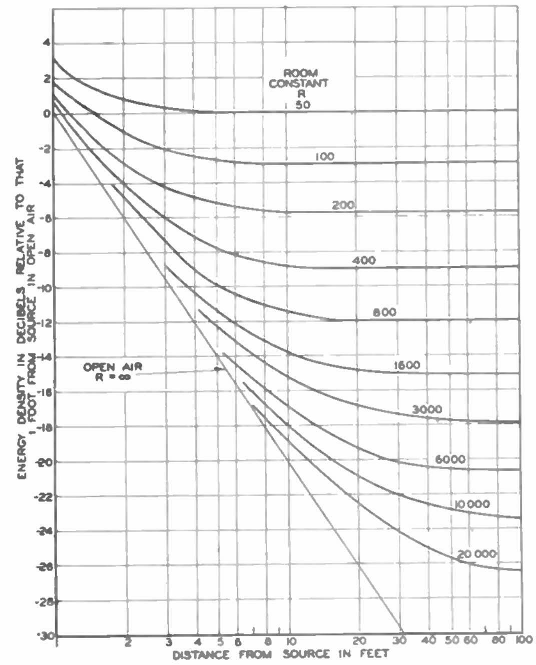

Fig. 12—Average energy density versus distance from the source for enclosures having various room coefficients relative to the energy density at 1 foot from the source in open air, expressed in decibels; point-source radiation assumed.

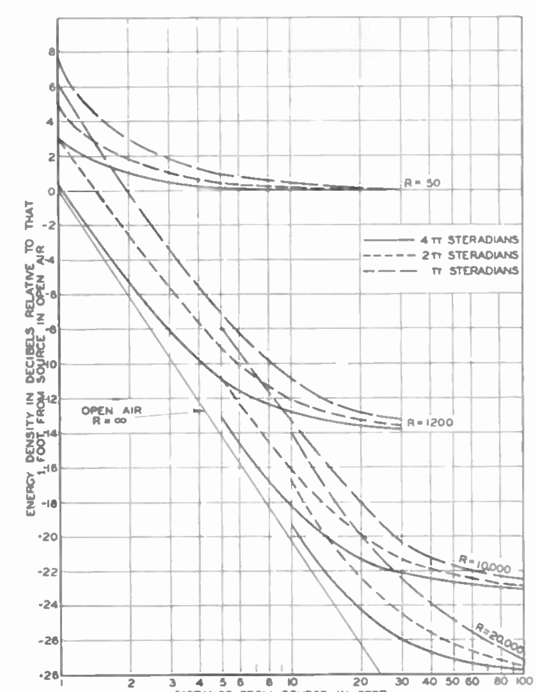

Fig. 13—Curves showing the effect of the directivity of the source on the average energy density in enclosures having various room coefficients.

the effects of the directivity of the source on these relationships must be determined before applying the results to practical conditions.

The effect of the directivity of the source may be determined readily by assuming various values of Q in (19). Curves illustrating this effect have been plotted on Fig. 13, for values of Q of 1, 2, and 4 for a range of room coefficients. Values of Q greater than 4 are not likely to be encountered in practice. It will be observed from the curves that the effect of directivity upon the energy density is small for distances of 30 feet or more, except for extremely large enclosures where R is large.

Consideration of this data indicates the fact that, if a representative distance from the source is selected, a gain in energy density over that for open air can be determined. It is observed that, beyond distances of 10 feet in small rooms and 30 feet in large rooms, the energy density remains constant and is practically all reflected energy. This fact is important because it permits the evaluation of the intensity throughout an enclosure from a single point observation. This suggests the use of 30 feet as a reference distance from the source. If the axial-pressure measurement p_{ax} of a loudspeaker under open-air conditions is made at a distance of 30 feet or corrected to the value that would exist at that distance, a room gain factor K_2, may be computed for any enclosure. This factor represents the gain in intensity level that would exist in an enclosure relative to that measured in open air at a distance of 30 feet for a given available power input.

The room gain factor may be computed as a function of room volume if it is assumed that all enclosures have an optimum reverberation time. The optimum reverberation time of enclosures has been determined by many investigators.[22–25] Average values of these data are shown on Fig. 14. It has been established that the shape of the room will have a negligible effect upon the results.

The value of K_2 may be obtained directly from Fig. 12 if the room coefficient R is known, or it may be computed in the following manner:

According to Eyring,[18] the reverberation time T of an enclosure in seconds is

$$T = \frac{0.05V}{-S_R \ln(1-\alpha)} \tag{20}$$

in which

V = volume of the room in cubic feet
S_R = the total surface area of the room in square feet
α = the average absorption coefficient.

[22] Watson Architecture, May, 1927.
[23] S. Lifschitz, "Acoustics of large auditoriums," *Jour. Acous. Soc. Amer.*, vol. 4, pp. 112–121; October, 1932.
[24] P. E. Sabine, "Acoustics of sound recording rooms," *Trans. Soc. Mot. Pic. Eng.*, vol. 12, pp. 809–813; September, 1928.
[25] W. A. MacNair, "Optimum reverberation time for auditoriums," *Jour. Acous. Soc. Amer.*, vol. 1, pt. 1, pp. 242–248; January, 1930.

Letting $\Delta = 0.05\,V/S_R$ and substituting in (20),

$$\alpha = 1 - e^{-\Delta/T}. \tag{21}$$

Since $R = \alpha S_R/1-\alpha$, by substitution

$$R = S_R(e^{\Delta/T} - 1). \tag{22}$$

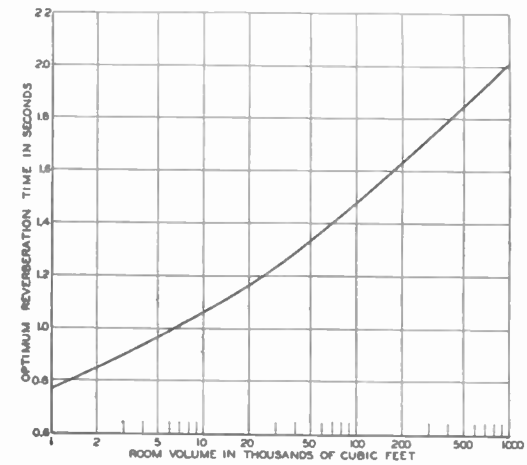

Fig. 14.—Optimum reverberation time as a function of room volume (512 c.p.s.).

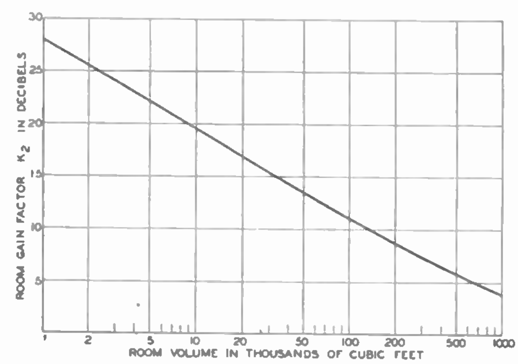

Fig. 15—Room gain factor for rooms having optimum reverberation times.

From Fig. 12, the ratio of the energy density at 30 feet to that at 1 foot in open air expressed in decibels is

$$L_0 = 10 \log_{10} \frac{1}{30^2} \tag{23}$$

$$= -29.54 \text{ db.}$$

From (19), (22), and (23) it is readily shown that, at a reference distance of 30 feet, the total gain in the energy density K_2 due to moving the source from open air to an enclosure is

$$K_2 = 10 \log_{10}\left[\frac{Q}{r^2} + \frac{16\pi}{S_R(e^{\Delta/T} - 1)}\right] - L_0. \tag{24}$$

If only the first three terms of the exponential series for e^x are employed, (24) reduces to

$$K_2 = 10 \log_{10}\left[\frac{Q}{r^2} + \frac{1000T}{V\left[1 + \frac{0.05V}{2S_R T}\right]}\right] - L_0. \quad (25)$$

The error introduced by this approximation will be a maximum of 10 per cent for a room volume of 10^6 cubic feet, and will be only 1 per cent for a room volume of 10^5 cubic feet. Values of K_2, computed from (25) for optimum reverberation time, are shown on Fig. 15. If the reverberation time differs from optimum by ± 30 per cent, the error will be 1.4 db. If the reverberation time is known, an approximate correction of the room factor may be obtained by assuming the acoustic power to be inversely proportional to the reverberation time for a given room volume.

It is now necessary to establish representative sound levels for the reproduction of speech and music in enclosures. Since Fletcher[6] has tabulated the required data, only a brief summary of this information is given in Table III.

As indicated in columns 1, 2, and 3 of this table, maximum peaks of speech or music are about 20 db above the long r.m.s. power indicated by a volume indicator or sound meter, while the maximum r.m.s. power is about 10 db above the volume-indicator value. Amplifier design is frequently based on the maximum r.m.s. power since the peaks that exceed this value are of short duration and occur during a very small percentage (1 to 5 per cent) of the time. The data for conversational speech were obtained at a distance of 20 feet and converted to the levels existing at 2½ feet, as shown in columns 4 and 5. An adequate speech level to be established within the enclosure has been selected as the level existing at 2½ feet from the lips of a person talking conversationally, as indicated in columns 6 and 7. The levels shown in column 6 of the table are applicable for amplifier design. From the table it is evident that a maximum r.m.s. intensity level of 78 db for speech, 96 db for small orchestras, and 106 db for large orchestras should be established in an enclosure for adequate reproduction.

From the above data, the acoustic power required for a specified intensity level in any enclosure may be computed. The maximum acoustic power radiated from a point source throughout a solid angle of 4π steradians in open air may be determined from (4) and (5). For the required intensity levels of 106, 96, and 78 db, the maximum r.m.s. acoustic power W_{a_1} is 41.2, 4.12, and 0.065 acoustic watts, respectively. Applying the room gain factor K_2, the maximum acoustic power W_{a_2} required for the desired levels of speech and music for any room volume having an optimum reverberation may be obtained from

$$W_{a_2} = W_{a_1} 10^{(K_2/10)}. \quad (26)$$

Values so computed are shown on Fig. 16. The optimum reverberation time at a frequency of 512 cycles was used in order to simplify the computation of power.

However, MacNair[25] has shown that, for the loudness of all pure tones to decay at the same rate as the sensation level at all frequencies, which is our premise, the optimum reverberation time for a given enclosure must change with frequency. It is constant between frequencies of 700 and 4000 cycles and about one-half of this value at a frequency of 100 cycles. Using MacNair's data and the midband frequencies of the equal loudness increments (Fig. 3), the values of K_2 were recomputed for various room volumes, and found to be within one-half of a decibel of the values obtained for a frequency of 514 cycles. Therefore, the ordinates of Fig. 16 are an adequate indication of required power based on loudness.

TABLE III

SOUND LEVELS FOR SPEECH AND MUSIC

	Data Extrapolated to 30 Feet from Measurements at a 20-Foot Distance Expressed in Decibels Relative to 10^{-16} watts/cm.^{2}			Desired Levels within Enclosure			
				Speech Levels Normal for 2.5-Foot Distance			
Type of sound	Maximum peak intensity level	Amplifier design maximum r.m.s. ⅛ second	Volume indicator reading long r.m.s.	Recommended for amplifier design	Volume indicator	Recommended for amplifier design	Volume indicator
	(1)	(2)	(3)	(4)	(5)	(6)	(7)
Conversational speech							
Men	66.5	56.5	46.5	78	68	78	68
Women	64.5	54.5	44.5	76	66		
Music							
1 voice	97	87	77			87	77
100 voices	117	107	97			107	97
75-piece orchestra	116	106	96			106	96
18-piece orchestra	106	96	86			96	86

In many cases, reproduction may be required in noisy places. It is always desirable to maintain the signal-to-noise ratio a maximum. However, when the noise level is very high (90 to 100 db), a signal-intensity level of 10 db above the noise level[26] is sufficient for adequate intelligibility of speech, in which case the 78-db noise level assumed for speech reproduction may have to be increased. The acoustic power required for this condition will then exceed the values shown on Fig. 16 by an equivalent amount.

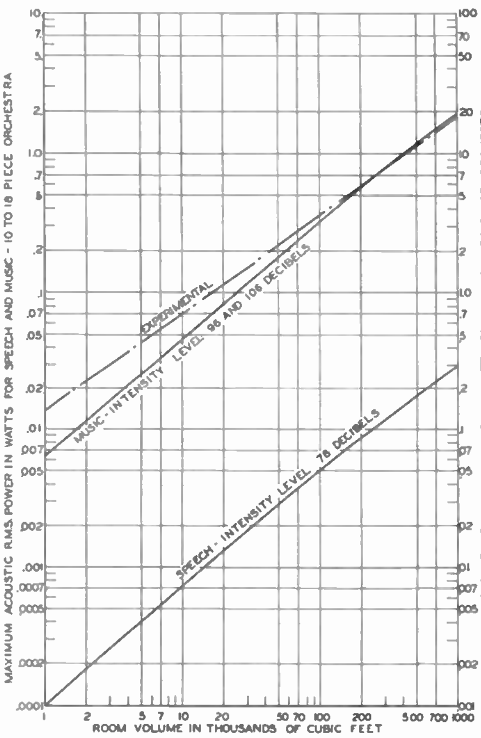

Fig. 16—Maximum r.m.s. acoustic power as a function of room volume for intensity levels of 78 db for speech and 96 and 106 db for music.

The solid curves of Fig. 16 show the computed relationship between room volume and the acoustic power required to reproduce speech and music at intensity levels of 78, 96, and 106 db. The curve indicated by the broken line of the figure represents the listening judgment of many observers on the basis of satisfactory or "pleasing" sound levels. It will be observed that close agreement between computed and empirical data exists at larger room volumes. The agreement is found to be somewhat poorer in the case of small rooms, where a maximum deviation of 3 db occurs. Since the empirical data is based on personal judgment, it is difficult to reconcile these differences. The computed levels of 78, 96, and 106 db for speech and music would appear to be at least adequate. In determining power requirements, however, it may be well to bear in mind that in small rooms a somewhat higher power may be necessary for satisfactory psychological effects. Fortunately, the deviation becomes appreciable only in small rooms where relatively little power is needed.

Amplifier Power

The amplifier capacity required for various enclosures determined by converting the ordinate of Fig. 16 to electrical watts for various values of the loudness-efficiency factor LR are shown on Figs. 17 and 18. The

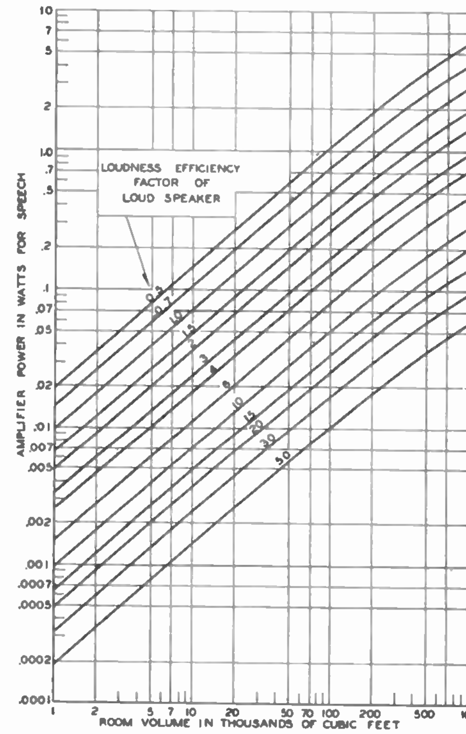

Fig. 17—Required amplifier power for speech as a function of room volume for various loudness-efficiency factors.

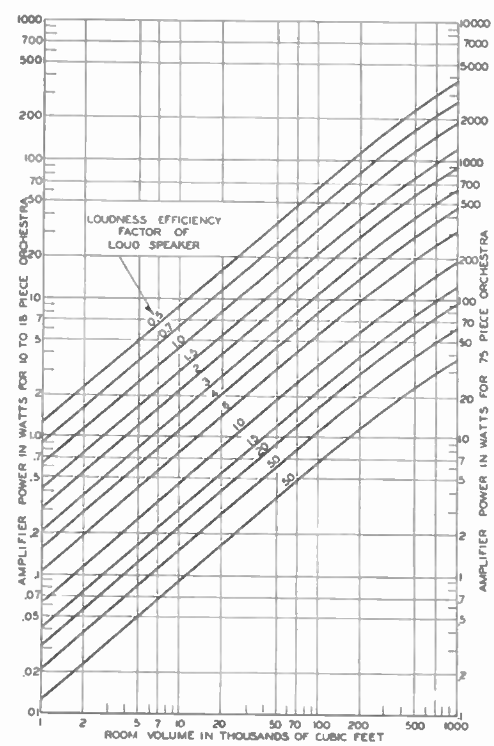

Fig. 18—Required amplifier power for music as a function of room volume for various loudness-efficiency factors.

[26] N. R. French, unpublished data.

number of loudspeakers required may be determined by dividing the required amplifier power by W_{ec}, the power capacity of the loudspeaker.

Application

The following illustrative example may serve to clarify the application of the foregoing proposals. A power source having a sweep-frequency rate of 5 to 10 per second and covering the frequency range from 300 to 3300 cycles, equalized in accordance with the characteristic shown on Fig. 6, is assumed. An open-circuit voltage of 31 volts is applied to the test circuit of a 12-inch direct-radiator loudspeaker having a rating impedance of 8 ohms. The available power will then be 30 watts, which is assumed to be the power-capacity rating of the loudspeaker. Under this condition, the sound pressure at 6 feet on the axis of the speaker is measured in open air and found to be 50 dynes per square centimeter. The intensity level as indicated by a sound-level meter would be 108 db. From these data it is now possible to determine the loudness-efficiency factor of the loudspeaker, the amplifier power necessary for the prescribed levels of reproduction for speech and music in any enclosure, and the number of loudspeakers required.

A pressure of 50 dynes per square centimeter obtained at an axial distance of 6 feet corresponds to an intensity level of 94 db relative to 10^{-16} watts/cm.² when extrapolated for a distance of 30 feet. Since W_{ec} is 30 watts, $k=14.78$ db relative to 1 watt, and K_1 from Fig. 26 in the Appendix is 6.8 db, substituting in (9), (10), and (11).

$$\begin{aligned} L_e &= L_{ax} - 89.8 - k - K_1 \\ &= 94 - 89.8 - 14.8 - 6.8 \\ &= -17.4 \text{ db} \\ W_{Le} &= 10^{-17.4/10} \\ &= .0183 \\ LR &= 100 \times .0183 \\ &= 1.83 \text{ per cent.} \end{aligned}$$

For the reproduction of speech and music in a room having a volume of 10^6 cubic feet and an optimum reverberation time, Figs. 17 and 18 indicate that, when LR is 1.83 per cent, an amplifier power of 980 watts is required for music from large orchestras, 98 watts for small orchestras, and 1.53 watts for speech. One loudspeaker is sufficient for the reproduction of speech, three loudspeakers for music from small orchestras, while 33 loudspeakers are required when music from large orchestras is reproduced.

The application of these results would require that manufacturers of loudspeakers make certain measurements to determine the loudness rating of their product. Such a determination would require that the following conditions be met:

(1) Tests should be made in open air or in a dead room without reflections above 300 cycles.

(2) The pressure measurement should be made on the geometric axis of the loudspeaker at a distance that is at least three times the maximum transverse dimension of the radiating area. The resulting pressure should be corrected to that which would exist at a distance of 30 feet, employing the inverse-square law.

(3) The electrical supply for the test should be a 300- to 3300-cycle sweep-frequency tone, with the designated weighting equalizer in the circuit. The sweep-frequency source should have a reciprocating linear frequency change with time at a rate of 5 to 10 times per second to obtain an amplitude distribution of the components in accordance with that indicated on Fig. 7.

(4) The pressure p_{ax}, or the intensity level L_{ax}, should be obtained on the axis with a power supply sufficient to drive the loudspeaker at its rated power capacity. If powers lower than W_{ec} are used in making this measurement, the appropriate correction must be made in the formulas.

(5) The loudness rating LR of the loudspeaker may then be obtained from (9), (10), and (11).

It must be recognized that, in addition to loudness, many other factors must be considered in establishing a true merit rating for loudspeakers. Uniformity of frequency response, harmonic distortion, frequency range, intermodulation, damping, and uniformity of distribution all have their effects on the performance of an instrument. These factors are controlled by basic instrument design, and their magnitudes are established by laboratory measurements. Listening tests, if carefully performed, provide a practical method of evaluating the extent to which the factors have been controlled and the suitability of an instrument for its intended use. Instruments for special or scientific uses require, of course, more careful selection.

In spite of the various compromises and assumptions which had to made to arrive at a practical and simple factor for rating loudspeakers, it is believed that the proposed method should give a reasonably accurate measure of effective loudness efficiency. Its use in practice should materially simplify the problems of the sound-systems engineer.

APPENDIX

Derivation of Loudness-Directivity Index

Typical loudspeaker systems used in practice and the shape of their radiating areas are given in Table IV. The theoretical condition assumed to approximate each practical condition is also shown.

In the analysis of the theoretical conditions, it is assumed that the radiating surface vibrates axially, and that the distance from the source at which the acoustic power is computed is sufficient to insure that the pressure and particle velocity are in phase.

The directivity index may be derived as follows:

Referring to Fig. 19, consider the center of the radiating area to be located at the origin O. At a given distance r the effective pressure p and the particle velocity ξ are in phase. For this condition, p is equal to $\rho c\xi$ where ρ is the density of the medium and c is the velocity of sound.

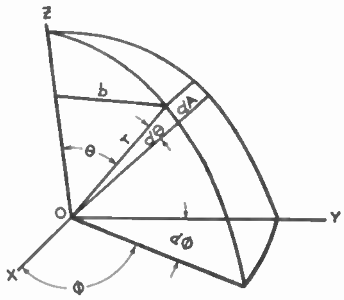

Fig. 19—Source of radiation.

TABLE IV

TYPES OF RADIATION FROM LOUDSPEAKERS

Practical Condition		Theoretical condition assumed to approximate practical condition
Type of loudspeaker	Approximate shape of radiating area	
A. *Baffled*		
Single Unit		
Horn	Circle	Rigid disk in infinite baffle
Direct radiator	Circle	Rigid disk in infinite baffle
Direct radiator on horn	Rectangle	Rigid rectangular plate in infinite baffle
Sectoral horn	Cylindrical sector	Sectoral radiation in infinite baffle
Multicellular horn	Rectangular spherical sector	Uniform radiation over equivalent solid angle in an infinite baffle
Two Unit		
Direct radiator on horn+sectoral horn	Rectangle+cylindrical sector	Rigid rectangular plate +sectoral radiation in infinite baffle
Direct radiator +sectoral horn	Circle+cylindrical sector	Rigid disk+sectoral radiation in infinite baffle
Direct radiator on horn+multicellular horn	Rectangle+rectangular spherical sector	Rigid rectangular plate +uniform radiation over equivalent solid angle in infinite baffle
Direct radiator +multicellular horn	Circle+rectangular spherical sector	Rigid disk+uniform radiation over equivalent solid angle in infinite baffle
B. *Unbaffled*		
Horn or direct radiator in small enclosure	Circle	Rigid disk set in sphere

The power radiated from a vibrating area in the XY plane, located at the center of a sphere, is $p^2/\rho c$ per unit area of the spherical surface. Then the power flowing through an elementary area dA of the spherical surface is

$$dP = \frac{p^2}{\rho c} dA. \tag{27}$$

From Fig. 19, the elemental spherical surface is

$$\begin{aligned} dA &= r \sin\theta d\Phi \cdot r d\theta \\ &= r^2 \sin\theta d\theta d\Phi. \end{aligned} \tag{28}$$

If the radiating area is symmetrical about the Z axis, the total power transmitted through the spherical surface is then

$$\begin{aligned} P_t &= \frac{r^2}{\rho c}\int_0^{2\pi} d\Phi \int_0^{\pi} p^2 \sin\theta d\theta \\ &= \frac{2\pi r^2}{\rho c}\int_0^{\pi} p^2 \sin\theta d\theta \end{aligned} \tag{29}$$

where p is the effective pressure at any angle θ.

Then the directivity index for radiation over a hemisphere (baffled condition) is

$$D.I._h = -10 \log_{10} \frac{\frac{2\pi r^2}{\rho c}\int_0^{\pi/2} p^2 \sin\theta d\theta}{\frac{2\pi r^2}{\rho c}\int_0^{\pi} p_{ax}^2 \sin\theta d\theta} \tag{30}$$

where p_{ax} is the axial pressure. Since p_{ax} may be considered constant, the denominator becomes $(4\pi r^2/\rho c)p_{ax}^2$. Let p_θ be the ratio of the pressure at any angle θ relative to p_{ax}; then

$$D.I._h = -10 \log_{10} \frac{1}{2}\int_0^{\pi/2} p_\theta^2 \sin\theta d\theta. \tag{31}$$

Then the directivity index for radiation over a sphere (unbaffled condition) is

$$D.I._s = -10 \log_{10} \frac{1}{2}\int_0^{\pi} p_\theta^2 \sin\theta d\theta. \tag{32}$$

The directivity index for the types of radiation assumed in Table IV may now be derived.

Rigid Disk in Infinite Baffle

When a rigid disk located in an infinite rigid baffle vibrates axially in a free fluid, the pressure ratio p_θ of the pressure in space at any angle θ from the normal to that existing at an equal distance on the axis, has been shown by Stenzel[9,10] to be

$$p_\theta = \frac{2J_1(ka \sin\theta)}{ka \sin\theta} \tag{33}$$

when the distance from the disk is at least five times the disk diameter.

$$J_1 = \text{Bessel's function of the first order}$$

$$ka = \frac{\pi df}{c} = 2.32 \cdot df \cdot 10^{-4}$$

where d = diameter of disk in inches
c = velocity of sound in inches per second
f = frequency in cycles per second.

Substituting (33) in (31) and integrating,[27] the directivity index for a rigid disk in a baffle is

$$D.I._h = -10 \log_{10} \frac{1}{(ka)^2}\left[1 - \frac{J_1(2ka)}{ka}\right]. \quad (34)$$

Rigid Rectangular Plate in Infinite Baffle

In this case, (31) for the directivity index cannot be used because the pressure is not uniform on a circle about the axis of the plate in any plane parallel to the plane of the plate. McLachlan[28] has shown that, when a rigid rectangular plate of length $2a$ and width $2b$ vibrates axially in an infinitely rigid baffle in the XZ plane, the pressure $p(r, \theta, \Phi)$ in space at an angle θ with respect to the Z axis and at an angle Φ with respect to the X axis in the XY plane, at a distance r from the origin, is

$$p(r, \theta, \Phi) = \frac{2\rho\ddot{\xi}ab}{r}\left[\frac{\sin(ka \sin\theta \cos\Phi)}{ka \sin\theta\cos\Phi}\right] \cdot \left[\frac{\sin(kb\cos\theta)}{kb\cos\theta}\right] \quad (35)$$

where ρ is the density of the medium and $\ddot{\xi}$, the acceleration of the plate. The axial pressure p_{az} on the Y axis is obtained when $\theta = \pi/2$ and $\Phi = \pi/2$, and is

$$p_{az} = \frac{2\rho\ddot{\xi}ab}{r}.$$

Therefore, the directivity index is

$$D.I._h = -10 \log_{10} \frac{\int_0^\pi \int_0^\pi p^2(r, \theta, \Phi) \cdot r^2 \sin d\theta d\Phi}{4\pi r^2 p_{az}^2}$$

$$= -10 \log_{10} \frac{1}{4\pi} \int_0^\pi \int_0^\pi \left[\frac{\sin(ka\sin\theta\cos\Phi)}{ka\sin\theta\cos\Phi}\right]^2 \cdot \left[\frac{\sin kb\cos\theta}{kb\cos\theta}\right]^2 \sin\theta d\theta d\Phi. \quad (36)$$

In an unpublished memorandum, C. T. Molloy has shown that (36) may be transformed into

$$D.I._h = -10 \log_{10} \frac{1}{ka} \int_0^{\pi/2} M(ka \sin\theta) \cdot \left[\frac{\sin(kb\cos\theta)}{kb\cos\theta}\right]^2 d\theta \quad (37)$$

in which

$$M(ka\sin\theta) = \frac{1}{2}\left[\int_0^{2ka\sin\theta} J_0(\lambda)d\lambda - J_1(2ka\sin\theta)\right].^{29}$$

The values of $D.I._h$ used in this paper were obtained by a "Simpson's Rule" numerical integration of (37).

Sectoral Radiation in an Infinite Baffle

For the case of radiation from a sectoral horn, rigorous analytical treatment is extremely difficult. With certain assumptions, however, a derivation which approximates the practical condition can be obtained. A radiating area in an infinite baffle was assumed to have a radiation pattern in which the pressure throughout the sectoral angle α is constant over an arc in any given plane parallel to the XY plane, and is zero outside the angle α on this arc. In the vertical direction, the radiation pattern was assumed to be that of a line radiator of length l lying on the z axis with its center at the origin. Further unpublished work of C. T. Molloy has shown that the total power P radiated at a distance r from the source is

$$P = \frac{\alpha r p_{az}^2}{\rho c} \int_0^\pi \left[\frac{\sin ka\cos\theta}{ka\cos\theta}\right]^2 \sin\theta d\theta \quad (38)$$

where θ is the angle between a radius vector from the origin to a field point, and the Z axis and $a = l/2$ and the power from a point source producing the same axial pressure is

$$P_{az} = \frac{4\pi r^2}{\rho c} p_{az}^2. \quad (39)$$

The directivity index is, therefore,

$$D.I._h = -10 \log_{10} \frac{\alpha}{4\pi} \int_0^\pi \left[\frac{\sin(ka\cos\theta)}{ka\cos\theta}\right]^2 \sin\theta d\theta, \quad (40)$$

which has been shown by Molloy to reduce to

$$D.I._h = -10 \log_{10} \frac{\alpha}{2\pi}\left[2\frac{Si(2ka)}{(2ka)} - \left(\frac{\sin ka}{ka}\right)^2\right].^{30} \quad (41)$$

[27] N. W. McLachlan, "Bessel Functions for Engineers," Oxford Press, New York, N. Y., 1934; p. 98.

[28] N. W. McLachlan, "Loudspeakers," Oxford Press, New York, N. Y., 1934; p. 101.

[29] *Method of Computation of* $\int_0^{2\alpha} J_0(\lambda)d\lambda$: In the interval $0 \leq \alpha \leq 5$ this function is tabulated in "Table of Integrals $\int_0^x J_0(t)dt$ and $\int_0^x Y_0(t)dt$" by A. N. Lowan and Milton Abramowitz, *Jour. Math. and Phys.*, vol. 22, May, 1943. In the interval $5 \leq \alpha \leq 25$, $\int_0^{2\alpha} J_0(\lambda)d\lambda =$

$$\int_0^{10} J_0(\lambda)d\lambda - \sqrt{2}[C(10)+S(10)] + \sqrt{2}[C(2\alpha)+S(2\alpha)]$$

where C and S are Fresnel Integrals, and are tabulated in "Functions and Tables" by E. Jahnke and F. Emde, p. 35, 1943, Dover Publications. In the interval $25 \leq \alpha$, use the same formula as for preceding interval, but compute C and S by the following asymptotic formulas:

$$C(2\alpha) = \frac{1}{2} + \frac{\sin(2\alpha)}{\sqrt{4\pi\alpha}} - \frac{\cos(2\alpha)}{4\alpha\sqrt{4\pi\alpha}}$$

$$S(2\alpha) = \frac{1}{2} - \frac{\cos(2\alpha)}{\sqrt{4\pi\alpha}} - \frac{\sin 2\alpha}{4\alpha\sqrt{4\pi\alpha}}.$$

The function $J_1(2\alpha)$ may be found in "British Association for the Advancement of Science Mathematical Tables, vol. VI, Bessel Functions," University Press, Cambridge, 1937.

[30] The integral sines, Si, are tabulated in Jahnke and Emde, p. 6.

If $\alpha=\pi$, the directivity index for a line radiator of length $2a$ results; thus,

$$D.I._h = -10 \log_{10} \frac{1}{2}\left[2\frac{Si(2ka)}{(2ka)} - \left(\frac{\sin ka}{ka}\right)^2\right]. \qquad (42)$$

When a is zero and α is equal to π, the radiator becomes a point source in a baffle. Then $D.I._h=3$ db.

Uniform Radiation Over a Portion of a Spherical Zone

For this case, the pressure is assumed uniform over an area of a sphere, intercepted by two planes at an angle α passing through the Z axis and two planes at an angle β passing through the X axis.

Referring to (29), the total power transmitted through this surface is

$$P_t = \frac{4p^2r^2}{\rho c}\int_{\pi/2-\alpha/2}^{\pi/2} d\phi \int^{\pi/2 \sin\theta d\theta} \cot^{-1}(\tan(\beta/2)\sin\phi) \qquad (43)$$

Integration of (43) yields:

$$P_t = \frac{4p^2r^2}{\rho c}\cdot\sin^{-1}\left[\sin\frac{\alpha}{2}\cdot\sin\frac{\beta}{2}\right]. \qquad (44)$$

The power from a point source producing the same axial pressure is

$$P_{ax} = \frac{4\pi r^2 p_{ax}^2}{\rho c}. \qquad (45)$$

Therefore, the directivity index is

$$D.I._h = -10\log_{10}\frac{1}{\pi}\sin^{-1}\left[\sin\frac{\alpha}{2}\cdot\sin\frac{\beta}{2}\right]. \qquad (46)$$

It is observed that the directivity index for this case is independent of frequency.

Piston Set in Sphere

In this problem, the directivity index may be obtained from (32). Morse[31] has established the equations for the pressure distribution in space resulting from a rigid piston vibrating axially, set in a sphere of radius a, and subtending an angle $2\theta_0$. Molloy has extended this work to include the actual distribution for a number of specific cases.

The spatial pressure is

$$|p(r,\theta)| = \frac{\mu_0}{2}\frac{(\rho c)}{(kr)}L(\theta) \qquad (47)$$

in which μ_0 = the maximum radial velocity of the piston, and the maximum axial pressure is

$$|p_{\max}| = \frac{\mu_0}{2}\frac{\rho c}{kr}\cdot L(0) \qquad (48)$$

[31] P. M. Morse, "Vibration and Sound," McGraw-Hill Book Co., New York, N. Y., 1936.

in which

$$L(\theta) = \frac{2}{\mu_0}\left|\sum_{m=0}^{\infty}\frac{U_m}{D_m}e^{i[\delta_m-(m+1)\pi/2]}\cdot P_m(\cos\theta)\right|$$

$$U_m = \frac{\mu_0}{2}\left[P_{m-1}(\cos\theta_0) - P_{m+1}(\cos\theta_0)\right]$$

$$D_m = \frac{1}{2m+1}\left\{\left[mj_m(ka) - (m+1)j_{m+1}(ka)\right]^2 + \left[mn_m(ka) - (m+1)n_{m+1}(ka)\right]^2\right\}^{1/2}.$$

$$\tan\delta_m = \left(\frac{mj_{m-1}(ka) - (m+1)j_{m+1}(ka)}{mn_{m-1}(ka) - (m+1)n_{m+1}(ka)}\right)$$

$P_m(\cos\theta_0)$ = Legendre Polynomial of m^{th} order

$$j_m(ka) = \sqrt{\frac{\pi}{2ka}}\cdot J_{(m+1/2)}(ka)$$

$$n_m(ka) = -1^{m-1}\sqrt{\frac{\pi}{2ka}}\cdot J_{-(m+1/2)}(ka)$$

$$p_\theta = \frac{1}{(L(0))}\cdot L(\theta)). \qquad (49)$$

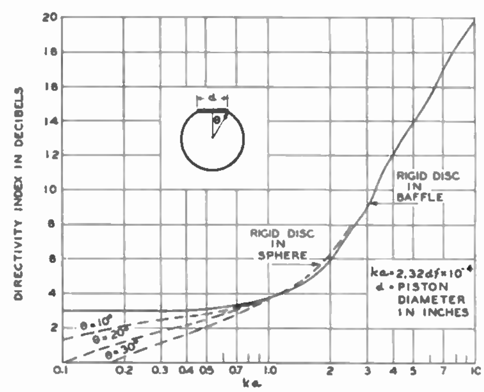

Fig. 20—Directivity index of a rigid disk vibrating axially in an infinite baffle and in a sphere.

Then, substituting in (32), the directivity index is

$$D.I._S = -10\log_{10}\frac{1}{2L^2(0)}\int_0^\pi L^2(\theta)\sin\theta d\theta \qquad (50)$$

or, if $\sigma(\theta)=L^2(\theta)$,

$$D.I._S = -10\log_{10}\frac{1}{2\sigma_0}\int_0^\pi \sigma(\theta)\sin\theta d\theta. \qquad (51)$$

Molloy has also shown that this may be reduced to

$$D.I._S = -10\log_{10}\frac{S_p k^2}{\pi\sigma_0}\cdot R_s \qquad (52)$$

in which

S_p = piston area in square centimeters

R_s = radiation resistance in $\dfrac{\text{dynes/cm.}^2}{\rho c\ \text{cm./sec.}}$

$k=\omega/c$.

Equation (51) may be evaluated by numerical integration.

The relation between the directivity index and the argument ka for each of the theoretical radiation conditions is shown on Figs. 20 through 24. These curves reveal interesting relationships between the various conditions of radiation.

Referring to Fig. 20, on which the directivity indexes for a rigid disk in a baffle and in a sphere are plotted, it is observed that, when ka exceeds unity, the directivity indexes are approximately the same for the two conditions. For lower values of ka, the directivity indexes for the rigid disk in a sphere approach those for the baffled condition when θ becomes very small. It is apparent from Fig. 21 that the curve shape for the rectangular

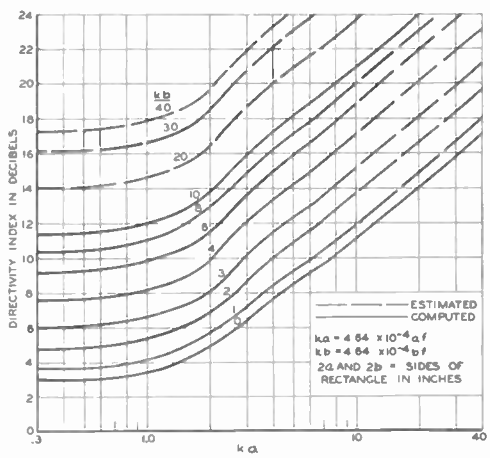

Fig. 21—Directivity index for radiation from a rigid rectangular plate in an infinite baffle.

plate is similar to that for the circular disk. The results of a further investigation of this similarity are shown on Fig. 22, where the directivity indexes are plotted in terms of radiating area for both the disk and the plate. It is observed that the directivity indexes for the circle, square, and rectangle are approximately the same for a given area. These data were computed for a frequency of 1000 c.p.s., but this conclusion applies at any frequency. From Fig. 23, which shows the directivity indexes for sectoral radiation, it should be noted that, when α is equal to 180 degrees, the directivity indexes

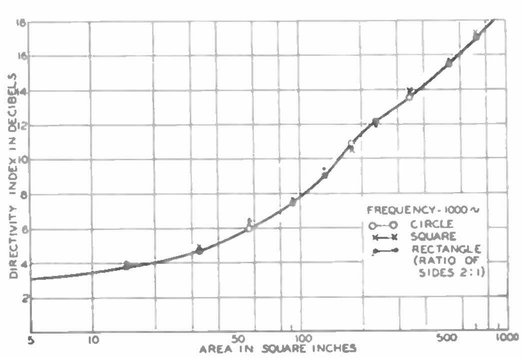

Fig. 22—Directivity index for radiation from a rigid disk and a rectangular plate in an infinite baffle as a function of area.

apply for radiation from a line of length l. Fig. 24 presents the directivity indexes for radiation from a portion of a spherical zone, a condition which approximates the multicellular horn. Since two variables, α and β, are involved, some simplification in succeeding computations

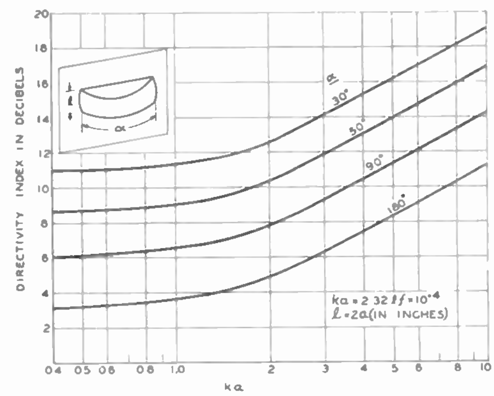

Fig. 23—Directivity index for sectoral radiation in an infinite baffle.

may be attained by expressing the directivity index in terms of the equivalent solid angle of radiation, values for which are shown on the figure.

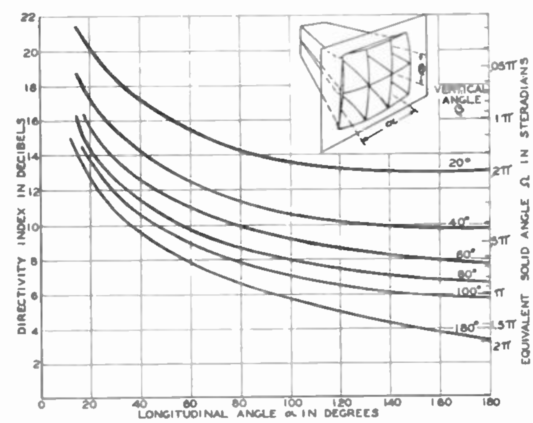

Fig. 24—Directivity index from a portion of a spherical zone in an infinite baffle.

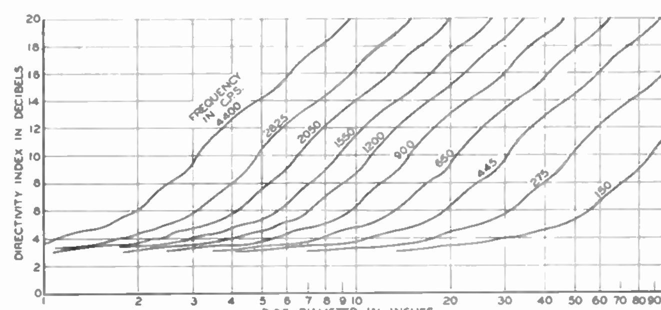

Fig. 25—Directivity indexes of rigid disks of various diameters in an infinite baffle at each of the midfrequencies of the ten equal-loudness bands.

The above data provide a basis for determining the loudness directivity indexes of the various loudspeakers listed in Table IV. The loudness-directivity index of

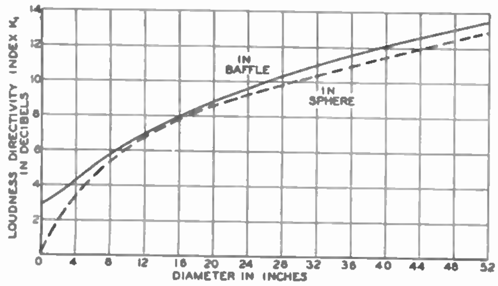

Fig. 26—Loudness-directivity index for radiation from a rigid disk in a sphere and in an infinite baffle (circular horn or direct radiator, baffled or unbaffled).

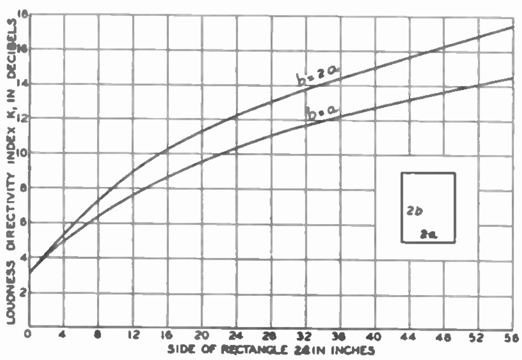

Fig. 27—Loudness-directivity index for radiation from a rigid rectangular plate in an infinite baffle (rectangular horn, baffled).

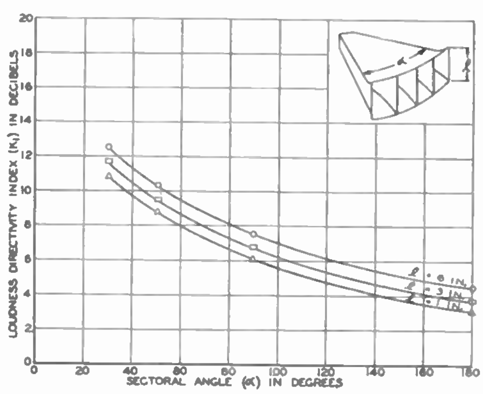

Fig. 28—Loudness-directivity index of sectoral horn.

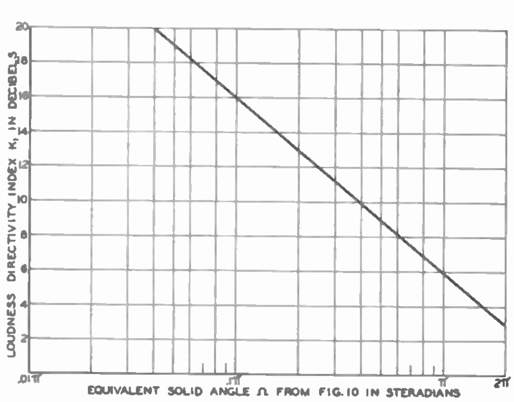

Fig. 29—Loudness-directivity index for radiation from a portion of a spherical zone in an infinite baffle (multicellular horn, baffled).

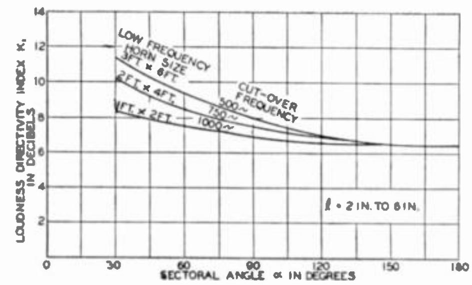

Fig. 30—Loudness-directivity index for a dual system involving sectoral radiation and radiation from a rigid rectangular plate in an infinite baffle (rectangular low-frequency horn and sectoral high-frequency horn, baffled).

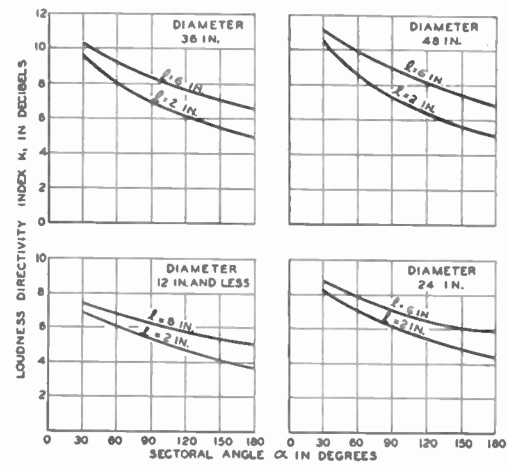

Fig. 31—Loudness-directivity index for a dual system involving sectoral radiation and radiation from a rigid disk in an infinite baffle (circular low-frequency direct radiator and sectoral high-frequency horn, baffled).

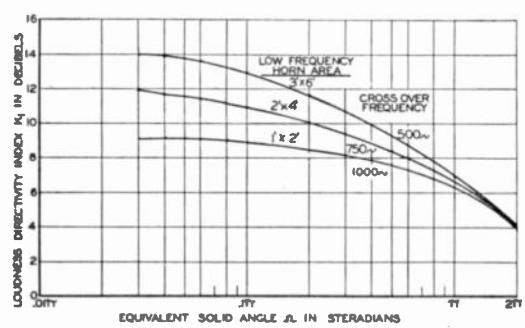

Fig. 32—Loudness-directivity index for a dual system involving radiation from a rigid rectangular plate and a portion of a spherical zone in an infinite baffle (rectangular low-frequency horn and a multicellular high-frequency horn, baffled).

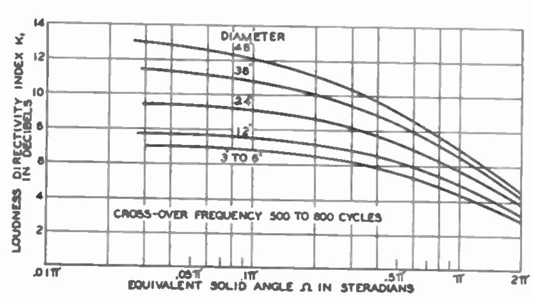

Fig. 33—Loudness-directivity index for a dual system involving radiation from a rigid disk and a portion of a spherical zone in an infinite baffle (circular low-frequency direct radiator and a multicellular high-frequency horn, baffled).

any radiating device may be defined as the average of the directivity indexes for the ten midfrequencies of the equal loudness bands, and may be expressed as follows:

$$K_1 = -10 \log_{10}\left(\frac{10^{-DI_{b1}/10}+10^{-OI_{f2}/10}+ \cdots\ 10^{DI_{f10}/10}}{10}\right) \quad (53)$$

where f_1 to f_{10} are the midfrequencies of the equal-loudness bands shown on Fig. 3. The loudness-directivity index may be obtained graphically by making use of curves such as those shown on Fig. 25 for the rigid disk in an infinite baffle. Thus, for various types of radiation the loudness-directivity index may be determined in terms of the dimensions of the radiating device. The loudness-directivity indexes for loudspeakers of the types listed in Table IV have been determined in this manner for a range of practical sizes, and they are shown graphically on Figs. 26 through 33.

Limiting Resolution in an Image-Orthicon-Type Pickup Tube*

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Summary—**An analysis is made of some of the factors which limit resolution in a television pickup tube of the image-orthicon type. Particular attention is given to effects of the glass-target characteristics and to departure from perfect focus in the image section, resulting from electron-emission velocities.**

It is shown that, as the width of lines in a resolution pattern is reduced, the signal generated tends to take the form of a small periodic modulation superimposed on a constant signal. This gives, in the reproduced image, a gradual reduction in contrast. A quantitative determination of limiting resolution, expressed as a number of lines, depends upon a physiological evaluation, not as yet available, of the minimum contrast in the reproduced image which will permit the pattern to be recognized as having a multiplicity of separate lines rather than a continuous blur.

The general effects of target thickness and of color of incident light discussed herein have been confirmed experimentally.

IN THE FUTURE development of television pickup tubes of the electron-image type, it is desirable to establish what limits, if any, can be set as to the possible improvement in tube performance, particularly as concerns the resolution capabilities of the tube. For example, it is desirable to know whether a tube having dimensions comparable to the present model (image orthicon[1] or image vericon[2]) can be designed and operated so as to give a picture having 1000-line resolution; and if so, what operating conditions must be met to achieve this result?

In an over-all television system using a pickup tube of this type, the resolution is limited by: (a) kinescope spot size; (b) amplifier pass band; (c) limitations in the scanning system of the pickup tube, including beam size and shape; (d) the characteristics of the pickup-tube target—resistivity and thickness; and (e) the resolution existing in the electron image formed on the target in the image section of the tube. This last item is externally affected by leakage into the image section of the magnetic and/or electric fields from the scanning-coil system, but it will be assumed that the effect of these stray fields can be completely eliminated by suitable shielding means, possibly combined with auxiliary field-bucking coils.

Scanning Section

If we disregard, for the moment, the image-resolution loss occurring in the image section, and that due to the glass target, the remainder of the system has the characteristics of a vericon or orthicon image-conversion system, with the same limitations. O. H. Schade[3] has demonstrated to the author the capabilities of such a system. He used an amplifier chain built to have a 20-Mc. pass band, and a kinescope specially selected for small spot size. With this system and a typical 2-inch orthicon (mosaic approximately $\frac{7}{8}\times 1\frac{1}{8}$ inches), he was able to show resolution in a wedge test pattern to about 900 lines. The limit here was set by the kinescope spot size, as he demonstrated by placing the signal on the tube during the retrace time. This extended the resolution to approximately 1400 lines. It would appear, then, that if the charge image on the scanned side of the target of an image tube has resolution of the order of 1000 lines, the scanning section should be able to transmit this information.

Target

The image-tube target acquires a positive charge distribution on the image side corresponding to the light distribution on the photocathode. During the time between scans, some of this charge diffuses through the target. The scanning beam deposits on each element enough negative charge to neutralize the positive charge which has diffused through the target, plus enough just

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[1] A. Rose, P. K. Weimer, and H. B. Law, "The image orthicon—a sensitive television pickup tube," Proc. I.R.E., vol. 34, pp. 424–432; July, 1946.

[2] A registered trade mark of Remington Rand, Inc.

[3] RCA Victor Division, Radio Corporation of America, Harrison, N. J.

to equal the remainder of the positive charge which has not yet reached the scanned side. The resistivity of the target must be sufficiently low to permit these free positive and negative charges to reach and neutralize each other during the time between successive scans. Higher resistance than this will cause the signal from a moving object to be greater than that for the same object stationary.

We can estimate the resistivity requirements quantitatively by considering an element of the target as equivalent to a plane-parallel capacitor shunted by a resistor. The time constant of this combination should be small compared with a frame time. If the target thickness is d and the area of an element is A, then the capacitance and resistance are given respectively by: $C_1 = 1.1KA/4\pi d \times 10^{-12}$ farad and $R_1 = \rho d/A$ ohms, dimensions being in centimeters. Hence, the time constant for leakage through the target material is

$$\tau_1 = R_1C_1 = \frac{1.1K\rho}{4\pi} \times 10^{-12} \text{ sec.}$$

To have charge neutralization within a frame time, 1/30 second, we must have $\tau_1 \ll 1/30$ sec. Solving for ρ, using $K=6$ for the dielectric constant of the glass, we have $\rho \ll 6\times10^{10}$ ohms cm. We may select as an upper acceptable value for the resistivity $\rho = 2\times10^{10}$ ohm cm. This is realized by Corning type-015 glass at temperature 46°C., or by Corning type-008 glass at temperature 72°C.

The charge deposited on an element of the target by the photoelectrons can diffuse laterally through the glass as well as directly to the opposite face. The result of this lateral diffusion is to cause gradual spreading of the images of small picture elements, with consequent loss of resolution. This loss of resolution due to charge spreading in the target is significant only at low light levels. At high-light-level operation, the spreading charge is continuously neutralized by collection of the secondary electrons released when the photoelectrons strike the target. The following discussion applies, therefore, only to low-light-level operation. The rate of spreading will decrease as the target thickness is reduced. We can set up the condition that the amount of charge which leaks in this manner to a distance as great as a picture element-width within one frame time should be small.

Quantitatively, consider a strip element of the target having width of one picture element σ and unit length, receiving a charge from the photoelectrons. The resistance between this strip and the uncharged region of the target at a distance σ from this strip is given by $R_2 = \frac{1}{2}(\rho\sigma/d)$ ohms. The capacitance of this strip to the target screen at a distance s from the target ($s \gg d$, and $\sigma > s$) is

$$C_2 = \frac{1.1\sigma}{4\pi s} \times 10^{-12} \text{ farad.}$$

Hence the time constant for charge leakage along the target is given by

$$\tau_2 = R_2C_2 = \frac{1.1\rho\sigma^2}{8\pi sd} \times 10^{-12} \text{ sec.,}$$

or, since

$$\rho = 2\times10^{10} \text{ ohm cm.,} \qquad \tau_2 = 8.8\frac{\sigma^2}{sd} \times 10^{-4} \text{ sec.}$$

In order to have small leakage to a distance equal to a picture-element width in a frame time, we must have $\tau_2 \gg 1/30$ sec. Solving for target thickness d, $d \ll 2.6\times10^{-2}\sigma^2/s$. For example, if the scanning pattern is 0.825 inch high and resolution of 500 lines is required, $\sigma = 1.65\times10^{-3}$ inch. The screen can not be mounted uniformly much closer than 10^{-3} inch from the target surface. Then, for these conditions, $d \ll 7.2\times10^{-5}$ inch. Since these figures involve considerable approximation, it may be concluded that target thickness should be of the order of 0.0001 inch.

For a better evaluation of the resolution capability of a target of given thickness, a more detailed analysis is indicated. We may consider the projected electron image as having the form of uniform strips separated by blank strips of the same width (alternate white and black bars) and examine the charge distribution on the target at the end of a frame time. Consider a section of the target taken across the resolution pattern, as represented in Fig. 1. The shaded sections represent regions

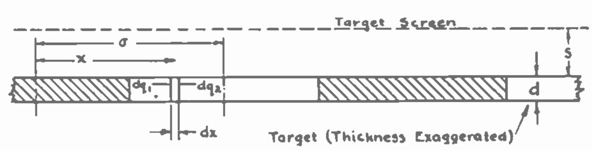

Fig. 1

on which charge is deposited. This charge flows into the clear regions, changing the potential distribution in both clear and shaded regions. Let us examine the charge accumulated in some element of width dx across the strips and of unit length along the strips. Charge will flow into the element, from the left, at a rate determined by the potential gradient at the left end of the element, and out of the element at a rate determined by the potential gradient at the right end of the element. The accumulation of charge in the element is at a rate given by

$$\frac{\partial q_x}{\partial t} = \frac{\partial q_1}{\partial t} - \frac{\partial q_2}{\partial t}.$$

But

$$\frac{\partial q_1}{\partial t} = \frac{-1}{R}\frac{\partial V_1}{\partial x}$$

and

$$\frac{\partial q_2}{\partial t} = \frac{-1}{R}\frac{\partial V_2}{\partial x}$$

where R is the resistance of the target material per unit area, $R=\rho/d$. Hence,

$$\frac{\partial q_x}{\partial t} = -\frac{1}{R}\left(\frac{\partial V_1}{\partial x} - \frac{\partial V_2}{\partial x}\right) = \frac{1}{R}\frac{\partial^2 V}{\partial x^2}dx.$$

There will also be a charge accumulation from electrons impinging on the target, at a rate

$$\left(\frac{\partial q_e}{\partial t}\right)_x dx$$

depending on the position of the element. This will have a constant value in the shaded regions and be zero in the clear regions. Hence,

$$\frac{\partial q_x}{\partial t} = \left[\frac{1}{R}\frac{\partial^2 V}{\partial x^2} + \left(\frac{\partial q_e}{\partial t}\right)_x\right]dx.$$

Now the potential of the element is determined by the charge q_x and the capacitance C_x of this element to the screen. $q_x = V_x C dx$, where C is the capacitance of unit area of the target, and hence $C_x = Cdx$.

$$C\frac{\partial V}{\partial t}dx = \left[\frac{1}{R}\frac{\partial^2 V}{\partial x^2} + \left(\frac{\partial q_e}{\partial t}\right)_x\right]dx,$$

or

$$\frac{\partial V}{\partial t} = \frac{1}{RC}\frac{\partial^2 V}{\partial x^2} + \frac{1}{c}i_{ex}.$$

This equation, integrated over a frame time, gives the charge distribution across the target at the time of scanning. For evident reasons of symmetry, it is sufficient to consider the distribution between the center of a bright strip and the center of a dark strip.

The equation for charge distribution is identical in form with that for heat flow along a unidimensional conductor having a number of heat sources along its length.

The solution of this equation presents very severe difficulties. An approximate solution, however, may be obtained as follows:

At time t, increments of charge dq are added to the images of the illuminated strip. This produces an increase in potential $dV = dq/C_x$ in each element of these regions. These charges may be regarded as spreading into the unilluminated regions, as though no other charges existed, for the interval from time t to time τ, the end of a frame time. After this interval, the potential distribution due to these elements of charge will be $(V_t)_x$. These may then be integrated with respect to t, summing up the contributions of all deposited charges to the final potential. First we investigate the distribution of potential after time t when the initial distribution is in the form of bars and blank spaces of equal width σ. The equation of flow

$$\frac{\partial V}{\partial t} = k\frac{\partial^2 V}{\partial x^2} \quad (t > 0), \qquad k \equiv \frac{1}{RC} \tag{1}$$

must be solved so as to satisfy the boundary conditions

$$V = f(x) \quad (t = 0, -\sigma \leqq x \leqq \sigma) \tag{2}$$

and

$$\left.\begin{aligned} &(V)_{x=\sigma} = (V)_{x=-\sigma} \\ &\left(\frac{\partial V}{\partial x}\right)_{x=\sigma} = \left(\frac{\partial V}{\partial x}\right)_{x=-\sigma} = 0 = \left(\frac{\partial V}{\partial x}\right)_{x=0} \end{aligned}\right\} \quad (t > 0). \tag{3}$$

The initial potential distribution (strips) may be represented by a Fourier expansion

$$(V)_{t=0} = f(x) = a_0 + \left(a_1 \cos\frac{\pi x}{\sigma} - b_1 \sin\frac{\pi x}{\sigma}\right) + \left(a_2 \cos\frac{2\pi x}{\sigma} + b_2 \cos\frac{2\pi x}{\sigma}\right) + \cdots.$$

In this expansion, the coefficients are given by

$$a_0 = \frac{1}{2\sigma}\int_{-\sigma}^{\sigma} f(x)dx; \qquad a_n = \frac{1}{\sigma}\int_{-\sigma}^{\sigma} f(x)\cos nx dx;$$

$$b_n = \frac{1}{\sigma}\int_{-\sigma}^{\sigma} f(x)\sin nx dx.$$

A solution of (1), satisfying the initial boundary conditions, is

$$V = \sum_{n=0}^{\infty}\left(a_n \cos\frac{n\pi x}{\sigma} + b_n \sin\frac{n\pi x}{\sigma}\right)e^{-(kn^2\pi^2/\sigma^2)t}.$$

Now it is apparent from the symmetry of the problem that $V_x = V_{-x}$, for $t \geqq 0$. This condition is satisfied if $b_n = 0$ for all values of n.

Thus

$$V = \sum_{n=0}^{\infty} a_n \cos\frac{n\pi x}{\sigma} e^{-(kn^2\pi^2/\sigma^2)t}$$

$$= \sum_{n=\sigma}^{\infty} a_n \cos\frac{n\pi x}{\sigma} e^{-Kn^2 t}$$

where

$$K \equiv \frac{k\pi^2}{\sigma^2} = \frac{\pi^2}{RC\sigma^2}.$$

For this problem, the initial function is

$$f(x) = V_0\left(-\frac{\sigma}{2} < x < \frac{\sigma}{2}\right)$$

and

$$f(x) = 0\left(-\sigma < x < -\frac{\sigma}{2}; \frac{\sigma}{2} < x < \sigma\right).$$

Evaluating the coefficients,

$$a_0 = \frac{1}{2\sigma}\int_{-\sigma/2}^{\sigma/2} V_0 dx = \frac{V_0}{2}$$

$$a_n = \frac{V_0}{\sigma}\int_{-\sigma/2}^{\sigma/2}\cos\frac{n\pi x}{\sigma}dx = \frac{2V_0}{n\pi}\sin\frac{n\pi}{2}$$

$$a_1 = \frac{2V_0}{\pi}, \; a_2 = 0, \; a_3 = -\frac{2V_0}{3\pi}, \; a_4 = 0, \; a_5 = \frac{2V_0}{5\pi}, \text{ etc.}$$

Then

$$V = \frac{V_0}{2} + \frac{2V_0}{\pi}\cos\frac{\pi x}{\sigma}e^{-Kt} - \frac{2V_0}{3\pi}\cos\frac{3\pi x}{\sigma}e^{-9Kt} + \frac{2V_0}{5\pi}\cos\frac{5\pi x}{\sigma}e^{-25Kt} - \cdots$$

$$= \frac{V_0}{2} + \frac{2V_0}{\pi}\sum_{n=0}^{\infty}(-)^n \frac{\cos\frac{(2n+1)\pi x}{\sigma}}{(2n+1)} e^{-(2n+1)^2Kt}.$$

This is now to be integrated with respect to t, to find the potential distribution due to charges arriving at various times during the cycle.

In the time interval between t and $t+dt$, an increment of charge dq_0 is deposited in the illuminated strip. This causes a potential rise dV_0 along this strip, given by $dV_0 = dq_0/C_1$ where C_1 is the capacitance between the illuminated strip and the target screen. It may be written

$$dV_0 = \frac{1}{C_1}\frac{dq_0}{dt}dt = \frac{I_0}{C_1}dt$$

with I_0 representing the current flowing to the illuminated strip.

The charge spreads along the target, in a direction normal to the strip, for the time interval from t to τ, the end of a cycle, at which time the entire charge is removed by the scanning beam. At the time τ, the contribution to the potential of an element at distance x from the center of the illuminated strip, made by this charge element dq_0, will be

$$(dV_x)_t = \frac{I_0}{C_1}\left[\frac{1}{2} + \frac{2}{\pi}\sum_{n=0}^{\infty}(-)^n \frac{\cos\frac{(2n+1)\pi x}{\sigma}}{(2n+1)} e^{-(2n+1)2K(\tau-t)}\right]dt.$$

Integrating on t, between limits o and τ, we get the total potential at the point x:

$$V_x = \frac{I_0}{C_1}\left[\frac{\tau}{2} + \frac{2}{K\pi}\sum_{n=0}^{\infty}(-)^n \frac{\cos\frac{(2n+1)\pi x}{\sigma}}{(2n+1)}\left(1 - e^{-(2n+1)2K\tau}\right)\right].$$

This gives (approximately) the potential distribution in the target at the time of scanning, resulting from uniform deposition of charge in the resolution pattern strips. The actual resolution obtained may be estimated by first expressing this potential distribution as a modulation

$$M = \frac{V_{x=0} - V_{x=\sigma}}{V_{x=0}}.$$

Then

$$M = \frac{2\left[(1 - e^{-K\tau}) - \frac{1}{27}(1 - e^{-9K\tau}) + \frac{1}{125}(1 - e^{-25K\tau}) - \cdots\right]}{\frac{\pi K\tau}{4} + \left[(1 - e^{-K\tau}) - \frac{1}{27}(1 - e^{-9K\tau}) + \frac{1}{125}(1 - e^{-25K\tau}) - \cdots\right]}.$$

The modulation, as a function of $K\tau$, is plotted in Fig. 2.

Estimate of thickness of target required is now dependent upon the assumed minimum useful modulation in the image. From the curve, a corresponding value of $K\tau$ may be determined. Then

$$K\tau = \frac{\pi^2\tau}{RC\sigma^2} = \frac{\pi^2 d \times 4\pi s\tau}{1.1\rho\sigma^2 \times 10^{-12}}.$$

Solving for d,

$$d = \frac{2.66\rho(K\tau)\sigma^2}{s} \times 10^{-13}.$$

If, for example, a modulation of 30 per cent gives sufficient contrast to be resolvable, then, for $M=0.3$, $(KT)=6.9$, and $d=1.00\times10^{-4}$ inch, or a glass target 0.10-mil thick would be sufficient to give 500-line resolution, s and σ having the same values as above.

It is convenient to reverse this formula and determine the number of lines resolution as a function of modulation, using a target region 0.825 inch high, for various target thicknesses.

Thus

$$\sigma = \sqrt{\frac{ds \times 10^{13}}{2.66\rho(K\tau)}}.$$

$$N = \frac{0.825}{\sigma} = 0.825\sqrt{\frac{2.66\rho(K\tau)}{ds \times 10^{13}}} \text{ (dimensions in inches).}$$

For

$$s = 10^{-3} \text{ inch}, \qquad \rho = 2 \times 10^{10} \text{ ohm cm.},$$

$$N = 1.90\sqrt{\frac{K\tau}{d}}.$$

For

$$d = 10^{-4} \text{ inch}, \qquad N = 190\sqrt{K\tau}.$$

Resolution, as a function of modulation, is plotted in Fig. 3 for representative values of ρ and d.

It may be noted that the target thickness required for a given resolution is proportional to the target resistivity. If, for example, the resistivity is lowered from 2×10^{10} ohm cm. to 1×10^{10} ohm cm., the target thickness required for a given resolution is halved.

The Image Section

In the image section of the tube, electrons emitted by the photocathode have a random distribution of initial velocities, both as to direction and speed. Because

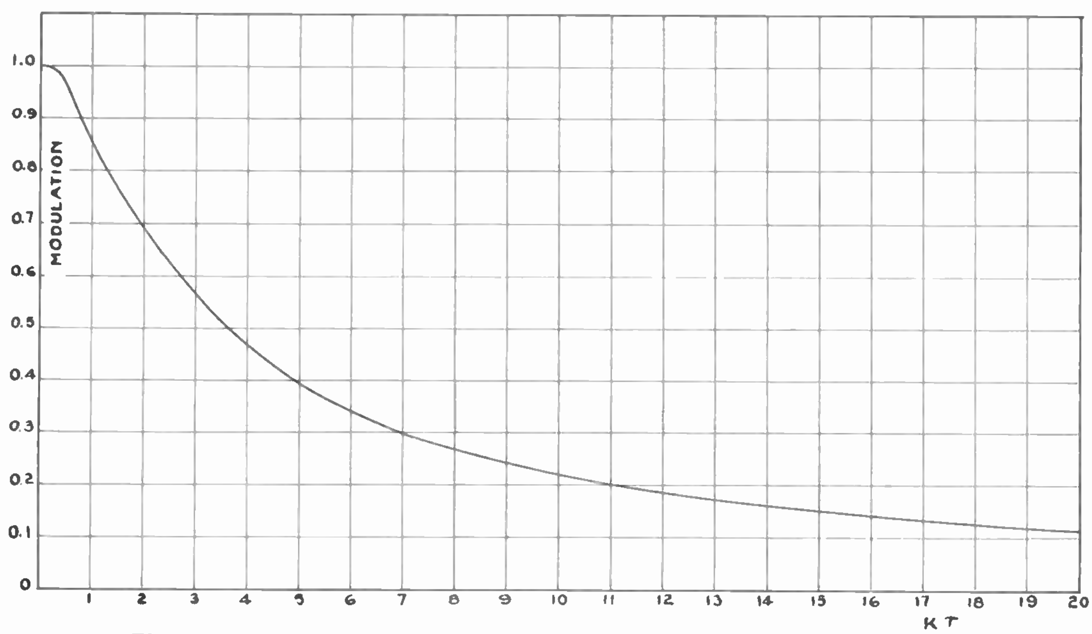

Fig. 2—Modulation in charge image on target caused by spreading through target.

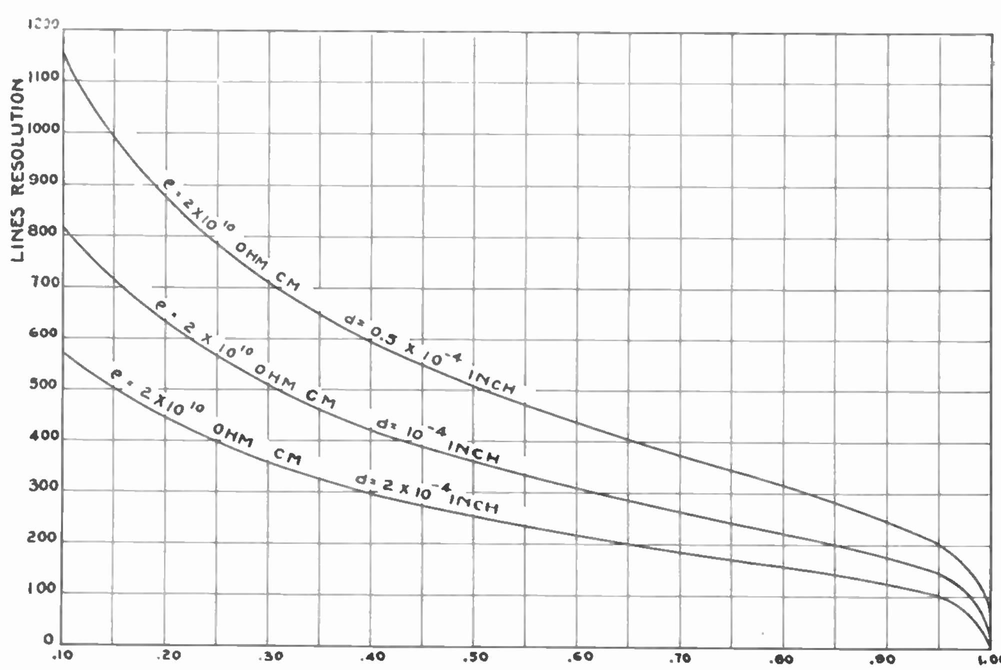

Fig. 3—Resolution limit determined by target leakage.

of this, the electrons from a point source can not be refocused to a mathematical point image at the target surface. The size of the image obtained will set a limit to the possible resolution obtainable. In operation of the tube under high-light-level conditions, the initial charge distribution obtained in the focused electron image will be modified by the redistribution of the returning secondary electrons. The following discussion is therefore limited to the case of operation under low light level, or "full-storage" conditions.

To simplify the problem as much as possible, assume that the accelerating field is plane parallel and that the magnetic focusing field is uniform as in Fig. 4. Both conditions will be met reasonably well over the greater

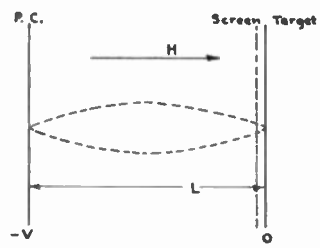

Fig. 4—Plane-parallel accelerating field and uniform magnetic field condition.

part of the image. The transit time for an electron moving from the photocathode to the target is given by

$$T = \frac{L}{v_{av}} = \frac{2L}{v_m + v_{il}}$$

where v_{av} = average velocity
m = maximum velocity
v_{ie} = initial velocity along tube axis.

From the well-known equations of motion for an electron in an electrostatic field, we have

$$v_m = \sqrt{\frac{2\,V_e}{m} + v_{ie}^2}.$$

If the electron is emitted at an angle θ to the tube axis, with an energy V_i electron volts, then the initial velocity is

$$v_i = \sqrt{\frac{2V_{ie}}{m}}$$

and the longitudinal and radial components are, respectively,

$$v_{ie} = \sqrt{\frac{2V_i e}{m}} \cos\theta \quad \text{and} \quad v_{ir} = \sqrt{\frac{2V_i e}{m}} \sin\theta.$$

Then

$$v_m = \sqrt{\frac{2Ve}{m} + \frac{2V_i e}{m}\cos^2\theta}$$

and

$$T = \frac{2L}{\sqrt{\frac{2Ve}{m} + \frac{2V_i e}{m}\cos^2\theta} + \sqrt{\frac{2V_i e}{m}\cos^2\theta}}.$$

For an electron with zero initial velocity, the transit time is

$$T_0 = \frac{2L}{\sqrt{\frac{2Ve}{m}}}.$$

Then the fractional reduction in transit time corresponding to a given initial velocity becomes

$$\frac{T_0 - T}{T_0} = 1 - \frac{1}{\sqrt{1 + z^2} + z}$$

where

$$z^2 = \frac{V_i}{V}\cos^2\theta.$$

Since $z \ll 1$, we may use the approximation

$$\frac{1}{\sqrt{1 + z^2} + z} \cong 1 - z.$$

Hence,

$$\frac{T_0 - T}{T_0} = z = \sqrt{\frac{V_i}{V}}\cos\theta.$$

It will be convenient, in the following discussion, to express all velocities in terms of equivalent electron volts.

The projected path, at the target, of an electron having an initial radial velocity component corresponding to V_{ir} volts, moving under the influence of magnetic field of H gauss in the image section of the tube, is a circle having a diameter

$$d = 6.7\,\frac{\sqrt{V_{ir}}}{H}\ \text{cm.},$$

traversed at uniform rate and described in time

$$T_1 = \frac{3.55 \times 10^{-7}}{H}\ \text{sec.}$$

The circle is tangent to a magnetic flux line passing through the point of origin of the electron at the photocathode.

Now,

$$T_0 = \frac{2L \times 10^{-7}}{5.93\sqrt{V}}\ \text{sec.}$$

With $L=4.5$ cm., and $V=300$ volts,

$$T_0 = 8.77 \times 10^{-9} \text{ sec.}$$

Since the time for one loop of focus is

$$T_1 = \frac{3.55 \times 10^{-7}}{H} \text{ sec.,}$$

there will be $2.47\times10^{-2}H$ loops of focus. For one loop of focus between photocathode and target, the magnetic field should be

$$H = \frac{10^2}{2.47} = 40.5 \text{ gauss.}$$

As the numerical values given are representative operating conditions for the tube, it may be concluded that there will be one loop of focus in the image section of the tube, under usual operating conditions.

If the transit time T differs from T_0, the electron will describe less than a full circle (projected path) and will arrive at the target at a distance r from the true focus, as shown in Fig. 5.

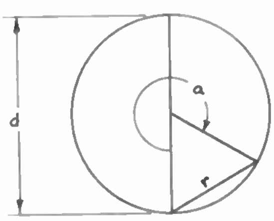

Fig. 5—Electron path as a consequence of transit time.

The angle α is given by

$$\alpha = 2\pi\left(1 - \frac{T}{T_1}\right) = 2\pi\left(\frac{T_1 - T}{T_1}\right)$$

and

$$r = 2\left(\frac{d}{2}\right) \sin \frac{\alpha}{2} = d \sin \pi\left(\frac{T_1 - T}{T_1}\right).$$

Since it has been assumed that conditions have been established for focusing the zero initial-axial-velocity electrons, $T_1 = T_0$
and

$$r = d \sin \pi\left(\frac{T_0 - T}{T_0}\right) = d \sin\left(\pi\sqrt{\frac{V_i}{V}} \cos\theta\right).$$

Now,

$$V_r = \frac{mv_r^2}{2e},$$

so that

$$d = \frac{6.7\sqrt{\dfrac{mv_r^2}{2e}}}{H} = 6.7\,\frac{\sqrt{V_i}}{H} \sin\theta.$$

Hence,

$$r = 6.7\,\frac{\sqrt{V_i}}{H} \sin\theta \sin\left(\pi\sqrt{\frac{V_i}{V}}\cos\theta\right).$$

Actual operating focusing conditions will differ slightly from those appropriate to the zero initial-axial-velocity electrons, in the sense that the charge distribution derived here will be slightly more diffuse than that actually obtained. However, this departure should be small enough not to affect the results seriously.

At this point it is appropriate to make a numerical estimate of V_i.

The general equation of photoemission, $V_m e = h\nu - W = h(\nu - \nu_0)$, gives the maximum emission velocity of photoelectrons in terms of the incident light frequency ν and the work function W, or equivalent threshold frequency ν_0, characteristic of the photosurface. For a silver-oxygen-cesium photosurface, the threshold frequency varies considerably, but, a reasonable value may be taken as corresponding to a wavelength of 10,000 Angstrom units. On the other hand, the image tube is generally employed under conditions utilizing considerably shorter wavelengths in the visible portion of the spectrum. An incident light frequency corresponding to a wavelength of 5500 Angstrom units, the peak sensitivity of the eye, may be taken as a useful region for discussion. Then the maximum emission velocity is derived from

$$V_m e = hc\left(\frac{1}{5.5 \times 10^{-5}} - \frac{1}{10^{-4}}\right) = 1.60 \times 10^{-12} \text{ erg.}$$

$$V_m = 1.00 \text{ volt.}$$

As this value is small compared with the 300-volt accelerating potential, we may, in the above expression for r, replace

$$\sin\left(\pi\sqrt{\frac{V_i}{V}}\cos\theta\right)$$

by the argument, giving

$$r = 6.7\,\frac{\sqrt{V_i}}{H} \sin\theta \cdot \pi\sqrt{\frac{V_i}{V}}\cos\theta.$$

$$= RV_i \sin 2\theta,$$

where

$$R = \frac{10.5}{H\sqrt{V}} = \frac{L}{V}.$$

Let us now follow an argument substantially identical with that first used by Fry and Ives.[4]

[4] A. L. Hughes and L. A. DuBridge, "Photoelectric Phenomena," McGraw-Hill Book Co., New York, N. Y., 1933, pp. 129–131.

From an element dA of the photocathode, there will be emitted $I_0 dA$ electrons. Of those emitted at angle θ to the normal, those with initial velocities between V_i and V_i+dV_i will strike the target at radii, from perfect focus, between r and $r+dr$. Thus, the permitted range in emission velocities for this region is $(dV_i/dr)dr$. Of the electrons emitted, a fraction $p(\theta)d\theta$ will start within the angular range between θ and $\theta+d\theta$, and of these a fraction $p(V_i)dV_i$ will have velocities to permit falling within the target range under consideration. Hence, the total number of electrons falling between r and $r+dr$ from the focus will be

$$I_0 dA\, p(\theta)d\theta p(V_i)\frac{dV_i}{dr}dr.$$

These electrons fall upon a target region having the area $2\pi r dr$. Hence, the current density at radius r from the focus will be

$$i_r=\frac{I_0 dA}{2\pi r}\int_{\theta_{\min}}^{\theta_{\max}} p(\theta)d\theta p(V_i)\frac{dV_i}{dr}\cdot$$

The integration limits are the extreme emission angles within which some electrons can land within the target region. The lower limit $\theta_{\min}$ is determined as the smallest angle for which electrons with the largest initial velocity V_{im} can strike the target at the given value of r from focus:

$$\sin 2\theta_{\min}=\frac{r}{RV_{im}}\cdot$$

The upper limit $\theta_{\max}$ is given by

$$\theta_{\max}=\frac{\pi}{2}-\theta_{\min},$$

as is apparent from the expression

$$\frac{r}{RV_i}=2\sin\theta\cos\theta.$$

Now it has been found experimentally that $p(\theta)$ and $p(V_i)$ can be represented by

$$p(\theta)=2\sin\theta\cos\theta=\sin 2\theta \qquad \text{(Lambert distribution)}$$

$$p(V_i)=\frac{6}{V_{im}}\left[\frac{V_i}{V_{im}}-\left(\frac{V_i}{V_{im}}\right)^2\right] \quad \text{(empirical form).}$$

Since

$$V_i=\frac{r\csc 2\theta}{R}, \qquad \frac{dV_i}{dr}=\frac{\csc 2\theta}{R}\cdot$$

Hence,

$$i_r=\frac{I_0 dA}{2\pi r}\int_{\theta_{\min}}^{\theta_{\max}}\sin 2\theta\cdot\frac{6}{V_{im}}\left[\frac{V_i}{V_{im}}-\left(\frac{V_i}{V_{im}}\right)^2\right]d\theta\cdot\frac{\csc 2\theta}{R}$$

$$=\frac{3I_0 dA}{\pi R^2 V_{im}^2}\int_{\theta_{\min}}^{\theta_{\max}}\csc 2\theta\left(1-\frac{r\csc 2\theta}{RV_{im}}\right)d\theta.$$

Let $RV_{im}\equiv a$, which is the maximum distance from true focus at which an electron can land.
Then

$$i_r=\frac{3I_0 dA}{\pi a^2}\int_{\theta_{\min}}^{\theta_{\max}}\csc 2\theta\left(1-\frac{r\csc 2\theta}{a}\right)d\theta$$

$$=\frac{3I_0 dA}{\pi a^2}\left[\log\tan\theta+\frac{r}{a}\cot 2\theta\right]_{\theta_{\min}}^{\theta_{\max}}.$$

Substituting for the limits $\theta_{\max}$ and $\theta_{\min}$, as given above,

$$i_r=\frac{3I_0 dA}{\pi a^2}\left[\log\left(\frac{a+\sqrt{a^2-r^2}}{r}\right)-\frac{\sqrt{a^2-r^2}}{a}\right]$$

$$=\frac{3I_0 dA}{\pi a^2}\left[\operatorname{sech}^{-1}\left(\frac{r}{a}\right)-\frac{\sqrt{a^2-r^2}}{a}\right].$$

This equation gives the charge-density distribution at the target corresponding to electrons emitted with a charge density I_0 from an element dA of the photocathode.

Consider now the charge-density distribution in the image of an illuminated strip of width W on the photocathode. In computing the current-density distribution in the image, we can distinguish three cases: $W\geqq 2a$, $2a\geqq W\geqq a$, and $W\leqq a$.

Case I. $W\geqq 2a$, Outside Boundary of Strip

The current density at the point P is found by summing the contributions from all elements of the photocathode in the illuminated strip, such that the projected distance from the point P on the target is less than or equal to a, since none of the emitted electrons can strike the target beyond $r=a$. (See Fig. 6.) The

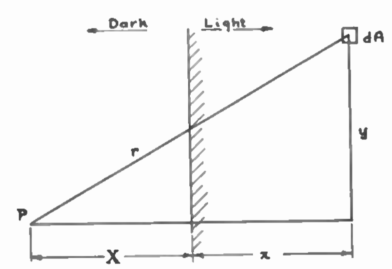

Fig. 6—Charge density outside the boundary of a wide illuminated strip.

contribution at the point P of charge emitted by the element dA is

$$di_P=\frac{3I_0}{\pi a^2}\left[\operatorname{sech}^{-1}\left(\frac{r}{a}\right)-\frac{\sqrt{a^2-r^2}}{a}\right]dxdy, \quad dA=dxdy$$

$$=\frac{3I_0}{\pi a^2}\left[\operatorname{sech}^{-1}\left(\frac{\sqrt{(X+x)^2+y^2}}{a}\right)\right.$$

$$\left.-\frac{\sqrt{a^2-(X+x)^2-y^2}}{a}\right]dxdy.$$

Since integration is over that part of the illuminated strip for which $r \leqq a$, the integration limits are $y = \pm\sqrt{a^2-(X+x)^2}$ and $x=0$ to $x=a-X$. For reasons of symmetry, the y integral can be represented as twice the integral from $y=0$ to $y=\sqrt{a^2-(X+x)^2}$. Then the current density at the point P is

$$I_P = \int_x\int_y di_P = \frac{6I_0}{\pi a^2}\int_x\int_y \left[\operatorname{sech}^{-1}\left(\frac{\sqrt{(X+x)^2+y^2}}{a}\right) - \frac{\sqrt{a^2-(X+x)^2-y^2}}{a}\right]dx\,dy.$$

By making the substitutions $y/a = v$ and $x/a = w$, this equation may be integrated, with the result

$$I_P = \frac{I_0}{2}\left(1-\frac{X}{a}\right)^3.$$

Simplifying the notation,

$$\frac{I_P}{I_0} = P\left(\frac{X}{a}\right)$$

where

$$P\left(\frac{X}{a}\right) \equiv \frac{1}{2}\left(1-\frac{X}{a}\right)^3.$$

This gives the charge density outside the boundary of the image of a wide illuminated strip. We have next to consider the distribution inside the strip boundary, and also that for a strip which is not wide.

Case II. Inside Strip Boundary

$$\frac{I_P}{I_0} = \frac{6}{\pi a^2}\int_0^X\int_0^{\sqrt{a^2-(X-x)^2}}\left[\operatorname{sech}^{-1}\frac{\sqrt{(X-x)^2+y^2}}{a} - \frac{\sqrt{a^2-(X-x^2)-y^2}}{a}\right]dxdy$$
$$+\frac{6}{\pi a^2}\int_X^{X+a}\int_0^{\sqrt{a^2-(x-X)^2}}\left[\operatorname{sech}^{-1}\frac{\sqrt{(x-X)^2+y^2}}{a} - \frac{\sqrt{a^2-(x-X)^2-y^2}}{a}\right]dydx.$$

which may be integrated, with the result

$$\frac{I_P}{I_0} = 1 - P\left(\frac{X}{a}\right).$$

If, now, the strip width W is less than $2a$, the distribution inside the image boundaries is modified, while if $W<a$ the distribution outside the boundary is also modified. (See Fig. 7.)

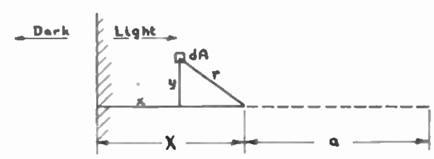

Fig. 7—Charge distribution inside the strip boundary.

Case III. Outside Boundary of Narrow Strip.

For values of X such that $(X+W)/a > 1$, we have the same value of charge as for Case I,

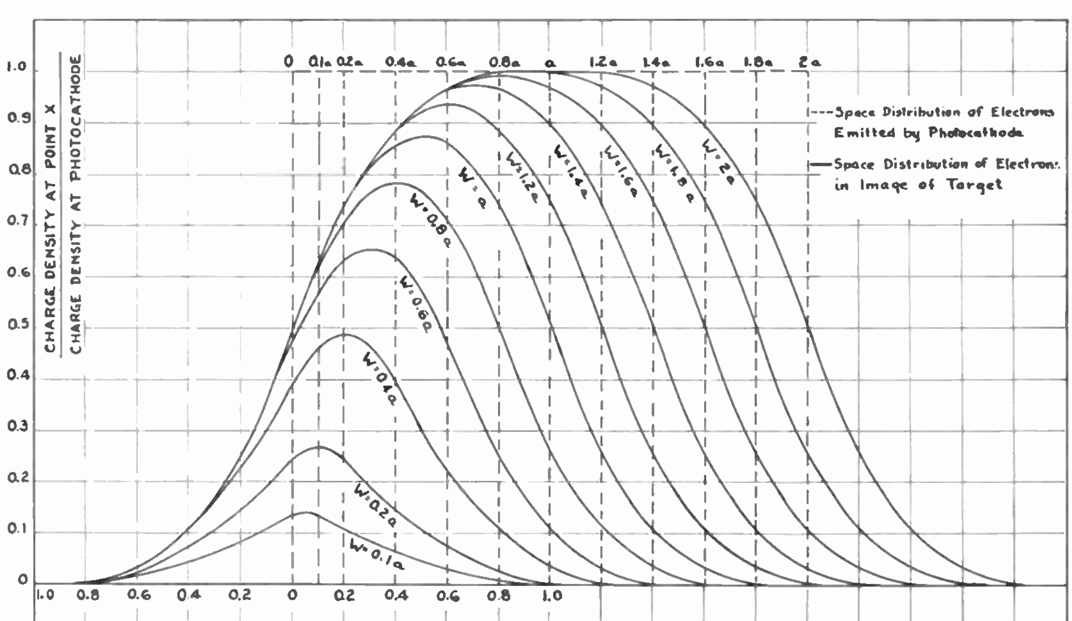

Fig. 8—Curves of charge density at the target of the image vericon with slit of light of width W on photocathode.

$$\frac{I_P}{I_0} = P\left(\frac{X}{a}\right).$$

When $X+W<a$, the integration limits on x are changed from

$$\begin{cases} x = a - X \\ x = 0 \end{cases} \quad \text{to} \quad \begin{cases} x = W \\ x = 0 \end{cases}.$$

With these limits, the integration gives

$$\frac{I_P}{I_0} = P\left(\frac{X}{a}\right) - P\left(\frac{W+x}{a}\right), \quad (0 \leqq X \leqq a - W).$$

Case IV. Inside Boundary of Narrow Strip.

Here we have the same two integrals as in Case II, with the difference that the limits on the second are changed from

$$\begin{cases} x = a + X \\ x = X \end{cases} \quad \text{to} \quad \begin{cases} x = W \\ x = X \end{cases}.$$

Carrying out the integration, we obtain

$$\frac{I_P}{I_0} = 1 - P\left(\frac{X}{a}\right) - P\left(\frac{W-X}{a}\right), \qquad (W \leqq 2a).$$

Summarizing the expressions for charge distributions in the image of an illuminated strip of width W:

For

$W \geqq 2a$,

$$\frac{I_P}{I_0} = P\left(\frac{X}{a}\right) \quad \text{outside boundary} \quad (0 \leqq X \leqq a)$$

$$\frac{I_P}{I_0} = 1 - P\left(\frac{X}{a}\right) \quad \text{inside boundary} \quad (0 \leqq X \leqq a).$$

For

$W < 2a$,

$$\left.\begin{aligned} \frac{I_P}{I_0} &= P\left(\frac{X}{a}\right) - P\left(\frac{W+X}{a}\right), && (0 \leqq X \leqq a - W) \\ &= P\left(\frac{X}{a}\right), && (a - W \leqq X \leqq a) \end{aligned}\right\}$$

outside boundary

$$\frac{I_P}{I_0} = 1 - P\left(\frac{X}{a}\right) - P\left(\frac{W-X}{a}\right), \quad (0 \leqq X \leqq W)$$

inside boundary.

In the above expressions, $a = RV_{im}$ cm. where V_{im} is the maximum initial velocity expressed in volts, and $R = L/V$.

For an image orthicon operating at $L = 4.5$ cm. and $V = 300$ volts, $R = 1.50 \times 10^{-2}$. Hence, for $V_{im} = 1.0$ volt, $a = 1.50 \times 10^{-2}$ cm. $= 0.15$ mm.

Since the charge distribution in the image is the same for any strip width for which $W \geqq 2a$, it is only necessary to extend calculations to this width. The curves in Fig. 8 give results for charge distributions, calculated from the formulas above.

In order to resolve two strips of a given width, their separation must be great enough that the charge distribution formed by their overlapping image patterns gives a double-humped, rather than a single-humped, total distribution. For the separation of two parallel bright strips of a given width, this means that, in the charge distribution for each, the peak charge plus the charge at a distance S from the center must be greater than twice the charge at a distance $S/2$.

The common resolution test for pickup tubes uses a wedge of black and white strips of equal width. The center-to-center distance S is twice the width W of each illuminated strip. Furthermore, since there is a multiplicity of strips, the image charge patterns may overlap to a considerable extent, as shown in Fig. 9.

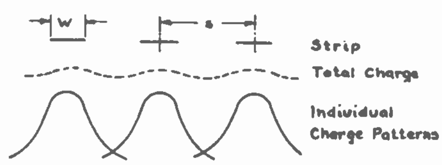

Fig. 9—Image charge pattern overlap.

It can be seen that the peak charge is given by the peak charge for one strip plus twice the charges at distances from the strip boundary given by

$$\frac{X}{a} = \frac{2nW}{a} - \frac{W}{2a} = \frac{4n-1}{2}\,\frac{W}{a}.$$

This must include contributions from all regions such that $(X/a) < 1$, $n = 1, 2, 3, \cdots$. Similarly the minimum, or valley, charge is given by twice the charges at distances from the strip boundary for which

$$\frac{X}{a} = \frac{2nW}{a} + \frac{W}{2a} = \frac{4n+1}{2}\,\frac{W}{a}, \qquad n = 0, 1, 2, \cdots.$$

For the resulting total charge distribution, we shall define the modulation M by

$$M = \frac{\text{charge at peak} - \text{charge at valley}}{\text{charge at peak}}.$$

The curve of Fig. 10 shows modulation as a function of strip width.

The problem of the limiting resolution obtainable in the image section now resolves itself into the question of the minimum observable modulation in the reproduced picture. No data are available to answer this question.

Furthermore, there is a loss in contrast in the reproduced picture caused by noise, by the finite bandwidth of the transmission system, and by contrast loss in the kinescope. This last loss is due partially to finite scanning-spot size and partially to light scattering. Estimates of the minimum modulation that would enable one to recognize the separate lines of the resolution pattern vary; but perhaps 20 per cent ± 10 per cent may be taken as a reasonable assumption.

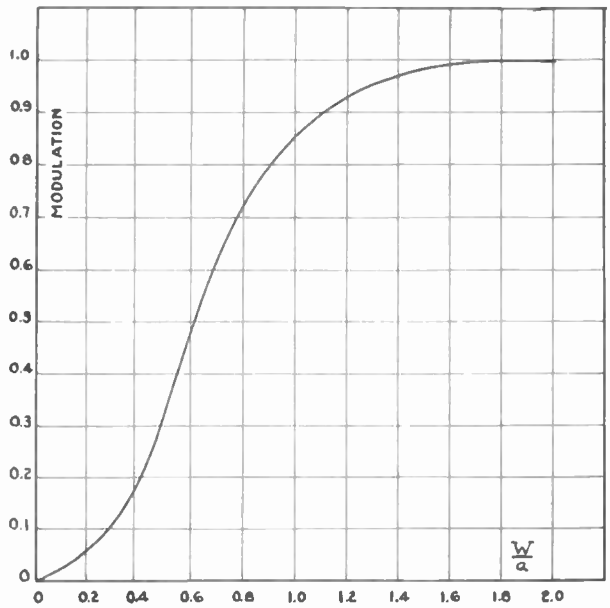

Fig. 10—Modulation in electron image of resolution pattern.

We have, for minimum line width as given by the modulation curve,

Modulation 10 per cent—$W_L = 0.28$
Modulation 20 per cent—$W_L = 0.41$
Modulation 30 per cent—$W_L = 0.48$.

We have already seen that for an image vericon or image orthicon in normal operation, with −300 volts on the photocathode, $a = 1.50 \times 10^{-2}\, V_{im}$ cm. where V_{im} is the maximum initial velocity. The scanning raster in these tubes is approximately 21×28 mm. The limiting resolution is given by Res. $= 2.1/W_L$ lines. (See Table I).

TABLE I

Minimum Modulation (per cent)	W_L Centimeters	Res. Lines	Res. at 5500 Angstrom Units ($V_{im} = 1.00$ volt)
10	$0.42 \times 10^{-2}\, V_{im}$	$\frac{500}{V_{im}}$	500 lines
20	$0.62 \times 10^{-2}\, V_{im}$	$\frac{340}{V_{im}}$	340 lines
30	$0.72 \times 10^{-2}\, V_{im}$	$\frac{290}{V_{im}}$	290 lines

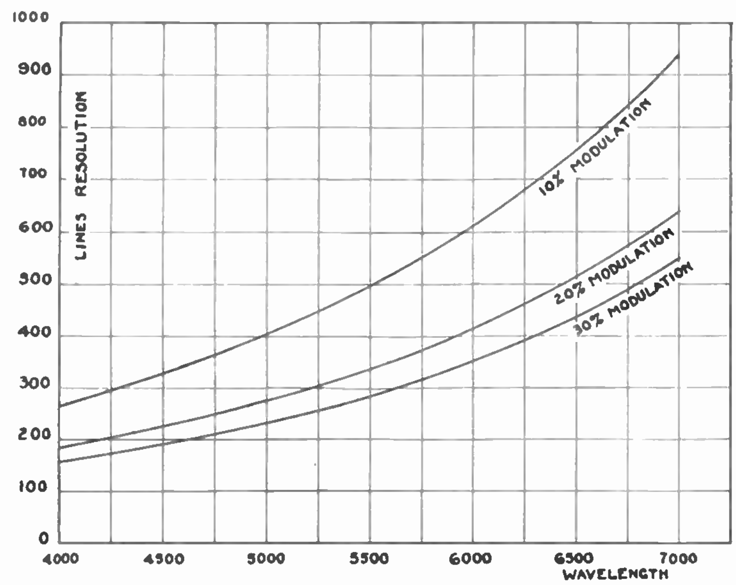

Fig. 11—Modulation in image on target of image vericon.

The effect on limiting resolution of the wavelength of the incident light may be seen in the curves of Fig. 11, based upon $\lambda_0 = 10{,}000$ Angstrom units.

Conclusion

The analysis herein indicates the manner in which the resolution capability of an electron-image type of television pickup tube is affected by target material and thickness and by the color of light employed, taken in conjunction with the cutoff characteristic of the photosensitive surface. The results have been confirmed in a qualitative fashion by tests with tubes having different target characteristics and by tests using different color filters in the optical path.

In order to predict quantitatively the resolution to be expected with any given tube design, it would be necessary to know the modulation in a picture which is required in order that a fine-line pattern will just be discernible as discrete lines. In the absence of such information, the analyses given in this paper serve to indicate means whereby resolution can be improved and provide an estimate of the improvement to be expected from any given change.

Acknowledgment

The author wishes to express appreciation for a number of valuable suggestions made by Albert Rose of RCA Laboratories Division, Radio Corporation of America.

Reflections of Very-High-Frequency Radio Waves from Meteoric Ionization*

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Summary—**The reflection of v.h.f. radio waves by ionization, produced by the passage of meteors through the upper atmosphere, results in echoes of short duration. For frequencies between 40 and 50 Mc. the duration is usually not more than a few tenths of a second, although infrequently one of several seconds duration is observed. The characteristics of the echoes as determined by measurements are discussed, and correlations shown between the distributions of occurrence and the theoretical and observed distributions of meteors. The observations indicate that the reflection cannot be treated as specular, but that the reflected waves form a lobe in the direction of the reflected ray which would exist if the reflection were specular.**

THE DATA from which this paper was prepared were obtained under a project for the continuous recording of selected f.m. and television stations in the frequency range from 42 to 84 Mc. After some preliminary tests at a temporary site in Washington, D. C. in 1942, four continuous chart recorders, operating from a single half-wave antenna, were set up at the Federal Communications Commission's Monitoring Station at Laurel, Md., and recordings were begun on February 1, 1943. During the summer and fall of 1943 additional equipment became available, and recorders were installed at the monitoring stations at Allegan, Mich., Grand Island, Neb., Atlanta, Ga., and Portland, Ore. The stations were chosen so as to provide data on the magnitude of the effects of the lower atmosphere at various distances and times, which would furnish a measure of the reliability of the Commission's theoretical signal-range charts, and to obtain recordings of signals reflected from sporadic patches of abnormally high ionization in the *E* layer, which were known to occur on frequencies up to about 60 Mc.

In addition to the signals which travel via the lower atmosphere, commonly called tropospheric signals, and the sporadic-*E* signals, an unexpected signal of different type was observed, which was termed a "burst" because of its characteristic rapid rise and short duration. The bursts varied in intensity from bare audibility above receiver noise, which would produce no motion of the recording pen, to indicated values of about 70 microvolts per meter. The duration was usually not more than a few tenths of a second, although infrequently one of several seconds duration was observed. In the great majority of the bursts the signal was apparently undistorted, and short passages of music or speech would be heard with great clarity. Occasionally, however, noticeable distortion occurred and in the bursts of longer duration a flutter would sometimes be present, indicating the arrival of waves by two or more paths. At times when bursts were received simultaneously with the tropospheric wave from relatively near stations, Doppler whistles of extremely short duration, and usually of descending pitch, were heard. There were also rumbles and other multipath distortion at such times.

When the bursts were first identified as being something other than peaks of rapidly fading tropospheric waves, several possible causes were advanced by those who had observed them, such as reflection from aircraft, from undulating tropospheric discontinuities, or from patches of ionization. A further extension of the latter theory was to the effect that the patches of ionization were caused by the passage of meteorites through the upper atmosphere. The ionization effects of meteors in connection with radio propagation at lower frequencies had been investigated by A. M. Skellett,[1] J. A. Pierce,[2] and others, and the occurrence at 10 Mc. of bursts of somewhat longer duration than those at 40 to 50 Mc. had been attributed to meteoric ionization by Pierce. Short-distance scatter effects at 10 Mc., which consisted of bursts of from one-half to one second duration, had also been investigated by T. T. Eckersley[3] and attributed to patchy *E*-layer ionization, for which meteors had been advanced as a probable cause. Upon the installation of recorders at Allegan, Atlanta, and Grand Island, signal bursts from frequency-modulation station WGTR at Paxton, Mass., were found to be received at all points. Since the distance from WGTR to Grand Island, 1370 miles, is about the limit of distance for single reflections from the *E* layer, theories of comparatively low-level reflections, such as from aircraft and the troposphere, were abandoned in favor of ionospheric reflection.

In order to determine the propagation path lengths of the burst pulses, a series of pulse tests was made at Laurel in conjunction with station W2XMN at Alpine, N. J., 197 miles away, in October 1943. This station was selected because a relatively steady signal was needed for reference pulses, between which the burst pulses would appear, if there was any difference in path length. A method of pulsing was used which will be explained in connection with Fig. 1. It consisted in frequency-

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† Federal Communications Commission, Washington 25, D. C.

[1] A. M. Skellett, "The ionizing effect of meteors in relation to radio propagation," PROC. I.R.E., vol. 20, pp. 1933–1941; December, 1932.

[2] J. A. Pierce, "Abnormal ionization in the *E* region of the ionosphere," PROC. I.R.E., vol. 26, pp. 892–909; July, 1938.

[3] T. T. Eckersley, "Analysis of the effect of scattering in radio transmission," *Jour. I.E.E.*, vol. 86, pp. 548–567: June, 1940.

modulating the transmitter ±75 kc. by a continuous tone of 170 c.p.s. The f.m. signal was received on a f.m.

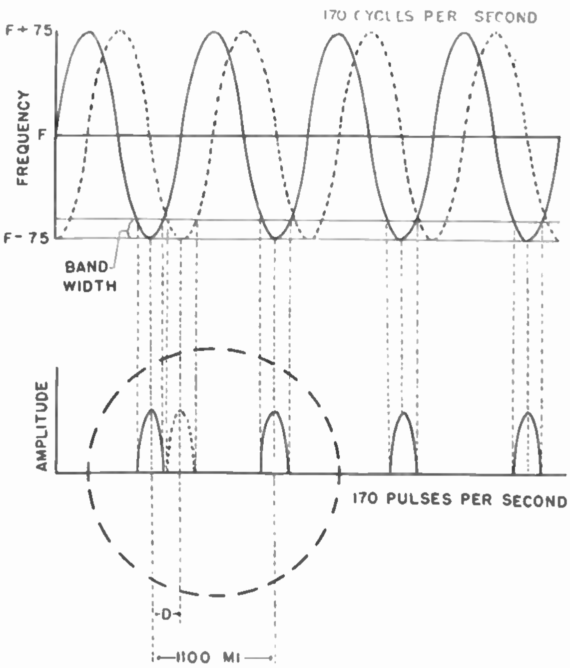


Fig. 1—Method of measuring path lengths of v.h.f. bursts.

receiver, the i.f. of which was passed to an a.m. receiver tuned to a narrow pass band at the lower end of the swing. This produced narrow pulses of tone frequency in the i.f. output of the a.m. receiver which were placed on the vertical plates of an oscilloscope, indicated diagrammatically by the dashed circle. The horizontal sweep was set at one-half tone frequency, so that two reference pulses appeared simultaneously on the screen. Any difference in path length D caused the burst signal to be delayed by an interval D/c, where c is the velocity of propagation, so that the frequency deviations of the delayed burst occurred between those of the ground-wave signal. In Fig. 1, the deviations of the burst signal are shown as a dashed curve lying between the cycles of the solid wave of the ground-wave signal, the burst pulse on the oscilloscope screen appearing at a distance D to the right of the zero reference pulse.

During the tests the estimated path differences ranged from about 150 to 900 miles, corresponding to total path lengths of from 350 to 1100 miles. The estimation of the shorter distances was made somewhat difficult by the pulse width, which tended to merge the delayed pulse with the reference pulse, but the values obtained indicate that the medium responsible for the shorter paths was separated from the sea-level great-circle path by a

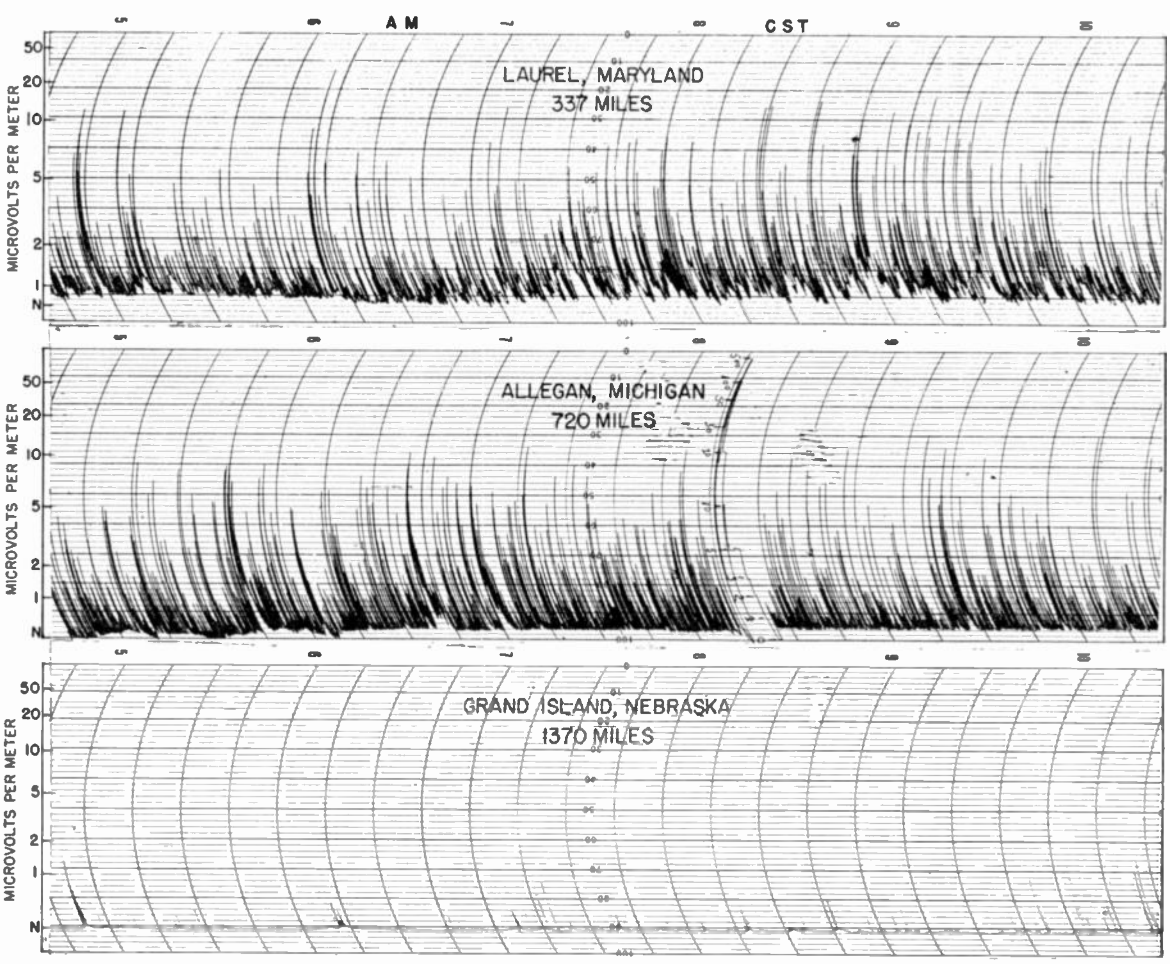


Fig. 2—Bursts from station WGTR, Paxton, Mass. (44.3 Mc.; 340 kw.) recorded simultaneously at three locations (September 17, 1943).

distance comparable to the height of the E layer. The greater distances can be interpreted as reflections from higher media, or from media of height comparable to the E layer, but lying to each side of the great-circle plane. In the light of subsequent information, the latter interpretation is felt to be correct.

Simultaneous recordings of bursts on 44.3 Mc. received at Laurel, Allegan, and Grand Island from f.m. station WGTR are shown in Fig. 2. The effective radiated power of the station is 340 kw. in the equatorial plane of its 10-bay turnstile antenna. The maximum burst intensities for this period were 30 μv/m. at Laurel, 12 at Allegan, and 1.4 at Grand Island, revealing a decreasing intensity with increasing distance. However, the record for Allegan shows a greater number of bursts at medium levels and indicates that the distribution of intensities versus numbers was not the same for Laurel and Allegan.

The results of an analysis of six morning hours of data recorded simultaneously at Laurel and Allegan to determine the intensity distributions are shown in Fig. 3. The numbers indicated along the abscissa are the cumulative numbers for all six hours which had intensities equal to or exceeding the intensities indicated by the ordinates. As in the preceding sample, bursts of greater intensity were recorded at Laurel, but at somewhat below the greatest intensities more bursts were

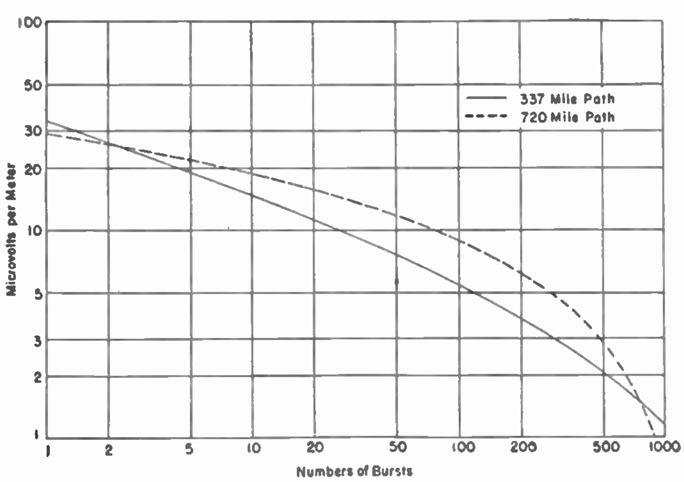


Fig. 3—Distributions of intensities versus numbers of bursts.

recorded at Allegan. While the lobe structures of the transmitting and receiving antennas over existing terrain, and some other factors, are not known with sufficient detail to obtain a mathematical verification of the distributions, they are in qualitative agreement with what would be expected for reflecting media principally of E-layer height and random distribution. Assuming that the principal lobe of the transmitting antenna encounters the E layer at distances between 500 and 700 miles, it will cover more area which is common to a receiving antenna 700 miles distant than one 300 miles distant, if the principal lobes of the receiving antenna are effective over a radius of 300 miles at E-layer height. This will account for the greater numbers of bursts of medium intensity. Infrequently a meteor will penetrate below E-layer level and encounter a region common to the two antenna patterns, which will produce a shorter path length and a higher burst intensity for the 300-mile spacing.

The numbers of bursts per hour exceeding 5-μv. recorder input (3.3 μv./m.) received at Laurel from station WGTR have been determined for the period from February, 1943, through May, 1944, for comparison with observed and theoretical variations in the occurrence of meteors. In Fig. 4, curve (a) shows the distribution with time of day of the average of 12 monthly

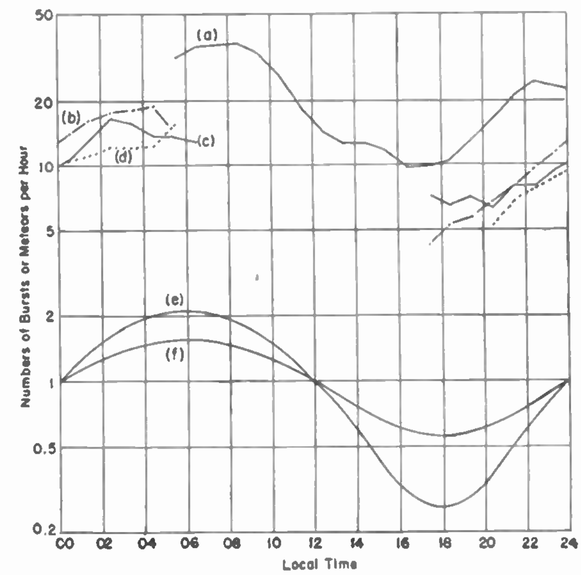


Fig. 4—Diurnal distributions of the hourly numbers of bursts and meteors.

median values of the numbers of bursts per hour for the period February, 1943, through January, 1944. Curves (b), (c), and (d) are the average observed numbers of meteors per hour during night hours as reported by three workers in the field, Schmidt, Coulvier-Gravier, and Hoffmeister.[4] The two smooth curves (e) and (f) are theoretical distribution curves for 40° North Latitude computed for meteors with parabolic orbits (e) and with hyperbolic orbits (f). The agreement in the range of variation is very good, the range in the numbers of the bursts (3.7) being between the hyperbolic distribution (2.8) and the parabolic (8.0). The time of the observed maximum and minimum do not coincide exactly with the theoretical, but the shift of the minimum to an earlier hour is consistent with the better radio propagation conditions which are found to prevail in the late afternoon and early evening hours, which will

[4] C. P. Olivier, "Meteors," Williams and Wilkins, Baltimore, Md., 1925; p. 182.

tend to increase the numbers of bursts exceeding a fixed reference level. This may also affect the range of variation somewhat, but good propagation conditions usually exist at the time of maximum in the morning. A similar

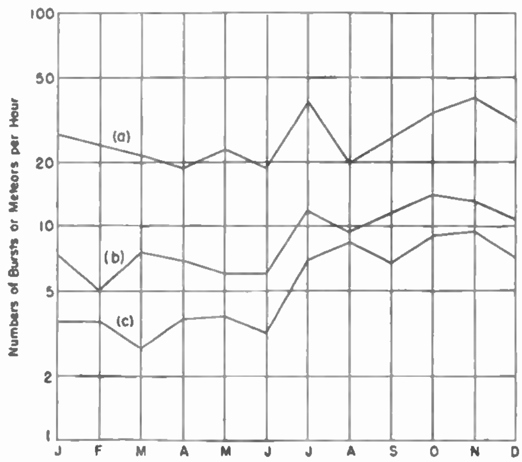


Fig. 5—Annual distributions of the monthly numbers of bursts and meteors.

The annual distribution of the monthly median of the numbers of bursts occurring during the hour 11–12 P.M. is shown by curve (a) of Fig. 5. The period covered is from June, 1943, through May, 1944, over which time station WGTR was operating with substantially the same output power and antenna pattern. Curves (b) and (c) are observed monthly averages according to Hoffmeister and Coulvier-Gravier.[4] While there is not consistent agreement in the relative monthly levels, a general agreement in the trends is noted. Differences between observations and between observed numbers and bursts can conceivably be due in part to variations from year to year. Note the July maximum in the numbers of bursts which agrees with the observations of Hoffmeister, while Coulvier-Gravier shows the maximum in August. A similar August maximum was reported by four other observers.[4]

In Fig. 6 is shown the annual distribution of the numbers of bursts occurring during the hour 12–1 A.M. in comparison with the principal meteor showers. The average duration and date of maximum of each shower

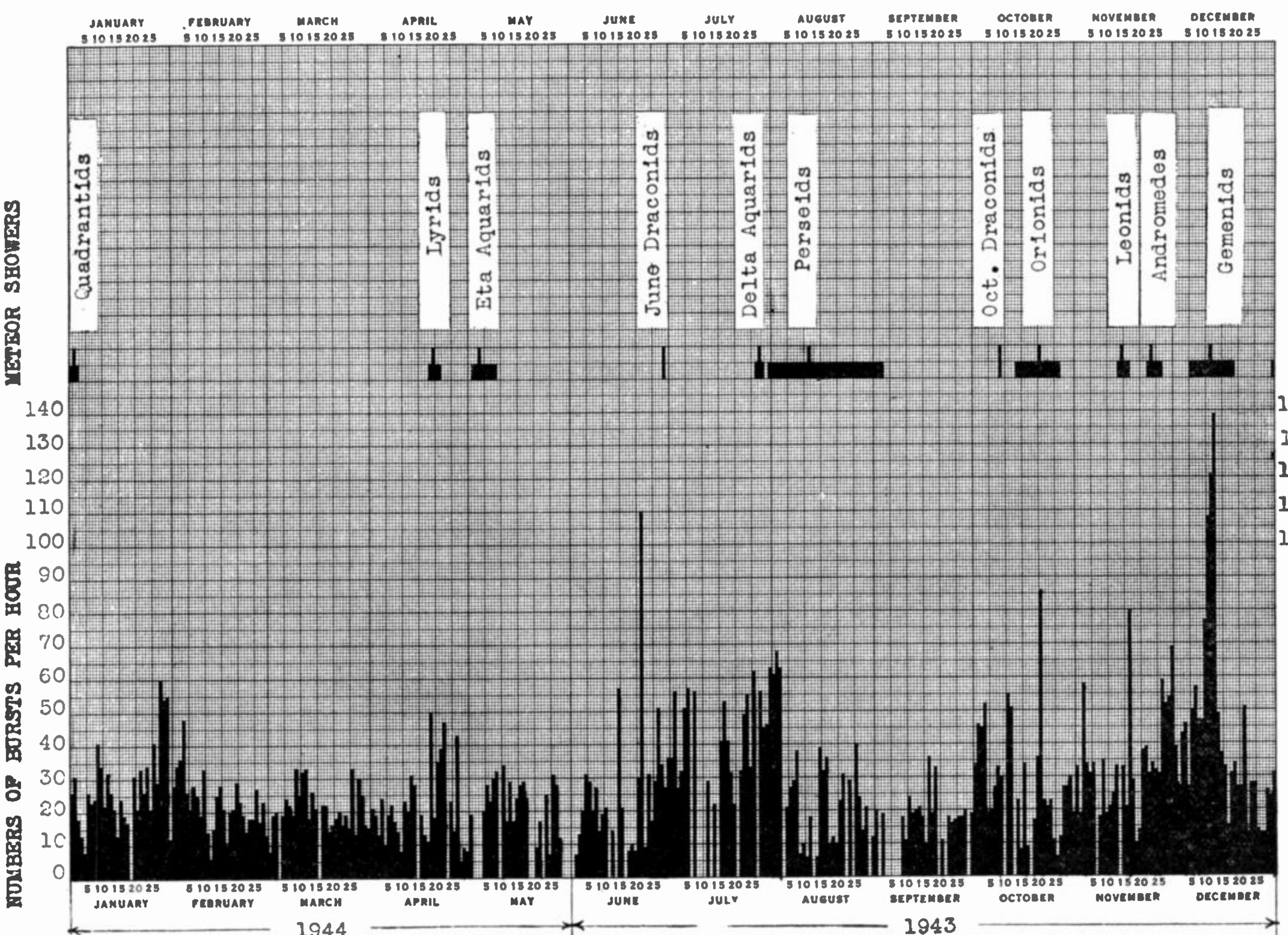


Fig. 6—Annual distributions of the hourly numbers of bursts and of meteor showers.

diurnal distribution for short-distance scatter effects was obtained by Eckersley.[3]

has been indicated by the blocked-in portions of the chart, but there is considerable variation from year to year for most showers, and some showers do not appear every year, so that exact coincidence with the maxima of recorded bursts is not to be expected. Other factors which can be expected to affect the correlation are the time of day during which the shower occurs at the path under consideration and the radiant or direction of arrival of the meteors. Nevertheless, except for the period of high bursts in the latter part of January for which no meteor shower is recognized, and the unusually low numbers during the Perseid shower in August, the times of greatest activity are during or reasonably near to the times of meteor showers. The hours during which no bursts are shown are hours for which no data are available because the transmitter or receiver was not in operation. No zero counts were obtained.

During June, 1944, after the agreement between the occurrences of meteors and bursts had been established, a few observations were made at Laurel in an attempt to obtain coincidences between bursts and visible meteors. Several meteors were seen, but none having a proper direction of flight, so that the ionized track was capable of reflecting the signal to the receiving point. Beginning on August 1, 1944, a continuous watch was kept between 9 and 10:15 P.M., E.S.T. on favorable nights by two or more observers at the author's home. Two coincidences were observed on August 6, in which the meteor track was approximately transverse to the signal path. On August 8 and 11, two coincidences were observed each night in which the meteor track was along the plane of the signal path. The last meteor was of particular brilliance with a persistent visible train, and the signal was sustained for about ten seconds. Observations were continued throughout the early fall and were repeated the follower summer. A total of thirteen coincidences was obtained. In making these observations it became evident that reflection was taking place from the side of the ionized track of the meteor, rather than from the meteor or the head of the track.[5] All of the meteors which produced bursts were apparently traveling in such a direction as to provide this condition. Several bright meteors whose line of fall, if continued, would have terminated between the transmitter and receiver, produced no bursts. The existence of this condition is further borne out by observations on meteoric ionization by radar. The radar pips do not show large changes in range with time, as would be the case if the pulses were being reflected from the region near the head of a meteor having a component of motion relative to the radar set. Some slight shifts in indicated range have been noted, but these would be expected by reason of the shift of the point of reflection, owing to changes in shape of the ion contours.[6]

If the radio waves are returned by reflection from the side of the meteor track, the geometry of reflection in a simple case may be as shown in Fig. 7. Assume a meteor track MM^1 lying in the plane of the transmitter T and the receiver R, having a surrounding cloud of ionized air in which contours of equal ion density will have substantially the form shown. The contours will increase rapidly to maximum radius by reason of diffusion, and beyond this point decrease slowly, owing to further diffusion and recombination.[7] If the contour shown is of proper density it will reflect radio waves from the transmitter T to the receiver R, the incident and reflected waves subtending equal angles ϕ with the normal to the contour.

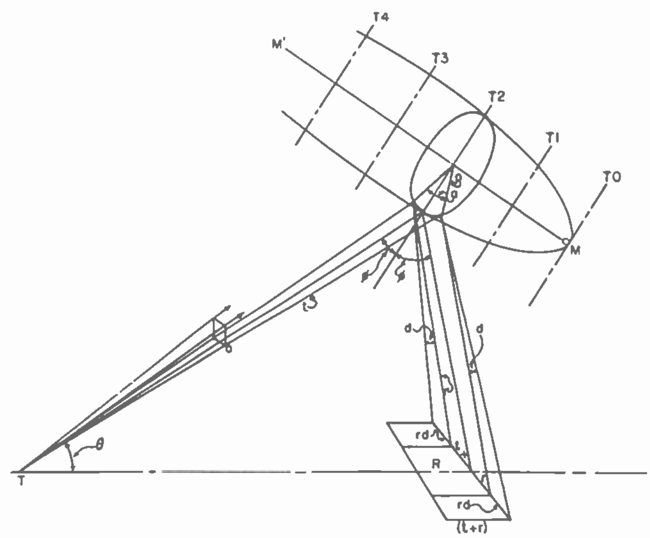


Fig. 7—Geometry of reflection of radio waves from ionized meteor tracks.

If reflection occurred from a plane surface, the area (square meters) at R covered by the energy passing through unit area O (one square meter) at unit distance (one mile) from T would be

$$(t+r)^2$$

(t and r expressed in miles), and the ratio of field intensities of the incident ray would be

$$\frac{E_0}{E_R} = (t+r).$$

However, owing to the finite radius p (meters) of the cloud cross section, the energy will suffer divergence through an angle d and will be spread over an area

$$(t+r+2rd)(t+r).$$

The ratio of field intensities will be

$$\frac{E_0}{E_R} = \sqrt{(t+r+2rd)(t+r)}.$$

[5] O. G. Villard, Jr. "Meteor detection by amateur radio," *QST*, vol. 31, pp. 13–18; July, 1947.

[6] O. P. Ferrell, "Meteoric impact ionization observed on radar oscilloscopes," *Phys. Rev.*, vol. 69, pp. 32–33; January 1 and 15, 1946.

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

Now

$$d = a \cos \phi$$
$$\approx \frac{t}{p} \cos \phi$$

and

$$\frac{E_0}{E_R} \approx \sqrt{\left(t + r + \frac{2tr \cos \phi}{p}\right)(t + r)}.$$

If the cloud is not cylindrical at the point of incidence, but has a curvature of radius q along its length, additional divergence will result and

$$\frac{E_0}{E_R} = \sqrt{\left(t+r+\frac{2tr \cos \phi}{p}\right)\left(t+r+\frac{2tr \sec \phi}{q}\right)}.$$

Although the above expressions are developed for the simple case in which the meteor track lies in the plane of propagation, similar but somewhat more complex expressions can be developed for cases in which the track crosses the plane. In each case, however, if the contours of equal ionization are uniform along the track, the geometry will be such that, at the point of reflection at a given instant, the contours will be normal to a line bisecting the angle between the paths of the incident and reflected waves. This will mean, in general, that the track of the meteor will also be substantially normal to the bisector.

Over the period from February, 1943, through May, 1944, the highest burst received at Laurel from station WGTR was 70 μv./m. Assuming that the highest bursts were recorded under conditions that were optimum; that is, that the reflecting point lay near the great-circle path between the transmitter and receiver, application of the above formula yields the following radii for limiting cases. For reflection at the midpoint of the path: $\phi = 67°$, $\theta = 21°$, $E_0 = 350$ millivolts per meter (calculated at 21° above horizontal for a 10-bay turnstile having a field of 2540 millivolts per meter at a mile in the horizontal plane), $p = 0.25$ meters for a cylindrical track, $N = 3.7 \times 10^6$ ions per cubic cm. For reflection at a point over the receiver: $\phi = 38°$, $\theta = 9°$, $E_0 = 640$ millivolts per meter, $p = 0.17$ meters for a cylindrical track, $N = 1.5 \times 10^7$ ions per cubic cm. Assuming (a) that the indicated radii are median for contours of all ion densities and that equal numbers of ions exist inside and outside of the contour of indicated radius, (b) that the initial meteor velocity is 40 kilometers per second, (c) that all of the kinetic energy of the meteor is converted into ionization, and (d) that the length of the meteor track is 100 kilometers, the above results indicate that the sizes of the meteoric particles are 2.9×10^{-7} and 9.8×10^{-7} grams, respectively. These particle sizes are much smaller than have hitherto been assumed as responsible for the larger meteors, and no reasonable assumptions as to the distribution of the ion densities within the cloud will yield values in line with the more reasonable estimates of 0.25 gram made by Pierce.[2]

It appears from the foregoing that the return should not be treated as a lossless specular reflection. At a frequency of 44 Mc. the losses due to absorption should be negligible. Specular reflection should not obtain unless the dimensions of the reflecting surface are very large compared with the wavelength, and while the above radii calculated for specular reflection are too small, the applicable radii are probably of the same order of magnitude as the wavelength. Considerable scattering of the energy must occur, with a possible field pattern consisting of a primary lobe of radiation centered on the line corresponding to the reflected ray. In this regard it should be possible to treat the ionized track as a traveling-wave antenna, with its primary lobe tilted forward in the direction of wave travel. This lobe would be very sharp for long meteor tracks, and it would not be possible to distinguish between the geometry of reception from such a lobe and the geometry of Fig. 7 by any observations which have been made to date. The field intensities produced by this lobe, while less than for specular reflection, would be dependent upon the same geometric characteristics of the meteor track; that is, the angle of departure for the radiation will be equal to the angle of the incident radiation, the sharpness of the lobe will depend upon the length of the track, the energy in the lobe will depend upon the energy intercepted by the track, and hence upon its radius and length.

Except for absolute intensity, Fig. 7 may be useful in arriving at certain characteristics of the bursts; for example, the variation of the intensity of the burst with time. Consider the successive cross sections of a meteor track at a high altitude, which are responsible for the intensity of the bursts at times T_0, T_1, T_2, T_3, T_4, etc., indicated in Fig. 7. The distributions of the ion densities at these times are plotted in a qualitative manner in Fig. 8(a), with ion density N as ordinate and radius p as

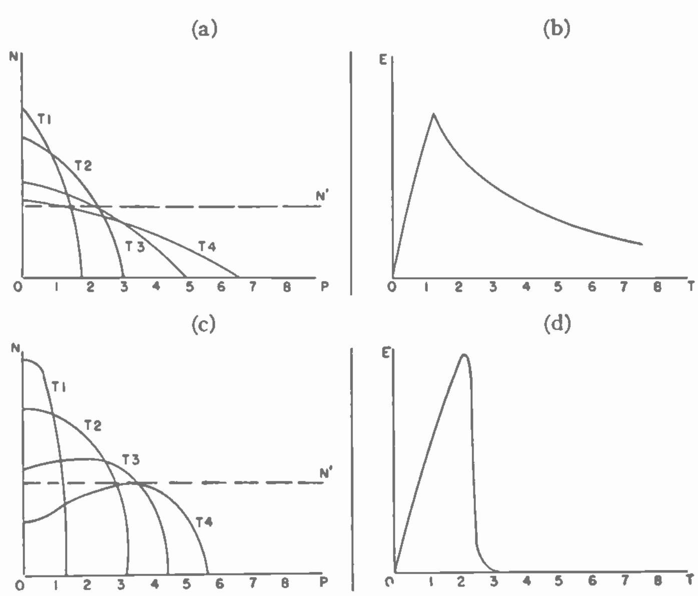


Fig. 8—Effect of height of meteor track on characteristics of bursts.
(a) and (b) Meteoric ionization at high altitudes.
(c) and (d) Meteoric ionization at low altitudes.

abscissa. Initially the central ionization is very intense and the gradient steep, but diffusion rapidly decreases both the maximum density and the gradient. The dashed line N^1 represents the density required to reflect the frequency on which the burst is received. It may be seen that the radius p rapidly increases to a maximum and then decreases, resulting in the envelope shown in Fig. 8(b). The maximum intensity of the burst, its time of occurrence, and the shape of the rise and decay slopes are dependent upon the level N^1 and hence upon the frequency of the reflected wave. Fig. 8(c) shows the effects of ionization at low altitudes. Owing to the greater density, diffusion is less rapid and recombination sets in more rapidly. Because of the lesser mobility of the positive ions, the recombination should take place initially at short radii, and may result in less ionization at the center of the cloud than near its outer boundary. This would result in the production of a burst having the shape of Fig. 8(d), with somewhat slower rise and an instantaneous decay of its trailing edge.

In preparation for the meteor shower of October 9, 1946, associated with the appearance of the Giacobini-Zinner comet, a continuous chart recorder was installed by Federal Communications Commission engineers at the Sterling (Va.) Laboratory of the National Bureau of Standards. Laboratory personnel had installed an SCR-270 radar for observation, and it was hoped that coincidences between bursts, radar indications, and visual observations could be obtained. No visual observations were possible because of rainy weather. A few coincidences between indications on the radar and the recorder were obtained on the night of October 9, when the recorder was tuned to the f.m. station at Paxton, Mass., and the radar was directed to the radiant of the shower at its predicted maximum, 320° true azimuth and 40° altitude. On October 10, the radar was oriented toward Paxton so that the propagation paths were more comparable, and although the number of bursts and radar returns were much less because of the decreased activity of the shower, a larger percentage of coincidences was observed.

Fig. 9 is a record of five bursts on 44.3 Mc. at a chart speed of three inches per minute, taken at Sterling, between 7:59 and 8:16 P.M., E.S.T., on the night of October 9, 1946. The first shows a comparatively slow rise and a slow decay time. The rise of the remaining four bursts is substantially instantaneous, two having rapid and two having slow decay times. From the foregoing analysis, bursts 1, 2, and 4 probably were produced by meteors at high level, and bursts 3 and 5 by meteors at a lower level. The bursts have serrated envelopes, which are probably due to interference between waves reflected from a plurality of points on an irregular ion cloud, or from a plurality of clouds caused by the simultaneous fall of a group of associated meteors.

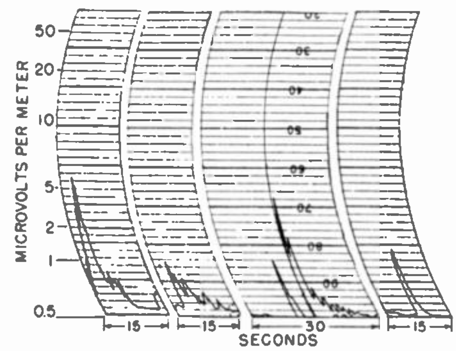


Fig. 9—Bursts on 44.3 Mc. recorded at fast chart speed.

Conclusion

The production of signal bursts in the frequency band near 40 Mc. by the reflection of radio waves from meteoric ionization has been established by observed coincidences between such bursts and visible meteor trains. A good correlation has been found between the diurnal distributions of the numbers of bursts, the observed numbers of meteors, and the theoretical distributions for randomly disposed meteors. A reasonably good correlation has also been found between the annual distributions of bursts and observed meteor numbers. The intensities of the bursts are lower than would be expected for specular reflection, and it is suggested by reason of the relatively small cross sections of the meteor tracks that a ray theory is not applicable, and that the reflection for simple cases comprises a primary lobe centered on the line of an assumed reflected ray.

Acknowledgment

The collection of the radio data on which this report is based was made possible by the co-operative efforts of my colleagues within the Commission, many of whom also made helpful suggestions as to the methods of analysis and the form of presentation of the data. Particular acknowledgment is made to E. W. Chapin, who suggested the pulse method and participated in the experiments, to W. K. Roberts, R. J. Renton, and G. L. Jensen, who also participated, and to G. L. Gadea, who performed much of the work of analysis. Acknowledgment is also made to N. Smith of the National Bureau of Standards, to C. P. Olivier of the Flower Observatory, and to O. P. Ferrell, for helpful discussion and reference to previous publications on meteoric ionization, and to E. H. Armstrong, in co-operation with whom the pulse measurements were made.

Rainfall Intensities and Attenuation of Centimeter Electromagnetic Waves*

RAYMOND WEXLER†, AND JOSEPH WEINSTEIN†

Summary—**The frequency distribution of rain intensities at four selected United States stations are analyzed with the view of determining the extent of rain attenuation on "X"-band radar. An areal study of the attenuation produced by heavy rain is also made.**

Purpose

THE MAXIMUM RANGE of radar in storm detection depends upon the radar parameters, the path attenuation, and the intensity of rain detected. For "X"-band radar, the atmospheric attenuation is approximately 0.01 db/km. More serious attenuation is caused by intervening rain to such an extent that heavy rain may occasionally block the radar from detecting storms beyond 25 miles. It is the purpose of this paper to determine the extent of rain attenuation, for "X"-band radar, to be expected at various localities in the United States.

Theory

On the assumption of complete interception of the radar beam by the rain storm, the maximum range R of a 3.2-centimeter radar set is given by[1]

$$R^2 = 1.34 \frac{AP_t d}{P_{r_{\min}}} Na^6 \, 10^{-0.2KR} \tag{1}$$

where A is the area of the antenna, P_t the peak power, $P_{r_{\min}}$ the minimum detectable signal, d the pulse length, Na^6 is the summation of the sixth power of the radii of the droplets per unit volume of rain cloud, and K is the path attenuation in db per unit distance. The term Na^6 in 10^{-10} centimeter^3 is related to the rain intensity I in mm. per hour by the regression equation

$$\log Na^6 = 1.441 \log I - 1.302 \tag{2}$$

as derived from the drop-size data of Laws and Parsons.[2,3]

Robertson and King[4] determined, experimentally, the attenuation of 3.2-centimeter electromagnetic waves to be approximately 0.03 db/km. per mm. per hour rainfall. There was considerable variation about this value. According to Ryde,[5] the theoretical attenuation for 3.2-centimeter waves, based on the mean drop-size data of Laws and Parsons, is about 0.02 db/km. per mm. per hour rainfall.

With the aid of (1) and (2), Fig. 1 is drawn indicating the theoretical maximum range of "X"-band (3.2-centimeter) radar as a function of the radar parameters, target intensity, and the path attenuation. A value of $AP_t d/P_{r_{\min}} = 10^{25}$ centimeters^3 corresponds approximately to that of the radar set AN/APQ-13, an "X"-band set widely used for storm detection. A value of

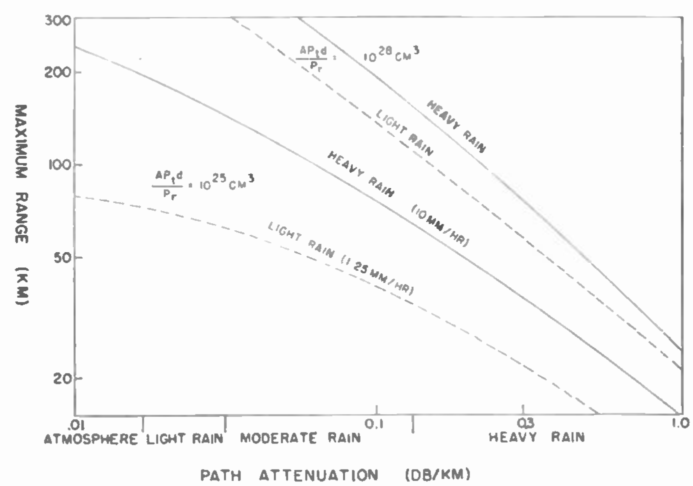


Fig. 1—Maximum range of X-band radar as a function of radar parameters, target intensity and path attenuation.

$AP_t d/P_{r_{\min}} = 10^{28}$ centimeters^3 is considered as a possible high-power "X"-band radar with large antenna area, long pulse length, and a more sensitive receiver. The abscissa in Fig. 1 indicates the regions of atmospheric attenuation due to oxygen and water vapor, attenuation due to light rain (<2.5 mm. per hour), moderate rain (2.5 to 7.5 mm. per hour), and heavy rain (>7.5 mm. per hour), based on Ryde's theoretical values for rainfall attenuation.

At a path attenuation of 0.3 db/km. in Fig. 1, the maximum range of the AN/APQ-13 is 35 km. for a target of heavy rain (10 mm. per hour) and 22 km. for a target of light rain (1.25 mm. per hour). In comparison, the high-power "X"-band radar should be able to detect heavy rain to a distance of 75 km. and light rain to a distance of 56 km. It is interesting to note that an increase in the radar parameters $AP_t d/P_{r_{\min}}$ of 1000-fold

* Decimal classification: R113.501. Original manuscript received by the Institute, June 30, 1947. Presented, American Meteorological Society, May, 1947.

† Signal Corps Engineering Laboratories, Belmar, N. J.

[1] R. Wexler and D. C. Swingle, "Radar storm detection," *Bull. Amer. Met. Soc.*, vol. 28, no. 1, pp. 159–167; April 1947.

[2] Dr. J. S. Marshall, Canadian Army Operational Research, derived, from his experimental data, an equation equivalent to

$$\log Na^6 \; 1.667 \log I - 1.348.$$

[3] O. J. Laws and D. A. Parsons, "The relation of raindrop-size to intensity," *Trans. Amer. Geophys. Union*, vol. 24, pp. 452–459; 1943.

[4] S. D. Robertson and A. P. King, "The effects of rain upon the propagation of waves in the 1- and 3-centimeter regions," Proc. I.R.E., vol. 34, pp. 178–180; April, 1946.

[5] Unpublished manuscript.

little more than doubles the maximum range of "X"-band radar through heavy rain.

If no intervening rain occurs, the path attenuation due to oxygen and water vapor is only about 0.01 db/km. The maximum range of the AN/APQ-13 would then be 240 km. for the detection of heavy rain and 80 km. for light rain. Theoretically, assuming complete interception, the high-power radar would be able to detect light rain out to 600 km., but earth shadow, and the fact that rain clouds usually do not extend to heights greater than 10 km., would effectively limit the range in most cases to below 300 km.

Fig. 1 is drawn on the assumption of complete interception of the radar beam by the rain cloud. Assuming that the lowest portion of the beam left the antenna at an angle of 0° with the earth's surface, the height above the earth at a distance of 240 km. would be about 3 km. If the height of the rain cloud at that range were only 9 km., then, only the lower half of the 3° beam width of the AN/APQ-13 would be intercepted by the cloud. The range of detection of heavy rain by the AN/APQ-13 would then be about 200 km., instead of the 240 km. indicated in Fig. 1. Considering the fact that the upper portion of a heavy rain cloud may not have the reflection characteristics of heavy rain that reaches the ground, the actual maximum range probably is considerably less.

Rainfall Intensities

In Fig. 2, the frequency distribution of rainfall intensities during the summer is given for Boston, Columbus, New Orleans, and Oklahoma City. The data for New Orleans has been obtained from a study by McDonald[6] based on 30 years of records, and the Oklahoma City data from Alexander[7] based on 25 years of records. For Boston and Columbus, the data has been compiled from the Monthly Meteorological Summaries for the years 1935 to 1946. Fig. 2 represents the frequency distribution of the amount of rain falling within a prescribed hourly period whether or not the rain was continuous during the period. It is seen that 60 to 80 per cent of all measurable summer rains are in the light-rain category (below 0.10 inch per hour). During the winter, 80 to 90 per cent of all rains are in this category.

It is seen from Fig. 2 that an intensity of 0.40 inch per hour or greater occurs during the summer at Boston, 2.4 per cent of the measurable rain hours (or about 3 hours per summer); at Columbus, 5.3 per cent (5 hours per summer); at Oklahoma City, 7.1 per cent (5 hours per summer); and at New Orleans, 11.2 per cent (13 hours per summer). If a rain of 0.40 inch (10 mm.) per hour occurred over an extended area, then, it may be seen from Fig. 1 that, at a value of 0.2 db/km., the range of the AN/APQ-13 would be reduced to 45 km. and that of the high-power "X"-band radar to 105 km. Using the experimental value of 0.3 db/km. for a 10 mm. per hour rainfall, the respective ranges would be 35 km. and 75

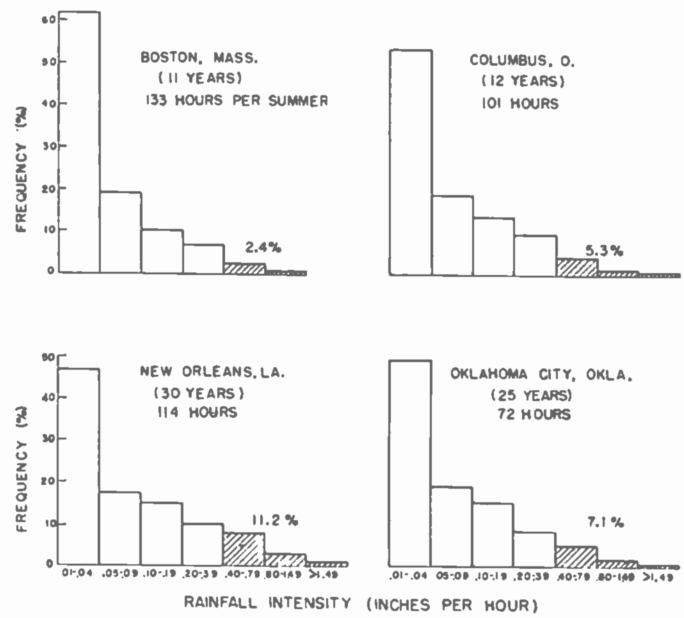


Fig. 2—Frequency distribution of rainfall intensities at four selected stations.

km. On the assumption that the heavy rains extend over long distances, the shaded portion in Fig. 2 would represent the percentage of measurable rain hours that the range of the specified "X"-band radars would be reduced below these distances.

Since most rains of heavy intensity are of short duration, it is probable that such storms are not widespread. At Boston in a 10-year period of summer rains, it was found that only thirty-five storms occurred which had at least one mean hourly intensity which could be classed as heavy (in this case, 0.31 inch per hour or greater). Four storms showed two consecutive hours of heavy rain, and one showed three consecutive hours of heavy rain. The remaining thirty storms had heavy rain within one hour only. This summary leads to the belief that the occurrence of heavy rains is itself infrequent and that heavy rains of a widespread character are rare.

In an effort to determine the extent of heavy rainfall, the rainfall data of four stations in an area around Houston, Tex., were studied. The four stations, Houston, Satsuma, Katy, and Alief, are all 13 to 15 miles apart, with the exception of Katy to Houston which is a distance of 25 miles. Hourly rainfall data for September for the years 1940 to 1945 for all four stations, as published in the *Hydrologic Bulletin*, were analyzed. September was chosen since it was felt that widespread heavy rain was most likely to occur during that month

[6] W. F. McDonald, "Hourly frequency and intensity of rainfall at New Orleans, La.," *Mon. Wea. Rev.*, vol. 57, pp. 1–8; January, 1929.

[7] H. F. Alexander, "A study of the hourly precipitation at Oklahoma City, Okla." *Mon. Wea. Rev.*, vol. 66, pp. 126–130; May, 1938.

in the hurricane season. The analysis is subject to the error that the same storm within the same hour may move over two or more stations and thus give the appearance of simultaneous heavy rains. If anything, this should give an overestimate of the simultaneous occurrence of heavy rains.

In the six years of September rains at any of the four stations, 54 hours were in the heavy rain (0.31 inch per hour or greater) category; only during one hour in six years did heavy rain occur simultaneously at all four stations. There were four hours of simultaneous heavy rain at three stations, nine hours at two stations, and forty hours at only one station. There were twelve hours during which heavy rain occurred at only one station and no rain whatsoever occurred at any of the other stations. It is apparent from this analysis that the heavy rains in the Houston, Tex., area generally were isolated and of small horizontal extent. The chances are better than 3 to 1 that the heavy rain will *not* extend 25 miles in diameter.

In an effort to determine the number of hours by which an "X"-band radar may be reduced below a certain range, the precipitation maps of the Muskingum River Watershed, Ohio, for the year 1938, were studied. The maps show precipitation amounts recorded in the area at half-hourly intervals. A mean intensity of 0.20 inch per half hour or over was considered as causing sufficient attenuation to reduce the range of a high-power "X"-band radar to less than 50 miles. Table I tabulates, by month, the total number of hours of mean precipitation intensity of 0.20 inch per half hour or greater from a point in the center of the area to 50 miles in any direction. Of the twenty hours of heavy rain attenuation over a 50-mile path in any direction, nineteen were in the nature of line squalls or a line of thunderstorms, and the remaining one hour was due to a single thunderstorm of an extremely heavy rain intensity and approximately twenty-five miles in diameter. There was only one hour in which a heavy rain line squall occurred on a line through the station causing attenuation in the two directions 180° apart. A storm, during August, 1938, was a frontal line squall about 20 miles wide of such heavy intensity as to cause attenuation through the width of the storm, causing four consecutive hours of heavy rain attenuation in some direction as the front moved from north to south past the station. The storm, during September, 1938, could be identified as two overlapping line squalls, one line passing to the east of the station, causing severe attenuation in the easterly sector for 2½ hours; the other line squall subsequently passed to the west of the station, causing another 2½ hours of rain attenuation in that sector.

TABLE I

Number of Hours of Mean Precipitation Intensity 0.20 Inches per Half Hour or Greater for 50 Miles in Any Direction in the Muskingum Valley During the Year 1938

Month	No. of Hours	No. of Consecutive Hours
April	½ hour	
May	1	
June	4	2
July	2	1
August	5½	4 and 1
September	5	5
November	2	2
Total	20 hours	

At no time was the rain intensity 0.20 inch per half hour or greater in all directions from the selected point. Since the major portions of heavy rain intensities occurred in line squalls, it is evident that a radar would be able to observe the entire length of the storm before arrival at the radar site; the range would then be reduced to below fifty miles along the frontal direction during the passage of the storm; after passage, the entire storm length would again be visible. The duration of serious rain attenuation would depend on the width of the storm and its velocity.

Conclusion

Heavy rains of 0.40 inch per hour or greater occur on the average from five to thirteen hours during the summer months for the four selected stations in the United States. Such rains, if occurring over the entire path, would generally reduce the range of a high-power "X"-band radar to below 50 miles.

From an analysis of heavy rains in the Texas area during September, it is evident that heavy rains are generally isolated and of small horizontal extent. An analysis of the precipitation in the Muskingum Valley, Ohio, shows that storms, during the year 1938, which would have reduced the range of a high-power "X"-band radar below 50 miles in any direction, totalled 20 hours. These storms were mostly of a line-squall nature. The attenuation produced by such line squalls may be turned to advantage by enabling one to determine, by radar, the mean precipitation intensity along the line squall. Such an analysis has been made by Wexler[8] on the heavy rains in a frontal storm.

The foregoing sample studies of rainfall-intensity distribution and area coverage indicates that the use of high-power "X"-band radar for storm detection is not seriously limited by rainfall attenuation in its function of safe guidance of aircraft through storms.

[8] R. Wexler, "Radar detection of a frontal storm, June 18, 1946," *Jour. Meteor.*, vol. 4, pp. 38–44; February, 1947.

An Inductance-Capacitance Oscillator of Unusual Frequency Stability*

J. K. CLAPP†, FELLOW, I.R.E.

Summary—**An *L-C* oscillator having unusual frequency stability is described and briefly analyzed. The circuit is similar to the familiar Colpitts, with an *L-C* series circuit replacing the inductor. Such a circuit has been used as a piezoelectric oscillator, with the quartz crystal replacing the *L-C* series circuit, but the circuit does not seem to have been previously described as an *L-C* oscillator.**

THE OSCILLATOR considered here has been in use for some years as a quartz-crystal oscillator.[1] Because of the inherent stability associated with the quartz crystal, not much attention has been paid to the possibilities of the circuit for use as an *L-C* oscillator. In such applications the circuit does not seem to have been described previously.

In the usual forms of pi-network oscillators, such as the Hartley and Colpitts, the plate and grid resistances of the tube are shunted across elements of the tuned circuit. Variations of these resistances, as by changes in electrode supply voltages, cause appreciable changes in the frequency of oscillation. In the Hartley circuit it is well known that tapping the tube across only a portion of the tuned-circuit inductance results in a substantial improvement in frequency stability. Such a circuit has been described by Lampkin.[2] This arrangement, indicated schematically in Fig. 1(a), has two practical disadvantages: first, it has a very strong tendency to generate spurious oscillations in the circuit formed by the tube capacitances and the tapped portion of the inductance; and second, the circuit is not too convenient for band-switching operation.

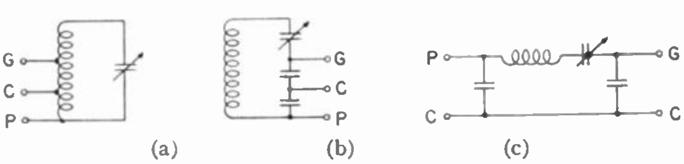


Fig. 1—Showing (a) tube tapped across part of inductive branch; (b) across part of capacitive branch; and (c) the circuit of (b) in filter representation.

The counterpart of this arrangement, with the tube tapped across a portion of the capacitive branch, is indicated in Fig. 1(b). With moderate care in keeping the connections to the tube short, there is practically no tendency toward spurious oscillation. In the configuration shown, only one point would have to be switched for multiband operation, but in practical circuits it is generally better to switch both ends of the inductor. When drawn in the form of a Colpitts circuit, the network appears as in Fig. 1(c), where it is evident that the low-pass (Colpitts) and high-pass (Hartley) arrangements have been replaced by a band-pass circuit.

A schematic diagram of a practical circuit is shown in Fig. 2. In effect, it is a grounded-plate arrangement, and no high d.c. voltages appear at any point of the tuned

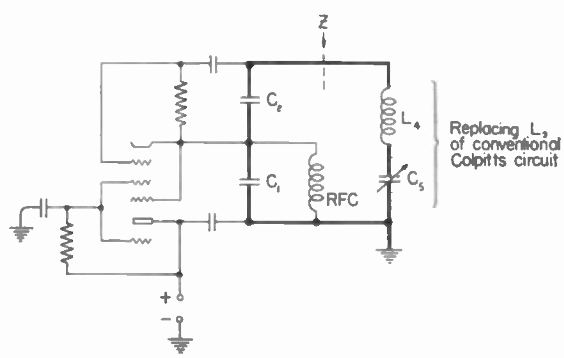


Fig. 2—Schematic diagram of circuit. Capacitances C_1 and C_2 are many times larger than those usually employed in a Colpitts oscillator.

circuit. For convenience, one side of the tuning capacitor is grounded. The coupling capacitances C_1, C_2 are large compared to the tuning capacitance C_5, and are huge compared to the tube capacitances. The radio-frequency choke provides the cathode d.c. return path; preferably the choke should be capacitive at the operating frequency. Variation of the screen-grid voltage provides a convenient means of adjusting the amplitude of oscillation. In normal operation the coupling capacitances are made just as large as reliable operation will permit. Under such conditions the grid and plate swings are only a very few volts.

It is informative to consider the circuits shown in Fig. 3. Starting from the simple circuit of Fig. 3(a), we find for the change in frequency caused by a change in tuning capacitance

$$\frac{df}{f} = -\frac{1}{2}\frac{dC_0}{C_0}. \tag{1}$$

To fix ideas, let the circuit resonate at a frequency of 1 Mc., in which case C_0 might be, conveniently, 100 $\mu\mu$fd.; df/f then is $-dC_0/200$.

* Decimal classification: R355.911.4. Original manuscript received by the Institute, April 16, 1947.
† General Radio Company, Cambridge, Mass.
[1] For example, the General Radio Type-475-C oscillator, used in broadcast frequency monitors, which was introduced in 1940.
[2] G. F. Lampkin, "An improvement in constant-frequency oscillators," PROC. I.R.E., vol. 27, pp. 199–201; March, 1939.

In the form of a Colpitts oscillator, the circuit of Fig. 3(b) might have $C_1 = C_2 = 200\ \mu\mu\text{fd}$. In this case, we will associate a change in capacitance with C_2, representing the input side of the tube. Then we have

$$\frac{df}{f} = -\frac{1}{2}\frac{C_1}{C_1 + C_2}\frac{dC_2}{C_2} = -\frac{1}{2}\frac{C_0}{C_2}\frac{dC_2}{C_2}. \qquad (2)$$

For the values given, $df/f = -dC_2/800$.

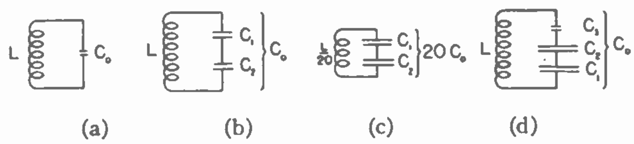


Fig. 3—Illustrating circuit configurations which successively reduce the variation in resonant frequency caused by a given change in capacitance. The given change in capacitance is associated with C_0 in (a) and with C_2 in (b), (c), and (d).

Keeping the same circuit, but utilizing $C_1 = C_2 = 40\ C_0$ in Fig. 3(c) with an inductance $L/20$, gives $df/f = -dC_2/16000$. This represents operation carried to the largest usable capacitances—a "High-C" circuit.

Turning now to the circuit of Fig. 3(d), representing Fig. 2, with $C_1 = C_2 = 40\ C_0$ and $C_3 = 20\ C_0/19$, the total capacitance is C_0. Then $df/f = -dC_2/320{,}000$, representing all of the improvement realized with "high-C" operation augmented by a factor due to the greatness of C_1, C_2 as compared with C_3. If we consider the frequency variation of Fig. 3(b), representing a conventional Colpitts oscillator, as unity, then the frequency variations of Fig. 3(b), (c), and (d) stand as 1, 1/20, 1/400, *for a given capacitance change.*

The circuit of Fig. 3(c) represents an improvement of up to twenty times or so over the conventional Colpitts, Fig. 3(b). In practice, this circuit would require a double variable tuning capacitance of large value. The circuit of Fig. 3(d) gives a further improvement of twenty times or so, using a single variable tuning capacitance of more usual value.

While the above development is by no means complete for determining the stability of frequency of an oscillator, it nevertheless indicates very well the relative improvement attainable with respect to changes in input capacitance of the tube as a result of temperature changes of tube structure or of changes in supply voltages.

A brief linear analysis is helpful in showing the improvement in stability of frequency which is possible using the circuit of Fig. 2. A simple approach is to divide the circuit at the dotted line and write the expression for the impedance Z seen looking into the circuit.

$$Z = Z_g + \frac{r_p Z_p}{r_p + Z_p} + \frac{\mu}{r_p} Z_g \frac{r_p Z_p}{r_p + Z_p} + Z_3. \qquad (3)$$

Taking Z_g as r_g in parallel with X_2, Z_p as X_1, alone, separating reals and imaginaries and placing them equal to zero, we have, for a conventional Colpitts circuit,

$$-\frac{\mu}{r_p} X_1 X_2 + \frac{X_1}{r_p}(X_3 - X_2) + \frac{X_2}{r_g}(X_3 - X_1) + R_3\left(1 - \frac{X_1 X_2}{r_g r_p}\right) = 0 \qquad (4)$$

$$-X_1\left(1 + \frac{R_3}{r_p}\right) - X_2\left(1 + \frac{R_3}{r_g}\right) + X_3\left(1 - \frac{X_1 X_2}{r_g r_p}\right) = 0. \qquad (5)$$

In (4) for reals, we can substitute with small error $(X_3 - X_2) = X_1$ and $(X_3 - X_1) = X_2$, and obtain

$$-\frac{\mu}{r_p} X_1 X_2 + \frac{X_1^2}{r_p} + \frac{X_2^2}{r_g} + R_3 = 0 \qquad (6)$$

where the first term represents the negative resistance developed in the circuit by the tube; the next two terms represent the equivalent series resistances of the tube resistances in parallel with the coupling capacitances; and the last term is the resistance of the inductor (neglecting the very small correction term).

In (5) for imaginaries, the terms in r_g are the ones causing a change in frequency with change in supply voltage. These are $X_1X_2X_3/r_g r_p$ and R_3X_2/r_g. With large reactances and low resistances, the first of these entirely overshadows the second. On reduction of the reactances to the lowest possible values, keeping the resistances of the tube as high as possible, however, the second becomes the predominant term.

An important analysis of the stability of this class of oscillators is given by Fair,[3] but the effects of the resistance of the inductor are not taken into account. Following Fair's method and including the coil resistance, we obtain

$$\frac{\delta\omega}{\delta V} = \frac{\dfrac{1}{\mu}\dfrac{\partial\mu}{\partial V}\left[\dfrac{X_1X_2X_3}{r_g r_p} + \dfrac{R_3 X_2}{r_g}\right]}{\left[1 - \dfrac{X_1}{\mu X_2} - \dfrac{R_3 r_p}{\mu X_1 X_2}\right]\left[-\left(1 + \dfrac{R_3}{r_p}\right)\dfrac{\partial X_1}{\partial\omega} - \left(1 + \dfrac{R_3}{r_g}\right)\dfrac{\partial X_2}{\partial\omega} + \dfrac{\partial X_3}{\partial\omega}\right]} \qquad (7)$$

for $\quad \mu = f_1(V), \quad r_g = f_2(a), \quad X_1 + X_2 + X_3 = f_3(\omega), \quad r_p = \text{constant}.$

[3] I. E. Fair, "Piezoelectric crystals in oscillator circuits," *Bell Sys. Tech. Jour.*, vol. 24, pp. 161–215; April, 1945.

This is in the form given by Fair and differs from his result only by the terms in R_3.

The term $-R_3r_p/\mu X_1X_2$, in the first factor in the denominator, is the ratio of the coil resistance to the negative resistance developed in the circuit. From the equation for reals it is seen that $\mu X_1X_2/r_p$ exceeds R_3 by only a small amount. The ratio is then near unity and the first factor reduces to $-X_1/\mu X_2$. The second factor in the denominator (neglecting the terms in R_3 which are small) is equal to $2L_3$. If we write X_3/Q_3 for R_3, (7) reduces to

$$\frac{\delta\omega}{\delta V}=\frac{\frac{1}{\mu}\frac{\partial\mu}{\partial V}\left[\frac{X_1X_2X_3}{r_gr_p}+\frac{X_2X_3}{Q_3r_g}\right]}{-\frac{X_1}{\mu X_2}[2L_3]} \tag{8}$$

$$=-\frac{X_2}{X_1}\frac{\partial\mu}{\partial V}\left[\frac{1}{2\omega C_1C_2r_gr_p}+\frac{1}{2Q_3C_2r_g}\right] \tag{9}$$

$$=-\frac{\partial\mu}{\partial V}\left[\frac{1}{2\omega n^2C_0^2r_gr_p}+\frac{1}{2Q_3nC_0r_g}\right] \tag{10}$$

for $C_1 = C_2 = nC_0$.

In the ordinary Colpitts circuit the first term predominates. If the tuning capacitances C_1, C_2 are increased as much as possible and maintain oscillation ($C_1=C_2$, L_3' adjusted for same frequency), the first term decreases as n^2 and stability is improved ("high-C" circuit).

If L_3' is replaced by L_4 and C_5 in series, (8) becomes

$$\frac{\delta\omega}{\delta V}=\frac{\frac{1}{\mu}\frac{\partial\mu}{\partial V}\left[\frac{X_1X_2(X_4-X_5)}{r_pr_p}+\frac{X_2X_4}{Q_4r_g}\right]}{-\frac{X_1}{\mu X_2}[2L_4]} \tag{11}$$

$$=-\frac{\partial\mu}{\partial V}\left[\frac{1}{2\omega n^2C_0^2r_gr_p}\frac{L_3'}{L_4}+\frac{1}{2Q_4nC_0r_g}\right]. \tag{12}$$

for $C_1 = C_2 = nC_0$

Note that, for $\omega=$constant, (X_4-X_5) in the numerator is equal to $X_3=\omega L_3'$ where L_3' is that value of inductance which would tune to the desired frequency with capacitances C_1 and C_2 only. L_4 can be many times L_3', in which case the effect of the first term is still further reduced in the ratio of L_3'/L_4. The second term s unaltered for the same value of C_2, as long as $Q_4=Q_3$. This condition may be difficult to meet as L_4 is made larger and larger compared with L_3'. However, a large reduction in the first term can be made, with small increase in the second term in any case.

Summarizing, these results show that in (11) the stability is improved when all reactances in the numerator are kept small, r_g and r_p are kept high, and the rate of change of reactance of the circuit, in the denominator, is made as large as possible. The Q of L_4 should be kept as high as possible.

This class of oscillator has been described by Jefferson[4] as being only "potentially stable," meaning that the change in frequency can only approach zero, but never reach zero, no matter how many elements are used in each branch of the pi network.

Llewellyn[5] has given many circuits with the conditions for frequency stabilization. In practice, however, many of these conditions are modified by the effects of tube capacitances, capacitances of coils, etc., so that either the conditions for zero frequency change are appreciably altered or become critical if the frequency is changed. Also, many of the circuits shown are not readily adaptable to variable-frequency or band-switching operation.

The result is that for many practical applications an oscillator circuit which can be made to approach perfect stability in a *noncritical* manner, even though perfect stability cannot be achieved, is preferable to a circuit which can be made perfectly stable but only by critical adjustments or by adjustments which must be changed when the frequency is changed.

The long-time stability of this circuit depends almost entirely on the permanence of the elements L_4, C_5, and their temperature coefficients. The variations in frequency caused by temperature changes will generally be found to be much simpler and more straightforward than in conventional circuits because of the fact that the tube effects have been effectively eliminated.

The circuit described above can be set up with reactances of 70 to 100 ohms for X_1 and X_2, these being about one-fortieth of the coil or tuning-capacitance reactance. The circuit has been operated successfully at frequencies ranging from 10 kc. to over 100 Mc. It has been applied in heterodyne frequency meters, oscillators for beat-frequency oscillators, master oscillators in amateur transmitters, and as a frequency-modulation generator. In the latter case, the frequency swing is caused by varying C_1, the center frequency being set by the series tuning capacitor C_5.

The stabilities obtained depend on the frequency and reactances used. Frequency changes of less than 1 part per million to a very few parts per million for changes in supply voltages of ± 15 per cent have been obtained. Interchanging tubes of the same type causes practically no change in frequency.

[4] H. Jefferson, "Stabilization of feedback oscillators," *Wireless Eng.*, vol. 22, pp. 384–389; August, 1945.

[5] F. B. Llewellyn, "Constant frequency oscillators," Proc. I.R.E., vol. 19, pp. 2063–2094; December, 1931.

The Comb Antenna*

RALPH GRIMM†, ASSOCIATE, I.R.E.

Summary—An analysis of a comb antenna for reception of vertically polarized medium-frequency waves is given. The formula can be applied to any array in which the elements are arranged in a line and in which the current ratios and phase relations are known. Close coupling of the elements is considered, rather than very loose capacitive coupling. The condition of close coupling greatly increases the directivity and signal level by adjusting the line velocity to an optimum value. The antenna is especially suited for reception of Loran signals.

THE PROBLEM of receiving weak vertically polarized ground waves or low-angle sky waves in the broadcast and medium-frequency bands has been one not easily solved. Most directional arrays become impractically large at these frequencies.[1-2] The Beverage or wave antenna will give good results if used over soil with rather poor conductivity, but even then it has the disadvantage of low output and greater response to low-angle sky waves than to the ground wave. This is highly objectional in the reception of 'phone or c.w. signals, as the area of selective fading moves closer to the transmitting station. This fading is not present in the reception of short-duration pulses, since the pulses are received in their entirety before the arrival of the sky-wave signal. Response to sky waves is highly undesirable when receiving weak signals, because the nighttime noise is propagated by low-angle reflections and the signal-to-noise ratio is decreased.

The comb antenna[3] is shown in Fig. 1. It consists of a number of vertical elements arranged in line with the transmitting station. A transmission line (coaxial) is used to connect all vertical elements together. This line is terminated at the end nearest the transmitter in an impedance to reduce standing waves, and a coupling unit is inserted at the other end to match the antenna to a transmission line. This impedance is not the characteristic impedance of the line, but rather a new impedance caused by the addition of the vertical elements. These elements will change the effective shunting impedance across the line and, thus, the line velocity. The base reactance of each element has a definite optimum value and will be described in detail. One must also consider that each element introduces a certain amount of loss.

A general equation will be obtained and assumptions then applied for particular cases. Comparison of the array will be made to a single element.

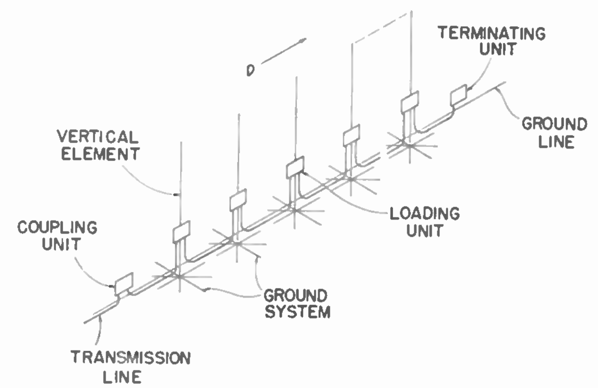


Fig. 1

Assume:

$$h \ll \lambda$$

$$a \ll \lambda$$

where h is the height of the elements and a is the spacing. All elements have identical base impedance.

Let

V = velocity of propagation along loaded line
V_0 = velocity of propagation along unloaded line
c = velocity of propagation in air.

Considering the array as being driven, the current in each element is

$$i_1 = I \tag{1a}$$

$$i_2 = I\epsilon^{(-\alpha a - j\theta)} \tag{1b}$$

$$i_3 = I\epsilon^{2(-\alpha a - j\theta)} \tag{1c}$$

$$i_n = I\epsilon^{(n-1)(-\alpha a - j\theta)}, \tag{1d}$$

where

$$\theta = \frac{2\pi a}{\lambda}\left(\frac{C}{V}\right)$$

and αa = attenuation between elements.

This notation is shown in Fig. 2.

The field intensity at a point P in space will be the vector sum of the fields produced by each element. In the line of the array, these fields are:

$$\mathcal{E}_1 = KI \tag{2a}$$

$$\mathcal{E}_2 = KI\epsilon^{(-\alpha a - j\theta + j(2\pi a/\lambda))} \tag{2b}$$

$$\mathcal{E}_3 = KI\epsilon^{2(-\alpha a - j\theta + j(2\pi a/\lambda))} \tag{2c}$$

$$\mathcal{E}_n = KI\epsilon^{(n-1)(-\alpha a - j\theta + j(2\pi a/\lambda))} \tag{2d}$$

* Decimal classification: R125.1×R325.11. Original manuscript received by the Institute, August 16, 1946; revised manuscript received, September 26, 1947.

† Clark Instrument Corporation, Silver Spring, Md.

[1] G. C. Southworth, "Certain factors affecting the gain of directive antennas," PROC. I.R.E., vol. 18, pp. 1502–1537; September, 1930.

[2] E. J. Sterba, "Theoretical and practical aspects of directional transmitting systems," PROC. I.R.E., vol. 19, pp. 1184–1216; July, 1931.

[3] The antenna described here is similar, except for polarization and phase velocity, to that described in the paper by H. H. Beverage and H. O. Peterson, "Diversity receiving system of RCA Communications Inc. for radiotelegraphy," PROC. I.R.E., vol. 19, pp. 31–562; April, 1931.

where K is a constant depending on element radiation efficiency.

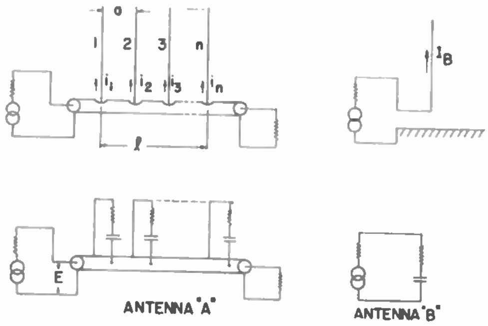


Fig. 2

The addition of these fields is shown in Fig. 3, which is a section of a logarithmic spiral in which

$$\frac{\rho_0}{\rho_1} = \frac{\rho_2}{\rho_3} = \frac{\rho_4}{\rho_5} = \cdots$$

$$\frac{\rho_0}{\rho_1} = \frac{\mathcal{E}_1}{\mathcal{E}_2} = \frac{\mathcal{E}_2}{\mathcal{E}_3} = \cdots$$

$$\phi = \phi_1 = \phi_2 = \cdots$$

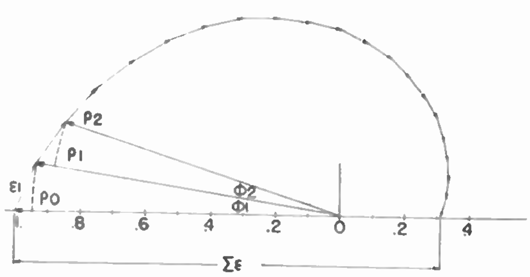


Fig. 3

where ϕ = phase difference between fields at point of reception.

From the law of cosines,

$$\mathcal{E}_1^2 = \rho_0^2 + \rho_1^2 - 2\rho_0\rho_1 \cos\phi \tag{3a}$$

$$\left(\frac{\mathcal{E}_1}{\rho_0}\right)^2 = 1 + \left(\frac{\rho_1}{\rho_0}\right)^2 - 2\frac{\rho_1}{\rho_0}\cos\phi \tag{3b}$$

$$\left(\frac{\mathcal{E}_1}{\rho_0}\right)^2 = 1 + \epsilon^{-2\alpha a} - 2\epsilon^{-\alpha a}\cos\phi \tag{3c}$$

where α = attenuation in nepers per meter, or $\epsilon^{\alpha a}$ = current ratios in adjacent elements.

$$\sum \mathcal{E}^2 = \rho_0^2 + \rho_n^2 - 2\rho_0\rho_n \cos n\phi \tag{4a}$$

$$\left(\frac{\sum \mathcal{E}}{\rho_0}\right)^2 = 1 + \left(\frac{\rho_n}{\rho_0}\right)^2 - 2\frac{\rho_n}{\rho_0}\cos n\phi \tag{4b}$$

$$\left(\frac{\sum \mathcal{E}}{\rho_0}\right)^2 = 1 + \epsilon^{-2n\alpha a} - 2\epsilon^{-n\alpha a}\cos n\phi. \tag{4c}$$

Combining (3c) with (4c),

$$\frac{\sum \mathcal{E}}{\mathcal{E}_1} = \sqrt{\frac{1 + \epsilon^{-2n\alpha a} - 2\epsilon^{-n\alpha a}\cos n\phi}{1 + \epsilon^{-2\alpha a} - 2\epsilon^{-\alpha a}\cos\phi}}. \tag{5}$$

ϕ has been defined as the phase difference between the fields from individual elements as measured at point P. If this point is in the line of the array, ϕ is determined by the line velocity. In other directions this phase difference becomes smaller by

$$\frac{2\pi a}{\lambda}\cos\delta$$

where δ = angle of departure. Then (5) becomes

$$\sum \mathcal{E} = \mathcal{E}_1 \sqrt{\frac{1 + \epsilon^{-2n\alpha a} - 2\epsilon^{-n\alpha a}\cos n\left[\frac{2\pi a}{\lambda}(1-\cos\delta) + \phi\right]}{1 + \epsilon^{-2\alpha a} - 2\epsilon^{-\alpha a}\cos\left[\frac{2\pi a}{\lambda}(1-\cos\delta) + \phi\right]}}. \tag{6}$$

This is the general formula for the radiation pattern of a comb antenna. If the attenuation can be neglected, one obtains:

$$\sum \mathcal{E} = \mathcal{E}_1 \frac{\sin\frac{1}{2}n\left[\frac{2\pi a}{\lambda}(1-\cos\delta) + \phi\right]}{\sin\frac{1}{2}\left[\frac{2\pi a}{\lambda}(1-\cos\delta) + \phi\right]}, \tag{7}$$

or, if the number of elements is large,

$$\sum \mathcal{E} = \mathcal{E}_1 n \frac{\sin\left[\frac{\pi l}{\lambda}(1-\cos\delta) + \frac{\beta}{2}\right]}{\frac{\pi l}{\lambda}(1-\cos\delta) + \frac{\beta}{2}} \tag{8}$$

where $\beta = n\phi$ and l = length of array.

Considering now the field-strength gain in the line of the array as compared to a single element, refer to Fig. 2.

The power input to antenna A is

$$P_A = \frac{E^2}{Z} = \frac{(IX_A)^2}{Z}$$

where

X_A = base reactance of each element

I = element current

Z = line impedance.

The power to antenna B is

$$P_B = I_B{}^2 R_B.$$

The two antennas produce equal field intensities when

$$I_B = I\sqrt{\frac{1+\epsilon^{-2n\alpha a}-2\epsilon^{-n\alpha a}\cos n\phi}{1+\epsilon^{-2\alpha a}-2\epsilon^{-\alpha a}\cos\phi}}. \tag{9}$$

Let the radical term be K:

$$I_B = IK.$$

The power ratio is then

$$\frac{P_B}{P_A} = \frac{Z}{I^2 X_A{}^2} I^2 R_B K^2$$

$$= \frac{Z R_B}{X_A{}^2} K^2. \tag{10a}$$

Since the radiation resistance of a short vertical above ground[4] is

$$R_B = 395\left(\frac{h}{\lambda}\right)^2,$$

the power gain is

$$\frac{P_B}{P_A} = 395\left(\frac{h}{\lambda}\right)^2 \frac{Z}{X_A{}^2} K^2. \tag{10b}$$

If the attenuation can be neglected, K is reduced and the power gain becomes

$$\frac{P_B}{P_A} = 395\left(\frac{h}{\lambda}\right)^2 \frac{Z}{X_A{}^2}\left(\frac{\sin\frac{1}{2}n\phi}{\sin\frac{1}{2}\phi}\right)^2. \tag{10c}$$

Further, if the number of elements is great,

$$\frac{P_B}{P_A} = 395\left(\frac{h}{\lambda}\right)^2 \frac{Z}{X_A{}^2}\left(\frac{\sin\frac{1}{2}n\phi}{\frac{1}{2}\phi}\right)^2. \tag{10d}$$

Considering a practical example, let the length be 2λ, the attenuation zero, the number of vertical elements large, and the velocity along the line equal the velocity in space. The resulting horizontal characteristic is shown in Fig. 4. The effect of reducing the line velocity to produce 180° phase difference between the line wave and the space wave is shown in Fig. 5. The optimum value of phase shift for maximum discrimination against random distribution of noise is approximately 180°.[5]

[4] I.R.E. Standards on "Transmitters and Antennas, Methods of Testing," reprinted 1942. This assumes "effective height" is equal to ½ actual height.

[5] W. W. Hansen and J. R. Woodyard, "A new principle in directional antenna design," Proc. I.R.E., vol. 26, pp. 333–346; March, 1938.

A practical antenna will now be considered which was built and tested by the United States Coast Guard for use in its Loran system. The operating frequency is 1.95 Mc. and the bandwidth must be sufficient to prevent discrimination against side bands of a pulse 40 microseconds in width. The length of the array is determined

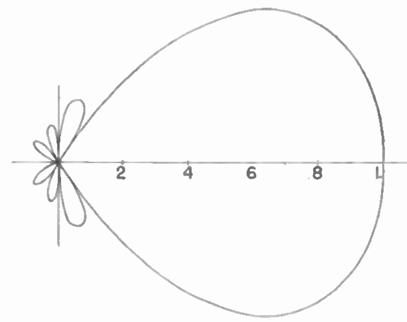


Fig. 4

by frequency and the characteristics of the transmission line. If the array becomes quite long, the optimum velocity of propagation along the loaded line approaches that of free space and tuning becomes difficult, while a short array suffers from poor directivity. A length of 1200 feet was chosen with elements spaced 60 feet apart, or a total

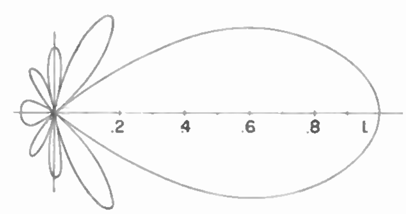


Fig. 5

number of 21 elements. For reasons of economy the vertical elements were supported by standard telephone poles 35 feet in length. A lower-impedance element was obtained by using an 18-foot whip on top of the pole, giving a height of 45 feet. It was not desired that the length of cable-connecting elements be equal to the spacing, as this would require supporting poles along its length. The cable was then made 65 feet long. The total phase shift in the array for maximum directivity is 180°, or 9° between elements. A special type of flexible coaxial cable covered with a steel jacket and pitch was used so that conduit would not be required. Its characteristics are as follows:

Characteristic impedance = 52 ohms
Capacity per foot = 31 $\mu\mu$fd.
Velocity of propagation = 0.65 c.

The electrical length of the unloaded line between elements is then 71.5°, while the desired length is 51.8°. The velocity of propagation must then be increased to 89.7 per cent c or approximately 90 per cent c. This requires that each element present an inductive reactance of 85 ohms to the coaxial line. This loading increases the characteristic impedance of the line to 72 ohms, which is the proper load and terminating resistances. The measured impedance of an element was $29-j650$ ohms. The cable attenuation was reduced and the bandwidth increased by shunting each element with a 200-$\mu\mu$fd. capacitor. A loading coil of 735 ohms was then inserted between the element and the cable. The ground system consists of four 50-foot radials with outer ends terminated in 3-foot ground rods. The measured attenuation was approximately 10 db, and the calculated horizontal directivity is shown in Fig. 6. Unfortunately the terrain was dense swamp land and water, and a check of directivity could not be made; however, direct comparison was made to a 70-foot vertical antenna, and the signal-to-noise ratio in each antenna used to evaluate performance. Fig. 7 shows the signal-to-noise ratios of this type comb and vertical antenna with signals arriving from the Northeast; it is the average of readings during the first five months of 1947. This experimental antenna is located at the United States Coast Guard Loran station at Bodie Island, North Carolina.

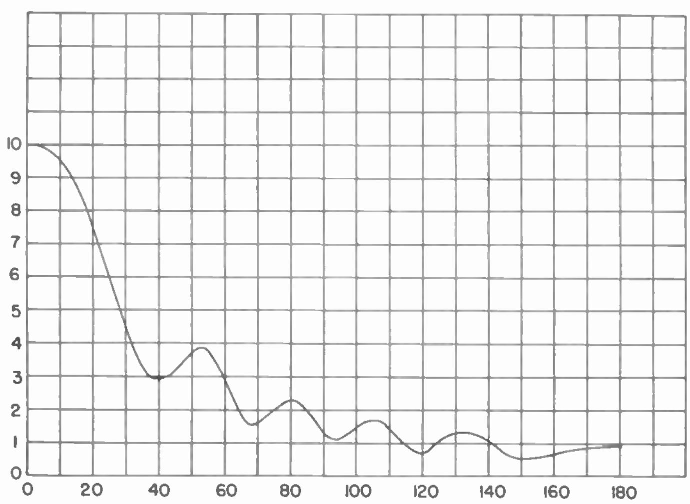


Fig. 6

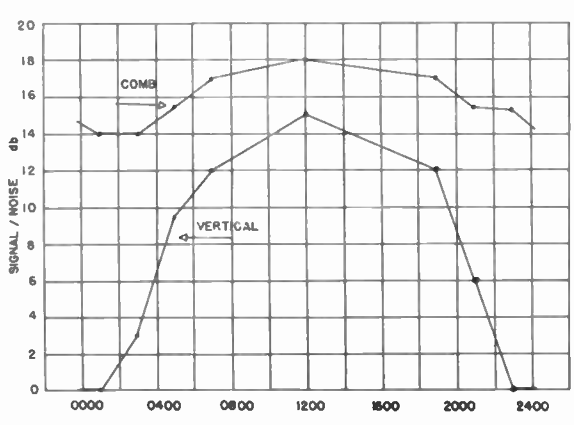


Fig. 7

As stated previously, the base resistance of each element is 29 ohms. This represents considerable loss, and it was found that the resistance could be reduced to 19 ohms by supporting the wire 4 feet from the pole. This type of construction is used in a more recent antenna, as shown in Figs. 8 and 9.

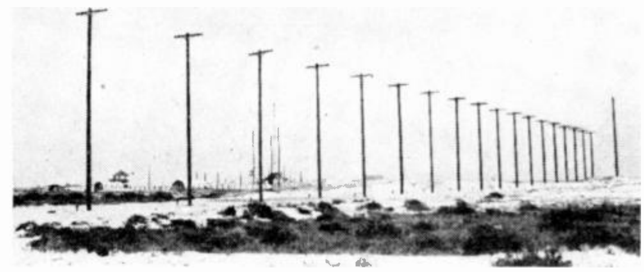


Fig. 8

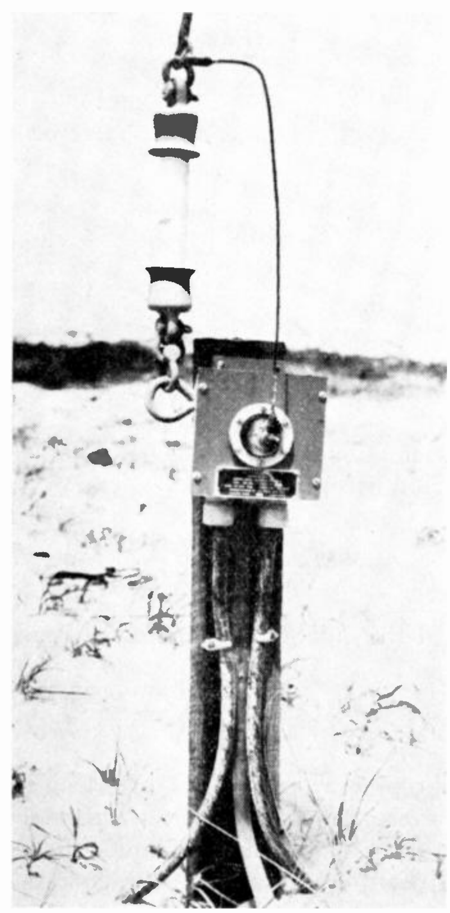


Fig. 9

ACKNOWLEDGMENT

The author wishes to express appreciation to the Communications Engineering Section of the United States Coast Guard for releasing this material for publication, and to their Electronics Field Test Station, with whose facilities experimental arrays were built and tested.

Correspondence

Low-Level Atmospheric Ducts*

We were very interested in the paper by Katzin, Bauchman, and Binnian[1] on the effect of low-level atmospheric ducts on the propagation of 3- and 9-centimeter waves. In this paper the authors show that such low-level ducts varying between about 20 and 50 feet in height do exist over the ocean, and that they give abnormally long ranges for centimetric equipment located low enough to be within the ducts. It may be of interest to describe how we were forced to the conclusion that such low-level ducts must exist over the sea.

We have made a systematic study of the records of field strength taken over a period of three years for a 60-mile overseas path in the Irish Channel. We could get a consistently good correlation between theoretical and experimental results only when we assumed the presence of low-level surface ducts for up to about 70 per cent of the time. The interesting feature about this result, which was published in an official report, is that we arrived at it purely from a study of the field-strength measurements. Low-level meteorological measurements had been made for part of the period by the Royal Naval Meteorological Service, but they were not available when the radio results were being analyzed. Later, when they were obtained, the presence of the low-level ducts, their heights of up to about 40 feet, and the percentage of time during which they occurred, all agreed well with the values predicted previously.

Some of the other results obtained by us do not quite agree with those found by the authors; for example, we did not observe that high winds were associated with higher ducts, nor did we find that there was any critical difference between the propagation of 3- and 9-centimeter waves. From our measurements we concluded that the presence of low-level ducts materially affects the propagation of both meter- and centimeter-wavelength radiation for transmission paths both below and immediately above the ducts. Incidentally, we also found some correlation between the difference between air and sea temperature and signal strength. This last result follows, of course, from the type of duct formed under such meteorological conditions. We hope shortly to publish our results in greater detail.

We are indebted to the Chief Scientist, Ministry of Supply, for permission to publish this note.

J. S. McPetrie
B. J. Starnecki
Signals Research and Development Est.
Somerford, Christchurch
Hants, England

* Received by the Institute, November 4, 1947.
[1] Martin Katzin, Robert Bauchman, and William Binnian, "3- and 9-centimeter propagation in low ocean ducts," Proc. I.R.E., vol. 35, pp. 891–906; September, 1947.

Radar Reflections from the Lower Atmosphere*

A letter from W. B. Gould,[1] describes 1.25-cm. "Angels." Mr. Gould says: "The short duration [of these echoes] may, in part, be explained by the possible motion of the reflecting medium through the relatively narrow radar beam produced by the equipments."

3-cm. and 10-cm. "Angels" do indeed move, beyond any doubt whatever. They move in fairly straight and level courses, usually running a little faster than the surface wind, sometimes running across or against it. Once in a while one stops for a few seconds. We have frequently tracked a single one of these echoes for 5 or 10 minutes until it vanished downwind. When a fully automatic tracking antiaircraft radar follows an "Angel," the motions are what you might call majestic.

Mr. Gould is to be complimented for presenting an excellent set of photographs of "Angel" signals.

Millard W. Baldwin, Jr.
Bell Telephone Laboratories, Inc.
New York, N. Y.

* Received by the Institute, October 23, 1947.
[1] W. B. Gould, "Radar reflections from the lower atmosphere," Proc. I.R.E., vol. 35, p. 1105; October, 1947.

Contributors to the Proceedings of the I.R.E.

Edward W. Allen, Jr.

Edward W. Allen, Jr. (M'44) was born on February 14, 1903, at Portsmouth, Va. He received the B.S. degree in electrical engineering from the University of Virginia in 1925, and the LL.B. degree from George Washington University in 1933.

From 1925 to 1927 Mr. Allen was employed by Westinghouse as a student engineer and research assistant. He joined the Chesapeake and Potomac Telephone Company, in Washington, D. C., in 1928 as engineering assistant. From 1930 to 1935 he worked for the United States Patent Office as junior and assistant patent examiner in telephony, telegraphy, facsimile, and television. Since 1935 Mr. Allen has been associated with the Federal Communications Commission as assistant chief of the technical information division in the engineering department. He is a member of Tau Beta Pi.

❖

J. K. Clapp

J. K. Clapp (A'24–M'28–F'33) was born on December 30, 1897, at Denver, Colo. He was with the Marconi Wireless Telegraph Company from 1914 to 1916 and served with the United States Navy from 1917 to 1919. From 1918 to 1919 he served the Foreign Service of the U. S. Government, and in 1920 became associated with the Radio Corporation of America. He received the B.S. degree from the Massachusetts Institute of Technology in 1923, and from 1923 to 1928 was an instructor in communications at this school, obtaining the M.S. degree in 1926.

Mr. Clapp has been with the engineering department, General Radio Company, Cambridge, Mass. from 1928 to date, working on frequency standards and measurements. He has served on various committees of the I.R.E. since 1931.

❖

Henry B. DeVore (A'35–M'40–SM'43) was born on December 20, 1907, at Monongahela, Pa. He received the B.S. degree in physics in 1926, and the M.S. degree in

Henry B. DeVore

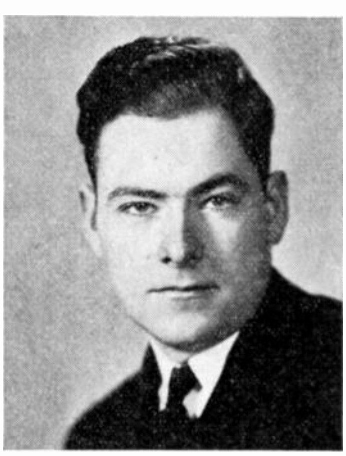
1927, from the Pennsylvania State College. From 1927 to 1931 he was employed at the experimental station of E. I. du Pont de Nemours and Company, and from 1934 to 1945 in the research laboratories of the Radio Corporation of America.

From 1945 to 1947, Dr. DeVore was associated with the Laboratory for Advanced Engineering of Remington Rand, Inc. Since May, 1947, he has been engaged in research at RCA Laboratories, Princeton, N. J. Dr. DeVore is a member of the American Physical Society and of Sigma Xi.

❖

Ralph Grimm (J'41–A'43) was born in Stanley, Va., on August 3, 1922. He was graduated from the Capitol Radio Engineering Institute in Washington, D. C., in 1941, and became a member of the instructing staff. He joined the United States Coast Guard in 1943, and served with the Communications Engineering Section until 1946. During most of his three and one-half years of service, Mr. Grimm specialized in the field application of the Loran system.

In 1946, Mr. Grimm joined the Air Track Manufacturing Company as project engineer. He is now engaged in electronic instrument development for the Clarke Instrument Corporation, in Silver Spring, Md.



Ralph Grimm

Harris F. Hopkins was born in Bath, Maine, on October 27, 1902. He received the E. E. degree from Brooklyn Polytechnic Institute in 1932. As a member of the technical staff of Bell Telephone Laboratories, Inc., he has been concerned chiefly with the development of electroacoustic instruments.

Harris F. Hopkins

❖

Joseph Weinstein was born in New York, N. Y., in 1915. He is a graduate of the College of the City of New York, receiving the B.S. degree in mathematics in 1936 and the M.S. degree in education in 1937. His postgraduate studies were in mathematics and statistics at New York University, the College of the City of New York, and Rutgers University.

Joseph Weinstein

Since 1942, Mr. Weinstein has been employed by the Signal Corps Engineering Laboratories as a research analyst and statistician. He was previously employed as a teacher of mathematics in New York schools and an employment interviewer in the New York State Department of Labor. He is a member of the American Statistical Association, the Institute of Mathematic Statistics, and the American Society for Quality Control.

Norman R. Stryker

Norman R. Stryker was born in Trenton, N. J., in November, 1899. He received the B.S. degree in electrical engineering from the University of Illinois in 1921. As a member of the technical staff of the Bell Telephone Laboratories, Inc., since graduation, he has been concerned with the development of sound-picture equipment and techniques, the application of acoustic principles to the telephone plant, and the development of electroacoustic instruments. During the war, he was assigned to a still-secret project for the Armed Forces.

❖

Raymond Wexler was born in Fall River, Mass., on July 12, 1914. He received the B.A. degree at Harvard University in 1936, and the M.S. degree in meteorology at the Massachusetts Institute of Technology in 1939.

From 1939 to 1941, Mr. Wexler was a meteorologist with Northwest Airlines at Spokane, Wash. During 1941, he taught meteorology to Air Corps Cadets at Hancock College, Calif., and in 1942, at the University of Chicago. Since 1943, he has been employed as a physicist on radar and meteorology problems at the Signal Corps Engineering Laboratories, Belmar, N. J. He is a professional member of the American Meteorological Society and the New York Academy of Sciences.

Raymond Wexler

Institute News and Radio Notes

1948 I.R.E. National Convention Program

HOTEL COMMODORE and GRAND CENTRAL PALACE—MARCH 22–25

Monday, March 22, 1948

9:00 A.M.–5:30 P.M.—Registration at Hotel Commodore and Grand Central Palace

10:30–12 A.M.—Annual Meeting; Principal Address: "An Engineer in the Electronics Industry—Prospects, Preparation, Pay," H. B. Richmond; Grand Ballroom, Hotel Commodore

11:00 A.M. Opening of the Radio Engineering Show at Grand Central Palace

11:00 A.M.–9:00 P.M.—Radio Engineering Show, Grand Central Palace

2:30–5:00 P.M.—"Frequency Modulation," "Networks," "Systems I," "Navigational Aids," and "Antennas I."

6:00–8:00 P.M.—Cocktail Party, Hotel Commodore

8:00 P.M.—Sections Committee Meeting, Hotel Commodore

Tuesday, March 23, 1948

9:00 A.M.–5:30 P.M.—Registration

10:00–12:30 P.M.—"Amplifiers," "Systems II," "Electronics I—Tube Design and Engineering," and "Antennas II."

10:00 A.M.–9:00 P.M.—Radio Engineering Show, Grand Central Palace

12:30 P.M.—President's Luncheon, honoring Dr. Shackelford, Grand Ballroom, Hotel Commodore. Featured speaker: Wayne C. Coy, Chairman, Federal Communications Commission. Guest of Honor, B. S. Shackelford; Toastmaster, W. R. G. Baker.

2:30–5:00 P.M.—"Superregeneration," "Transmission," "Nuclear Studies," "Electronics II—Industrial Application of Tubes and Electronic Circuits," and "Components and Supersonics."

8:00–10:30 P.M.—Symposium: Nuclear Science. (See page 366.)

Wednesday, March 24, 1948

9:00 A.M.–5:30 P.M.—Registration

10:00 A.M.–6:00 P.M.—Radio Engineering Show, Grand Central Palace

10:00 A.M.–12:30 P.M.—Symposium: Advances Significant to Electronics. (See page 366.)

2:30–5:00 P.M.—"Television," "Synthetic Crystals" (special added session), "Broadcasting and Recording," "Electronics III—Tube Manufacture," and "Measurements I—V.H.F., U.H.F., and S.H.F."

7:00 P.M.—Annual I.R.E. Banquet (dress optional), Hotel Commodore. Toastmaster: W. L. Everitt, University of Illinois. Awarding of the Medal of Honor, the Morris Liebmann Memorial Prize, the Browder J. Thompson Memorial Award, and Fellow Awards.

Thursday, March 25, 1948

9:00 A.M.–5:30 P.M.—Registration

10:00 A.M.–9:00 P.M.—Radio Engineering Show, Grand Central Palace

10:00 A.M.–12:30 P.M.—"Computers I—Computing Systems," "Propagation," "Electronics IV—New Forms of Tubes," and "Measurements II."

2:30–5:00 P.M.—"Computers II—Computer Components," "Microwaves," "Receivers," and "Active Circuits."

WOMEN'S ACTIVITIES

Monday, March 22, 1948

9:00 A.M.—Registration

2:15 P.M.—Sightseeing Trip of Lower New York $2.30

Tuesday, March 23, 1948

9:15 A.M.—Trip to United Nations; luncheon available at United Nations Cafeteria $2.25

or

10:30 A.M.—Trip to Frick Collection, with 3/4-hour art lecture No charge

3:30–5:30 P.M., Tea, I.R.E. Headquarters Building, 1 East 79 Street No charge

Wednesday, March 24, 1948

10:45 A.M.—Fashion talk by Miss Nash, of Bonwit-Teller, Inc., New York, N. Y., Hotel Commodore No charge

2:15 P.M.—Matinee, choice of "The Heiress" or "High Button Shoes" $3.00

Thursday, March 25, 1948

9:15 A.M.–3:30 P.M.—All-Day Trip to West Point, including luncheon at Thayer Hotel $6.00

ANNUAL MEETING

MONDAY, MARCH 22, 10:30 A.M.

This opening meeting of the convention is for the entire membership. The meeting will feature the following address:

AN ENGINEER IN THE ELECTRONICS INDUSTRY—PROSPECTS, PREPARATION, PAY

H. B. RICHMOND
(General Radio Co., Cambridge, Mass.)

A short review of the development of the electronics industry up to World War II, and the violent impact of that war on the industry, will be followed by an analysis of opportunities available in the industry during the next decade. The type of collegiate and supplemental instruction that should be given to fit engineers and research personnel for opportunities within the industry will be discussed with special reference to position adaptability. The place of industry-collegiate co-operative courses both from the student and employer viewpoint will be mentioned. A discussion of salaries and pay methods will conclude the paper.

SPECIAL SESSION

"Nuclear Science"

MARCH 23, 8:00–10:30 P.M.

Chairman, L. R. HAFSTAD
(Research and Development Board, Washington, D. C.)

1. The Atomic Energy Problem and the Engineer

L. R. HAFSTAD

2. The Program of the Atomic Energy Commission

(To be presented by a representative of the AEC. This paper will deal with the engineering aspects of atomic energy.)

3. Electronic Problems of the Atomic Energy Program

(To be presented by a representative of the AEC.)

4. Biological Effects of Radiation and Consideration of Protection Problems

J. Z. BOWERS
(Deputy Director, Division of Biology and Medicine, Atomic Energy Commission, Washington, D. C.)

"Advances Significant to Electronics"

MARCH 24, 10:00–12:00 A.M.

This special program will be addressed by five invited speakers, who will deal with outstanding advances significant to electronics and of real interest to engineers.

1. Cybernetics

The capacity of the individual to assimilate and apply information.

NORBERT WIENER
(Massachusetts Institute of Technology, Cambridge, Mass.)

2. Information Theory

Limitations on the transmission of information imposed by bandwidth, time, and signal-to-noise ratio.

CLAUDE SHANNON
(Bell Telephone Laboratories, Murray Hill, N. J.)

3. Computer Theory

The philosophy of computors as a substitute for the brain in repetitive and original thinking processes.

JOHN VON NEUMANN
(Institute for Advanced Study, Princeton, N. J.)

4. Electronics and the Atom

I. I. RABI
(Columbia University, New York, N. Y.)

5. Pulse Modulation

The broad significance of this form of modulation and its application to time-division multichannel systems.

E. M. DELORAINE
(International Telephone and Telegraph Corporation, New York, N. Y.)

COMMITTEE MEETINGS

March 22–25, 1948

Date	Parlor B	Parlor C	Parlor E	Parlor F	Parlor G	West Ballroom
MONDAY 10:00 A.M.			Electron Tube Subcommittee	Receivers Subcommittee	RMA TR4.1 ProgramTransmitters	
2:15 P.M.	Standards (A. B. Chamberlain, *Chairman*)	Radio Transmitters (E. A. Laport, *Chairman*)		RMA TR9 Component Standardization (J. B. Coleman, *Chairman*)		
8–10:30 P.M.						Sections (A. W. Graf, *Chairman*)
TUESDAY 10:00 A.M.	Railroad & Vehicular Communication (G. M. Brown, *Chairman*)	Navigation Aids (J. A. Pierce, *Chairman*)	Wave Propagation (S. A. Schelkunoff, *Chairman*)	RMA R. F. & I. F. Transformers		
2:15 P.M.	Radio Receivers (W. O. Swinyard, *Chairman*)	Antennas (P. S. Carter, *Chairman*)	Electroacoustics (E. Dietze, *Chairman*)	RMA R4 Television Receivers (I. J. Kaar, *Chairman*)	Education	
WEDNESDAY 10:00 A.M.	Electron Tubes (R. S. Burnap, *Chairman*)	Board of Editors (A. N. Goldsmith, *Chairman*)	Electronic Computers (J. R. Weiner, *Chairman*)	Audio Facilities		
2:15 P.M.	Audio-Video (H. A. Chinn, *Chairman*)	Board of Editors (A. N. Goldsmith, *Chairman*)	Indust. Electronics (G. P. Bosomworth, *Chairman*)	(R. A. Main, *Chairman*)	Membership (B. Dudley, *Chairman*)	
THURSDAY 10:00 A.M.	Television (P. J. Larsen, *Chairman*)	Research (F. E. Terman, *Chairman*)	Symbols (E. W. Schafer, *Chairman*)	Circuits (J. G. Brainerd, *Chairman*)		

TECHNICAL PROGRAM SCHEDULE

	Hotel Commodore Grand Ballroom	Hotel Commodore East Ballroom	West Ballroom	Grand Central Palace Maroon Room	Grand Central Palace Blue Room
Monday, March 22					
Morning	—	—	—	—	—
Afternoon 2:30–5 P.M.	Frequency Modulation	Networks	Systems I	Navigation Aids	Antennas I
Tuesday, March 23					
Morning 10 A.M.–12:30 P.M.	—	Amplifiers	Systems II	Electronics I—Tube Design and Engineering	Antennas II
Afternoon 2:30–5 P.M.	Superregeneration	Transmission	Nuclear Studies	Electronics II—Industrial Application of Tubes and Electronic Circuits	Components and Supersonics
Evening 8–10:30 P.M.	Symposium: Nuclear Science	—	—	—	—
Wednesday, March 24					
Morning 10–12:30 P.M.	Symposium: Advances Significant to Electronics	—	—	—	—
Afternoon 2:30–5 P.M.	Television	Synthetic Crystals	Broadcasting and Recording	Electronics III—Tube Manufacture	Measurements I—V.H.F., U.H.F., and S.H.F.
Thursday, March 25					
Morning 10–12:30 P.M.	Computers I—Systems	—	Propagation	Electronics IV—New Forms of Tubes	Measurements II
Afternoon 2:30–5 P.M.	Computers II—Components	—	Microwaves	Receivers	Active Circuits

SUMMARIES OF TECHNICAL PAPERS

NOTE

No papers are available in preprint or reprint form nor is there any assurance that any of them will be published in the PROCEEDINGS OF THE I.R.E., although it is hoped that many of them will appear in these pages in subsequent issues.

Frequency Modulation

1. F.M. DETECTOR TUBE WITH INSTANTANEOUS LIMITING AND SINGLE-CIRCUIT DISCRIMINATOR

Robert Adler

(Zenith Radio Corporation, Chicago Ill.)

Characteristics resembling a step function, suitable for instantaneous limiting, are obtained in a grid-controlled tube by using electron-optical principles. A practical form of the tube operates simultaneously as a limiter and, with the aid of one tuned circuit, as a discriminator. It provides a good and simple detector for f.m. and television sound.

2. A PROPOSED COMBINED F.M. AND A.M. COMMUNICATION SYSTEM

J. C. O'Brien

(General Railway Signal Company, Rochester, N. Y.)

A system using two simultaneous a.m. and f.m. communication channels with a single carrier halves the usual drift tolerance and guardband width per a.f. channel. Separation circuits developed for receivers, treating discriminators as bridge circuits, include a "dual" of the ratio discriminator, double-triode discriminators, and f.m. and a.m. degenerative i.f. circuits.

3. RATIO OF FREQUENCY SWING TO PHASE SHIFT IN PHASE- AND FREQUENCY-MODULATION SYSTEMS TRANSMITTING SPEECH

D. K. Gannett and W. R. Young

(Bell Telephone Laboratories, Inc., New York, N. Y.)

Data on the subject are derived by computation and simple voice-frequency experiment. The results vary with voices and with circuit conditions. With a carbon microphone used in the mobile radio system, the ratio of peak frequency swing to peak phase shift ranged from 1.1 to 1.5 kilocycles per radian for phase-modulation systems and from 0.6 to 1.2 for f.m. systems.

4. A NEW MAGNETRON FREQUENCY-MODULATION METHOD

P. H. Peters

(General Electric Company, Schenectady, N. Y.)

This paper describes a new method for frequency-modulating and stabilizing magnetrons in the range of 100–1000 Mc. Using a magnetron-type reactance section, the technique results in an 8 per cent carrier-frequency deviation with less than 10 per cent change in power output. A physical interpretation of the space-charge action producing the tuning is presented. Performance of an 850-Mc. 300-watt transmitter with 0.03 per cent center-frequency stability and 50 db signal-to-noise ratio is discussed.

5. I.F. DESIGN FOR F.M. RECEIVERS

K. E. Farr
(Hazeltine Electronics Corporation, Little Neck, N. Y.)

The paper will discuss the desired characteristics of an f.m. i.f. response, illustrating the importance of this response in determining tuning characteristic, drift response, distortion, and production economy. Also, typical design data will be presented for two- and three-stage i.f.-amplifier systems. The effect of production tolerances also will be reviewed.

Networks

6. PROPERTIES OF SOME WIDE-BAND PHASE-SPLITTING NETWORKS

D. G. C. Luck
(Radio Corporation of America, Princeton, N. J.)

Passive networks that yield polyphase output from single-phase input over wide frequency bands are discussed. A simple expression is derived for phase difference between currents in branches of a network as a function of frequency, from which overall operating properties are evident and direct circuit design from required performance is possible. Performance and design curves are presented.

7. THEORY AND DESIGN OF CONSTANT-CURRENT NETWORKS

C. S. Roys and P. H. Chin
(Syracuse University, Syracuse, N. Y.)

The fundamental theory of constant-current networks covering equivalent circuits, current regulation, losses, and efficiency is discussed, as well as design procedures involving either linear or saturating reactors. Predetermined and experimental results for given networks are compared.

8. NEW PARAMETER-ADJUSTMENT METHOD FOR NETWORK TRANSIENTS

M. J. Di Toro
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)
and R. C. Wittenberg
(Ford Instrument Co., Long Island City, N. Y.)

A new method by which parameters in networks such as television video amplifiers may be adjusted to give a "ramp" transient response shape is described. In certain applications it is found that the transient response to an input step function avoids undesirable overshoot and simultaneously gives a small build-up time or wide-band performance.

9. APPLICATION OF TCHEBYSCHEF POLYNOMIALS TO DESIGN OF BANDPASS FILTERS

M. Dishal
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

For multiple-tuned filters an elegant, straightforward, exact design procedure is presented using coefficients of the Tchebyschef polynomials. Characteristics of the optimum circuit-response curves for an n resonant-circuit bandpass filter and the exact design equations for filters using up to three coupled resonant circuits are given. Q distribution is considered.

10. A SIMPLIFIED NEGATIVE-RESISTANCE-TYPE Q MULTIPLIER

H. E. Harris
(Massachusetts Institute of Technology, Cambridge, Mass.)

This single-stage circuit makes use of positive feedback around a stabilized amplifier to produce a negative-resistance characteristic which neutralizes part of the losses in a conventional tuned circuit. The stability of the circuit is analyzed thoroughly from the standpoint of percentage change in effective Q for a given percentage change in the amplifier gain and also for the percentage change in gain tolerable with the complete oscillations. Experimental results are demonstrated showing effective Q's as high as 50,000 with good stability. Finally, the present circuit is compared with usual ones of good type as to stability and simplicity.

Systems I

11. TECHNICAL ASPECTS OF EXPERIMENTAL PUBLIC TELEPHONE SERVICE ON RAILROAD TRAINS

N. Monk and S. B. Wright
(Bell Telephone Laboratories, Inc., New York City, N. Y.)

Telephone service is extended experimentally to certain railroad trains through radio stations of the mobile telephone services of the Bell System. This paper describes component parts of the first public train telephone system, results of radio coverage tests on the routes involved, and devices employed to control two-way transmission.

12. REFLECTED-POWER COMMUNICATION

Harry Stockman
(Watson Laboratories, Cambridge, Mass.)

Point-to-point communication, with the r.f. power generated at the receiver and the transmitter replaced by a modulated reflector, represents a system which possesses new and different characteristics. Radio, light (infrared), or sound waves may be used under approximate conditions of specular reflection. This new communication principle may yield high directivity, automatic pinpointing, independence of atmospheric bending and fading, simple transmitter design without tubes, increased security, simplified means for identification and navigation, etc.

13. STATIC-FREE SYSTEMS OF DETECTION

D. L. Hings
(International Electronic Corporation, Indianapolis, Ind.)

The paper refers to recent developments in reduction of the reproduction of impulse energy in a demodulation system. Consideration of the limitations include the design requirements for reproduction of symmetrical waves from preamplifiers under high-amplitude impulse conditions and their effect on a.v.c. and high-Q circuits. A second detector including sideband-rejection circuits will be analyzed in relation to neutralization of sideband impulse energy.

14. SELECTIVE-SIDEBAND TRANSMISSION AND RECEPTION

D. E. Norgaard
(General Electric Company, Schenectady, N. Y.)

A communication system employing newly developed single-sideband techniques offers simplified apparatus and improved performance. Multiplex transmission and reception of two channels of any desired bandwidth is accomplished with better than 40-db channel separation. Distortion caused by selective fading of conventional transmissions is eliminated, and optional choice of the sideband received allows interference reduction. Binaural broadcasts may be received with conventional receivers as single-channel transmissions or may be resolved by simple twin-channel binaural receivers.

15. STATISTICAL METHODS IN THE DESIGN AND DEVELOPMENT OF ELECTRONIC SYSTEMS

L. S. Schwartz
(Hazeltine Electronics Corporation, Little Neck, N. Y.)

A study is made of the factors affecting tolerance assignment in the production and operational stages of an electronic system. The procedure adopted is first to review some of the fundamentals of the statistical control of quality and the assignment of valid, economic production tolerances, and second, to describe how the principles may be applied in the setting of some operational tolerances for an electronic system. The advantages to design and development derived from a knowledge of how tolerances, both productional and operational, are assigned and how they combine statistically are discussed.

16. BASIC PRINCIPLES OF DOPPLER RADAR

E. J. Barlow
(Sperry Gyroscope Company, Great Neck, N. Y.)

A discussion of the basic principles and the techniques of doppler radar is given, beginning with a simple doppler radar system and progressing to more complex systems which furnish more target information. Subjects discussed include the doppler effect, range measurements with c.w. systems, effect of a moving base for the radar system, and filter design required in the radar receiver. Two major applications of these principles are discussed: the detection of moving targets in the presence of much larger fixed targets, and the accurate determination of the velocity of projectiles, such as shells or rockets. Some other applications of these principles are pointed out.

Navigation Aids

17. THE RADIOVISOR LANDING SYSTEM FOR AIRCRAFT

D. G. Shearer
(Culver City, Calif.)
and W. W. Brockway
(Los Angeles, Calif.)

The radiovisor is a unique fundamental approach to a solution of the problem of landing an aircraft under adverse visibility due to weather conditions. A realistic virtual image of a landing area is presented to the pilot in such a manner that he actually sees the landing area as it would appear to him if normal vision were possible. A demonstration of the optical principles and the "type of presentation" will be available. Operational details of the complete "Radiovisor" system are discussed.

18. CONSIDERATIONS IN THE DESIGN OF A UNIVERSAL BEACON SYSTEM

L. B. Hallman, Jr.
(Communication and Navigation Laboratory, Wright Field, Dayton, Ohio)

Airborne beacons are an essential element in aircraft navigation and traffic-control systems since they provide a means of (a) radar range extension and (b) automatic intelligence transmission. However, it is important that a universal beacon be provided which will operate with all ground and airborne radar equipments regardless of the operating frequency of the primary radar equipment. Also, the airborne beacon must provide the maximum of facilities for intelligence transmission. The paper outlines the specifications for a proposed universal beacon system satisfying the above basic requirements and discusses certain design criteria for the several components of the proposed system.

19. SURVEILLANCE RADAR DEFICIENCIES AND HOW THEY CAN BE OVERCOME

J. W. Leas
(Air Transport Association of America, Washington, D. C.)

Ground surveillance radar is being installed at certain airports today and many more installations are planned. Every effort will be made to use the radars to the fullest extent in increasing safety and expedition in the control of air ftraffic. But certain major technical and operational deficiencies must be overcome before ground radar can be used as a primary traffic-control aid in civil operations. These limitations will be explained, and ways in which they can be overcome will be explored.

20. THE COURSE-LINE COMPUTER

F. J. Gross
(Civil Aeronautics Authority, Department of Commerce, Indianapolis, Ind.)

In an aircraft navigation radio receiver, the phase relation between 30-cycle voltages is proportional to the bearing from an omnirange station, and in the aircraft radar distance-measuring equipment a d.c. voltage is proportional to the distance to the same station. The course-line computer converts these quantities into course-deviation meter deflections (left-right meter) for any selected straight-line track passing within the service range of the transmitting equipment. The computer also indicates continuously the distance between the aircraft and any desired destination. The pilot selects a track and a destination by adjusting three dials on the computer. Results of extensive flight tests of a working model of the computer will be presented.

21. AIRCRAFT INSTRUMENTATION AND CONTROL

F. L. Moseley, J. A. Biggs, E. T. Heald, and J. C. McElroy
(Collins Radio Company, Cedar Rapids, Iowa)

Aircraft navigation systems and related instruments are now developing in the United States in a way which makes it possible to provide comprehensive aids to track flying, schedule maintenance, and traffic control, either in visual-manual or full automatic form. The United States policy decision to standardize the Omnidirectional Radio Range–Distance Measuring System provides a polar-diagram foundation of continuous position fixing upon which any desired system of tracks can be built. Through the addition of geometric and time-rate-distance computing systems, tracks, destinations, holding patterns, and required schedule speeds can be derived from the polar information supplied by the omnirange and distance-measuring facilities. Systems and apparatus recently developed to implement such a program are described in this paper.

Antennas I

22. PHYSICAL LIMITATIONS OF DIRECTIVE RADIATING SYSTEMS

L. J. Chu
(Massachusetts Institute of Technology, Cambridge, Mass.)

This paper deals with the relationship between the radiation gain and the optimum impedance bandwidth of an electromagnetic radiating system. For an arbitrary radiating system of given over-all dimensions, it is shown that a gain higher than the conventional value can be obtained only by sacrificing the optimum impedance bandwidth. The optimum impedance bandwidth is improved by the additional dissipation in the radiation system, with a reduction of the radiating efficiency of the system.

23. THE RADIATION RESISTANCE OF AN ANTENNA IN AN INFINITE ARRAY OR WAVEGUIDE

H. A. Wheeler
(Consulting Radio Physicist, Great Neck, N. Y.)

The electromagnetic field in front of an infinite flat array of antennas can be subdivided into wave channels, each including one of the antennas. Each channel behaves like a hypothetical waveguide similar to a transmission line made of two conductors in the form of parallel strips. A simple derivation then leads to the radiation resistance of each antenna and to some limitations on the antenna spacing. In the usual flat array of half-wave dipoles, each allotted a half-wave square area, and backed by a plane reflector at a quarter-wave distance, the radiation resistance of each dipole is $480/\pi = 153$ ohms. In a finite array, this derivation is a fair approximation for all antennas except those too close to the edge. This derivation also verifies the known formula for the directive gain of a large flat array in terms of its area. The same viewpoint leads to the radiation resistance of an antenna in a rectangular waveguide, which has previously been derived by more complicated methods.

24. REFLECTORS FOR WIDE-ANGLE SCANNING AT MICROWAVE FREQUENCIES

R. C. Spencer, Wade Ellis, and Ellen C. Fine
(Watson Laboratories, Cambridge, Mass.)

The analysis of spherical reflectors is simplified. Two improvements over that of the sphere are indicated when the scan is in one plane. The optimum reflector is a barrel, with axis perpendicular to the plane of scan and to the axis of the paraboloid; and with sections which are circles if the scan is symmetrical, and which are spirals of the form of $\rho = \rho_0 e^{k\psi}$ if the scan is asymmetrical. A value of $k = \frac{1}{4}$ is suggested for off-axis feeding.

25. MEASURED IMPEDANCE OF VERTICAL ANTENNAS OVER FINITE GROUND PLANES

A. S. Meier and W. P. Summers
(Ohio State University, Columbus, Ohio)

An investigation was made to obtain some fundamental information concerning the relation of the impedance of a vertical antenna over a finite ground plane as a function of the size and shape of the ground plane when dimensions are relatively small in terms of wavelength. It was found that the input impedance is a damped oscillating function of wavelength and ground-plane dimensions, the impedance of a circular ground plane varying from ± 5 to ± 20 per cent. Similar variations were observed on a square ground plane which were approximately 50 per cent of those of the circular ground plane except when the dimensions of the ground plane were small. In general, it was found that the impedance is quite critical with respect to the size and shape of the ground plane and relatively independent of the thickness of the antenna. Measurements were made at microwave frequencies by a modified Chipman method capable of determining small differences in antenna impedance.

26. CURRENT DISTRIBUTIONS ON AIRCRAFT STRUCTURES

J. V. N. Granger
(Harvard University, Cambridge, Mass.)

This paper treats an experimental technique which can be employed for obtaining r.f. current distributions on aircraft structures excited by transmitting antennas. After a brief discussion of the basic method, practical measuring devices are described. Current distributions and corresponding radiation patterns for the v.h.f. stub on a P-47 aircraft, a h.f. inclined-wire antenna on a bomber, and a h.f. tail-cap antenna on a DC-6 are presented as examples. Means of shaping current distributions to obtain desired radiation patterns are briefly discussed.

Amplifiers

27. LOW-NOISE AMPLIFIER

Henry Wallman, A. B. Macnee, and C. P. Gadsden
(Massachusetts Institute of Technology, Cambridge, Mass.)

This paper describes an amplifier circuit that yields very low noise factors, consisting of a grounded-cathode triode followed by a grounded-grid triode. The combination is entirely noncritical and provides the low noise factor of a triode with the high amplification and stability of a pentode. Noise factors averaging 0.25 db at a carrier frequency of 6 Mc. and 1.35 db at 30 Mc. have been achieved. Typical circuit details are given. With intermediate-frequency amplifiers employing this circuit as the first stage, it was possible to build 3000-Mc. (radar) receivers with over-all (radio frequency) noise factors of 8.7 db.

28. PHASE DISTORTION IN AUDIO SYSTEMS

L. A. deRosa
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

The effects of amplitude and frequency distortions on audio reproduction are well known. The importance of phase distortion, however, has not been fully recognized, except in several cases involving transient reproduction. It is shown that the ear acts as a transient analyzer for brief instants on the occasion of a change in envelope of the signal and before it acts as a harmonic analyzer. This transient-analysis function is performed for envelope changes of the order of 6–12 cycles depending on the modulated frequency. Phase distortion changes the intervals during which the ear is performing the sequential operations of integration and frequency analysis, and thus deteriorates the reproduction.

29. VISUAL ANALYSIS OF AUDIO-FREQUENCY TRANSIENT PHENOMENA

D. E. Maxwell
(Columbia Broadcasting System, Inc., New York, N. Y.)

There is presented an audio-frequency measuring technique based upon the transient application of a sine-wave voltage to the input of the system or device under measurement. A great advantage of this technique over other methods of transient analysis is the ease with which the results may be analyzed in terms of sine-wave performance. An essential part of the required measuring equipment is an electronic switch and synchronizer unit, which provides the switching, phasing, and synchronizing functions necessary for visual presentation of the transient phenomena on a cathode-ray oscillograph. Its theory of operation is described in detail. Several applications of the technique to typical problems are included.

30. SQUARE-WAVE ANALYSIS OF COMPENSATED AMPLIFIERS

P. M. Seal
(University of Maine, Orono, Maine)

The results of a complete analysis of a single-stage video-frequency amplifier are presented. Output wave shapes for a number of cases where the input voltage is a symmetrical square wave are drawn, both for high-frequency compensation and for low-frequency compensation. The corresponding frequency- and phase-response curves are drawn for comparison. The effect of the cathode impedance on the low-frequency-compensated case is considered briefly.

31. A NEW FIGURE OF MERIT FOR THE TRANSIENT RESPONSE OF VIDEO AMPLIFIERS

(R. C. Palmer and Leonard Mautner
(Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

The design of wideband amplifiers on a transient basis is becoming increasingly important, particularly with regard to television, radar, and applied pulse techniques. This paper describes a new method of evaluating the transient response of wideband amplifiers, using as examples a wide variety of types of both two- and four-terminal interstage networks, and considers both the rise time as well as the transient overshoot, arriving at a new figure of merit relating these networks. By the use of this method it is possible to compare the transient response of different networks even though their rise times and overshoots may be significantly different.

32. DISTRIBUTED AMPLIFICATION

E. L. Ginzton
(Stanford University, Stanford, Calif.)
W. R. Hewlett
(Hewlett-Packard Company, Palo Alto, Calif.)
J. H. Jasberg and J. D. Noe
(Stanford University, Stanford, Calif.)

A new principle in wide-band amplifier design is presented. It is shown that, by an appropriate distribution of ordinary vacuum tubes along artificial transmission lines, it is possible to obtain amplification over much greater bandwidths than would be possible with ordinary circuits. The ordinary concept of "maximum bandwidth versus gain product" does not apply to this *distributed amplifier*. The high-frequency limit of the distributed amplifier appears to be determined by the grid-loading effects. The general design considerations included are the effect of improper termination of transmission lines; methods for controlling the frequency-response and phase characteristics; the design which provides the required gain with fewest possible number of tubes; and a discussion of high-frequency limitations. The noise factor of the amplifier is evaluated.

Systems II

33. THEORETICAL STUDY OF PULSE-POSITION MODULATION WITHOUT FIXED REFERENCE

A. E. Ross
(Stromberg-Carlson Company, Rochester, N. Y.)

This paper discusses pulse-position modulation without fixed reference, a type of pulse communication in which the information is contained in the variation of the distance between the successive pulses. Because of the extreme nonlinearity of this type of modulation, its mathematical theory cannot be constructed in the usual manner; that is, with the harmonic analysis of the modulated pulse train as the starting point. The essential features and the type of regularity characteristic of this apparently random method of sampling the signal begin to appear when one realizes that the positions at which the signal is sampled are iterations of a one-to-one continuous transformation (induced by the method of sampling) of the interval 0 to 2π (the period of the signal). Making use of the properties of such transformations, one discovers that sampling positions depend only upon the ratio f_s/f_p of the signal frequency f_s to the pulse-repetition rate f_p and upon the amplitude of the signal. It is found that signals of certain frequencies f_s are sampled k times in x periods at (stable) positions depending on f_s, and that for fixed k and x there exist whole intervals (intervals of stability) of such values of f_s. Curves are included showing how the intervals of stability may be determined in typical cases and how fixed sampling points may be found for given value of f_s/f_p.

34. HIGH-QUALITY RADIO PROGRAM LINKS

M. Silver and H. A. French
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

The advent of high-fidelity broadcasting, especially f.m. and television, has stressed the need for proper aural program-supply circuits. Equipment at the medium and very-high frequencies has previously been used for this purpose. Improved results are obtained by operation in the u.h.f. band of 940–952 Mc. recently allocated by the F.C.C. for this purpose. A radio link (STL) equipment designed to meet the requirements of this service is described. System and component design considerations are discussed. Technical features of the equipment are outlined and performance characteristics given.

35. SIGNAL-TO-NOISE RATIO IMPROVEMENT IN A P.C.M. SYSTEM

A. G. Clavier, P. F. Panter, and W. Dite

(Federal Telecommunications Laboratories, Inc., Nutley, N. J.)

Pulse-count modulation (p.c.m.) makes use of three successive operations: sampling, quantizing of the sampled amplitudes, and coding of the quantized amplitudes. The present paper discusses the effect of fluctuating noise on the type of signals used in p.c.m. Assuming that the distribution of noise bursts obeys Poisson's law, the noise power affecting the decoded signals can be computed for a given signal-to-noise ratio in the code pulses. It is shown that for a binary code the signal-to-noise ratio in the decoded signals expressed in decibels is approximately equal to the signal-to-noise-power ratio of the code pulses. This relation is substantially independent of the number of digits in the binary code provided this number is sufficiently large; more precisely, larger than 3 or 4.

The distortion due to quantization varies considerably, however, with the number of discrete levels used, and consequently with the number of code digits. Sometimes a distortion power larger than noise power can be tolerated. When very high fidelity is considered, the number of digits should be increased to the point where the noise power in the decoded signals is equal to the distortion power. A curve is given showing the necessary number of digits for a given signal-to-noise ratio in the code pulses. It is of course unnecessary to increase the number of digits beyond that value, as any further reduction in distortion would be masked by the presence of a higher noise level. For instance, if a 60-db signal-to-noise ratio is required in the decoded signals, the maximum number of digits is found to be of the order of 11.

In case the communication system includes a number of relays, regenerative repeaters can be used; that is to say, repeaters in which the code pulses are reshaped and sent on practically undistorted to the next repeating point. A very small increase in the signal-to-noise ratio affecting the code signals is sufficient to overcome the cumulative effect of noise in the whole chain of repeaters. This is one of the most important properties of p.c.m. systems, both for radio or cable applications.

36. RADIO-WIRE LINKS FOR MULTICHANNEL TRANSMISSION

E. M. Ostlund and H. R. Hunkins

(Federal Telecommunications Laboratories, Inc., Nutley, N. J.)

The application of radio circuits in conjunction with wire lines for subcarrier multiplex telephone and telegraph transmission is increasing. This paper describes f.m. radio links and the line-carrier terminal equipment with which it is designed for use. The radio link is intended for operation over relatively short circuits involving a small number of, or no, relay points. Line-carrier terminal equipment designed to provide economical high-quality telephone and telegraph transmission over short-haul circuits is described. Application of radio links to operation in connection with long-haul circuits is discussed.

37. BANDWIDTH REDUCTION IN COMMUNICATION SYSTEMS

W. G. Tuller

(Melpar, Inc., Alexandria, Va.)

There are two possible methods of bandwidth reduction in communication systems. One of these, which has been thought about for years but worked on relatively little, takes advantage of the coherence of the message; i.e., the fact that from a knowledge of the past behavior of the message it is possible to predict its future behavior, in general, with a fair degree of accuracy.

The second method of bandwidth reduction has only recently been given public attention. In this method bandwidth is exchanged for signal-to-noise ratio in accordance with the relation.

$$H = BT \log (1 + S/N)$$

where H is quantity of information; B is the bandwidth; T is the time of transmission; and S/N is the signal-to-noise ratio in the transmission link.

The first method is discussed in some detail, showing how one may determine mathematically when all the bandwidth compression permitted by this method has been obtained. It is shown that the maximum bandwidth compression obtainable in this manner occurs when the signal has the statistical properties of random noise. Examples of past approximations to this technique have been the use of pre-emphasis and de-emphasis in recording and radio broadcast systems and the use of derivative control in servomechanisms. Further possibilities for the use of these techniques are pointed out.

Two possible methods of using the second or trading method of bandwidth reduction are discussed. The first of these employs techniques analogous to those of p.c.m.; however, in this case the coding is carried out in the inverse direction from that usually employed, so that bandwidth is gained at the expense of a loss in tolerable signal-to-noise ratio.

The second method for using this trading principle makes use of the fact that the output of a filter may be predicted if its input wave form and transient response are known. A possible system using this fact is outlined. It is shown that this system has the same limitations as does the "inverted" p.c.m. system, as would be predicted from the general theory. It is pointed out that these are typical rather than optimum practical systems, and the need for future engineering work along these lines is emphasized.

Electronics I

Tube Design and Engineering

38. THERMIONIC EMISSION FROM GRIDS IN VACUUM TUBES

M. Arditi and V. J. DeSantis

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

Primary emission of electrons from the control grids is often a serious limitation in the design of modern high-power u.h.f. tubes. Electron bombardment of the grids may produce partial or total dissociation of coatings on the surface of the grid, even though the temperature of the grid is well below the temperature corresponding to thermal dissociation in vacuo. This phenomenon may be harmful to cathode emission or grid-emission inhibitors. Its effect upon design dimension scaling and upon methods of testing materials for inhibiting grid emission is also discussed. Some experimental results on grid-emission inhibitors for tubes with oxide-coated and thoriated-tungsten filaments are reported.

39. THE NEGATIVE-ION BLEMISH IN A CATHODE-RAY TUBE AND ITS ELIMINATION

R. M. Bowie

(Sylvania Electric Products Inc., Flushing, N. Y.)

A critical review of the widely scattered and somewhat conflicting data regarding negative ions in cathode-ray tubes and blemish formation. By mass-spectrographic means, these negative ions have been studied by several observers with results which fall into agreement only after careful study of the experimental conditions. The use of a backing layer such as aluminum reduces the blemish but does not eliminate it, apparently owing to porosity of the backing layer. The ion trap removes the negative ions from the electron beam by electron-optical means, thus eliminating the blemish. A trap is characterized by three essentials: (1) A beam must be formed before reaching the trap. (2) A magnetic field having a component perpendicular to the direction of propagation of the beam must be provided. (3) A spot must be provided on which the ions may impinge while permitting the electrons to pass by.

40. WIDE-TUNING-RANGE CONTINUOUS-WAVE HIGH-POWER MAGNETRONS

P. W. Crapuchettes

(Litton Industries, San Carlos, Calif.)

The design and development of a typical 1-kw. magnetron capable of tuning ± 20 per cent at "S"-band frequency are outlined. Problems of design and construction are illustrated and techniques used in their solution are described. Variations in design (strapping, cavities, anode length, and loading) for other power levels and wavelengths are discussed. Comparison of magnetron performance with triodes by circuit analysis is indicated, the assumption being that the electrons contribute reactance to the circuit.

41. WIDE-RANGE TUNING SYSTEMS FOR MAGNETRONS

E. N. Kather

(Raytheon Manufacturing Company, Waltham, Mass.)

Various methods of obtaining wide-range tuning of magnetrons are described, including (1) external-reactance systems; (2) variable-capacitance systems such as the strap-capacitance or "cookie-cutter" tuner, the vane-capacitance tuner and the V-capacitance tuner; (3) variable-inductance systems such as the "crown of thorns" tuner; and (4) combination variable-vane capacitance and inductance systems. The problem of interfering resonances and magnetic-field distortion introduced by the tuning system are described and explained theoretically. Practical solutions of these and other mechanical problems are presented. Performance data are included to illustrate actual operation of magnetrons using the various tuning systems.

42. DESIGN CHARACTERISTICS OF HEARING-AID TUBES

G. W. Baker

(Chatham Electronics Corporation, Newark, N. J.)

Hearing-aid tubes are required to operate in a resistance-coupled-amplifier circuit over a very wide range of plate, screen, and filament supply voltages with fixed values of the circuit constants. Tube design characteristics that insure optimum operation over this wide range of supply voltages are described.

Antennas II

43. AN OMNIDIRECTIONAL HIGH-GAIN ANTENNA FOR CIRCULARLY POLARIZED RADIATION

A. G. Kandoian

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

It is advantageous, in a number of communication applications, to use circularly polarized radiation in the place of the more commonly used linear polarization. It has recently been indicated that better v.h.f. broadcasting coverage can be obtained by the use of circular polarization at the transmitting end. The present antenna development was carried on primarily for this and similar applications. Although there are several known forms of circularly polarized radiating elements, the problem of stacking them to obtain concentration of the vertical radiation pattern, and hence antenna power gain, has proved difficult to solve. The reason for this is the presence of the metallic antenna-supporting structure which distorts the phase of the vertical component of the radiation so that the desired quadrature relationship between the horizontal and vertical field no longer holds. The present design overcomes this difficulty by providing a simple control for both the amplitude and phase of the vertical component of the radiation without undue complication of the antenna feed system. Measured data on an array consisting of a stack of eight elements, as well as a discussion of the experimental technique used in these measurements, will be presented.

44. ANALYSIS OF EFFECT OF CIRCULATING CURRENTS ON THE RADIATION EFFICIENCY OF BROADCAST DIRECTIVE ANTENNA DESIGNS

G. D. Gillett

(Glenn D. Gillett and Associates, Washington, D. C.)

It apparently has not been generally recognized that, for any directive array which focuses most of the field from the array over a relatively narrow horizontal angle, there may be induced therein circulating currents of such magnitude as to multiply the parasitic losses by most surprising amounts and so reduce the radiation efficiency of the array to values that are neither economic nor sufficient to meet the F.C.C. minimum requirements. The purpose of this paper is to point out that there is produced as an inherent part of any such directive array criteria which accurately determine the effective magnitude of these circulating currents and from which their effect on the antenna efficiency can be accurately computed; that they are an inherent part of the design, and that they may be computed without any assumption or reference to the mutual impedance existing between any of the elements of the array. Basically, the magnitude of these circulating currents is determined by the ratio of the r.m.s. value of the unit pattern which results from the array to the r.m.s. sum of the unit vectors used in computing the pattern.

45. A MODEL STUDY OF RERADIATION FROM BROADCAST TOWERS

Andrew Alford and Henry Jasik

(Andrew Alford Laboratory, Boston, Mass.)

Towers of one broadcast station are sometimes present in the field of the antenna of another station. Reradiation from the towers in which currents are induced may have an undesirable effect on the null of the inducing directional antenna. A model study of currents induced in towers of various heights shows that, in general, the induced-current distributions are not even approximately sinusoidal. The induced-current distributions as well as the amplitude of the induced currents are materially changed by varying the impedance connected between the base of the tower and ground. Certain values of reactance produce a marked reduction in reradiated field from towers less than 0.6λ high at the frequency of the inducing field. Experimental evidence indicates that towers $\frac{3}{4}\lambda$ high or higher cannot be effectively detuned by connecting an impedance between the base of the tower and ground.

46. HELICAL BEAM ANTENNAS FOR WIDE-BAND APPLICATIONS

J. D. Kraus

(Ohio State University, Columbus, Ohio)

A helix is a fundamental form of antenna with many radiation modes. Loops and linear conductors can be regarded as special cases of the helix, since a helix of fixed diameter collapses to a loop as the spacing between turns approaches zero, and, on the other hand, a helix of fixed spacing straightens into a linear conductor as the diameter approaches zero. A circularly polarized mode, called the axial or beam mode, has maximum radiation in the direction of the helix axis. The conditions for this and other radiation modes are considered. Optimum dimensions for wide-band applications of the beam mode are discussed and design data given.

47. A CIRCULAR-POLARIZATION ANTENNA FOR F.M.

C. E. Smith

(United Broadcasting Company, Cleveland, Ohio)

and R. A. Fouty

(Ohio State University Research Foundation, Columbus, Ohio)

The use and advantages of circular polarization for f.m. will be discussed and a new antenna for circular polarization which has been installed by the United Broadcasting Company in Cleveland will be presented. Basically, the antenna consists of an array of vertical dipoles and longitudinal slots, fed to produce a uniform pattern in the horizontal plane with high directivity for the vertical field pattern. Development of the antenna and the specific data will be presented with slides. A demonstration with a one-eighth scale model will be presented, showing the uniformity of pattern in the horizontal plane and the circularity of polarization.

Superregeneration

48. SUPERREGENERATION AS IT EMERGES FROM WORLD WAR II

H. A. Wheeler

(Consulting Radio Physicist, Great Neck, N. Y.)

Superregeneration, previously not competitive, found new life shortly before the war, in the higher-frequency ranges, in the "walkie-talkie" and in the "transponder" of IFF. The first application of superregeneration to commercial broadcast receivers has just appeared. The ultimate limitations of a superregenerator are discussed by regarding it as a receiver having a radio-frequency amplifier modulated by pulses at the quench frequency.

49. THEORY OF THE SUPER-REGENERATIVE RECEIVER

W. E. BRADLEY

(Philco Corporation, Philadelphia, Pa.)

An analysis of operation of a superregenerative receiver is obtained in terms of "time-aperture function," which specifies the sensitivity to a short impulse of incoming signal. The Fourier transform of the time-aperture function is the frequency response of the receiver to a continuous carrier. Selectivity curves and time-aperture functions for various quench wave forms are given.

50. SUPERREGENERATION—AN ANALYSIS OF THE LINEAR MODE

H. A. GLUCKSMAN

(Watson Laboratories, Cambridge, Mass.)

A superregenerator operated in the linear mode is regarded as a tuned circuit with periodically varying decrement. Under certain assumptions, a solution is obtained. Sensitivity and selectivity as well as envelope form are discussed. It is shown that well-known properties of superregenerators, such as multiple resonance, are predicted. Test results are given.

51. EXTERNAL AND INTERNAL CHARACTERISTICS OF A SEPARATELY QUENCHED SUPERREGENERATIVE CIRCUIT

SZE-HOU CHANG

(Watson Laboratories, Cambridge, Mass.)

External characteristics of a superregenerative circuit tell how signal parameters affect performance, and internal characteristics give the effect of circuit parameters. External characteristics identify operational modes controlled by internal characteristics. Contour-diagram presentation for the internal characteristics separates the regions of linear and logarithmic mode by a line which is the locus of maximum output.

52. THE HAZELTINE FREMODYNE CIRCUIT

B. D. LOUGHLIN

(Hazeltine Electronics Corporation, Little Neck, N. Y.)

The Hazeltine FreModyne circuit, which can be used to give an inexpensive, but practical, f.m. receiver, contains one dual-triode tube operating as superheterodyne frequency converter, a superregenerative i.f. amplifier, and a side-tuned f.m. detector. The circuit and operational details of the FreModyne f.m. receiver are described. Typical performance characteristics are presented, together with a brief discussion of the circuit components which determine these characteristics.

Transmission

53. SIMPLIFIED PROCEDURE FOR COMPUTING THE BEHAVIOR OF MULTICONDUCTOR LOSSLESS TRANSMISSION LINES

S. FRANKEL

(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

A method is given for calculating the behavior of an arbitrary, uniform, lossless, multiconductor transmission line operating in the *TEM* mode. It is shown that the performance of the line depends essentially on the Maxwell's coefficients of capacitance for the line. The method is used to obtain characteristic impedance of lines, input admittances of unbalanced lines above ground, and a formula for "loop-to-loop" coupling between two wire lines. The method is based on the fact that a simple relation exists between the current and charge in the forward (or back) wave in any line.

54. OPTIMUM GEOMETRY FOR RIDGED WAVEGUIDE

W. E. WALLER, S. HOPFER, AND M. SUCHER

(Polytechnic Research and Development Co., Brooklyn, N. Y.)

The work of other authors on ridged waveguide is extended to yield more information on the bandwidth and other properties. Methods for calculating the attenuation and power-handling capacity are described, and curves relating the fundamental cutoff frequency, bandwidth, attenuation, and power-handling capacity to the parameters of the geometry are presented. Comparisons of these quantities with those for rectangular waveguide and coaxial line are made, and the region of usefulness of ridged waveguide is discussed on this basis.

55. FIELDS IN NONMETALLIC WAVEGUIDES

ROBERT N. WHITMER

(Rensselaer Polytechnic Institute, Troy, N. Y.)

The transmission of electromagnetic waves in dielectric rods has been treated by considering the case of a flat slab of dielectric with the electric field parallel to the faces. The type of field both inside and outside the dielectric is described both for uniform dielectric constant and one increasing toward the center. The calculations are extended to rods of circular cross section for the H_{on} type of mode. The propagation factors are discussed. The dielectric-waveguide wavelengths lie between the free-space wavelength and the length of a plane wave in the dielectric.

56. A WIDE-BAND WAVE-GUIDE-FILTER STRUCTURE

S. B. COHN

(Harvard University, Cambridge, Mass.)

This paper presents the theoretical analysis, design procedure, and experimental verification of a waveguide-filter structure. This structure is useful when a wide pass band is desired. The lower cutoff frequency of the pass band is the natural cutoff frequency of the waveguide itself. The upper cutoff is due to a succession of cavities and constrictions in the waveguide. Although the individual filter sections have additional pass bands at higher frequencies, it is possible by proper design of a multisection filter to eliminate all spurious responses up to several times the upper cutoff frequency of the principal pass band.

57. THE TRANSMISSION-LINE VECTOR DIAGRAM

W. C. BALLARD, JR.

(Cornell University, Ithaca, N. Y.)

The paper describes a graphical method for the solution of conventional transmission-line problems which requires no transmission-line charts and from which voltages and currents may be directly scaled and angles measured as in normal vector diagrams. Simple illustrations to the determination of standing-wave ratio of a line terminated in other than the characteristic impedance and to the input impedance of short-circuited lines are given. As a typical application of the system, it is shown how the length and position of the proper short-circuited stub for impedance matching may be obtained from a few voltmeter measurements along the uncompensated line.

Nuclear Studies

58. OSCILLATOR DESIGN FOR 130-INCH FREQUENCY-MODULATED CYCLOTRON

E. M. WILLIAMS AND H. E. DEBOLT

(Carnegie Institute of Technology, Pittsburgh, Pa.)

Electrical design problems in a wide-deviation frequency-modulated oscillator for the Carnegie Institute of Technology's 130-inch synchro-cyclotron are discussed. Some unique problems are involved because frequency requirements are higher than in other machines now in use or under construction. The load, primarily reactive, is about 500,000 kilovoltamperes; about 80 kilowatts are required for losses in the dee and supporting and tuning structures. Results of tests on a full-scale model are described.

59. AN ELECTRONIC INSTRUMENT FOR THE DETERMINATION OF THE DEADTIME AND RECOVERY CHARACTERISTICS OF GEIGER COUNTERS

L. COSTRELL

(National Bureau of Standards, Washington, D. C.)

An electronic instrument developed for the measurement of the deadtime and recovery characteristics of Geiger counters will be described. Measurement of the characteristics to an accuracy of 2 microseconds is possible. The instrument records the number of pulses that follow other pulses within a predetermined time interval.

60. ELECTRONIC CLASSIFYING, CATALOGUING, AND COUNTING DEVICES

J. H. Parsons
(Clinton Laboratories, Oak Ridge, Tenn.)

There are many physical phenomena that can be studied only by classifying each of the several parts according to the magnitude and determination of the number of parts in each increment of magnitude. Electronic systems have been developed that carry out the operations of classifying, cataloguing, and counting each of the parts, and perform these operations much faster than can be done manually. The systematic operations of these electronic devices will be discussed. These devices have many possible fields of application in the scientific and industrial world.

61. HEALTH PHYSICS PROBLEMS IN ATOMIC ENERGY

K. Z. Morgan
(Clinton Laboratories, Oak Ridge, Tenn.)

Instruments for personnel protection from radiation exposure will be described. Many new developments have been made, and pictures and brief descriptions of instruments will be given.

62. A SELECTIVE DETECTOR FOR HEAVY CHARGED PARTICLES

Keith Boyer
(Massachusetts Institute of Technology, Cambridge, Mass.)

Electronic circuits are used to count the number of protons or other heavy charged particles of a prescribed energy in the presence of very high backgrounds by placing requirements on the range and specific ionization of the particles. The particles to be detected pass through an aluminum absorber, two proportional counters, and stop in a third, giving rise to a current pulse, proportional in amplitude to the specific ionization of the particle, in each counter respectively. Pulses from each of the three counters are amplified and passed through a series of amplitude gates and coincidence circuits arranged so that a particle is recorded when three pulses of the correct amplitude occur in time coincidence. Thus the kind of particle selected is determined by adjusting the amplitude gates, and the energy by the amount of absorber the particle must traverse.

Electronics II

Industrial Applications of Tubes and Electronic Circuits

63. EXPERIMENTAL STUDY OF THE EFFECTS OF TRANSIT TIME IN CLASS-C POWER AMPLIFIERS

Oliver Whitby
(Harvard University, Cambridge, Mass.)

A description is given of the apparatus developed for measuring the transit angle of the plate and grid current pulses in a class-C triode power amplifier. The equipment also includes provision for observing the actual shape of the current pulses and for studying the interrelation between transit time, pulse shape, operating voltages, and plate load impedance. With the apparatus it is also possible to investigate these relations in a tetrode.

64. NEW RECEIVING TUBES FOR INDUSTRIAL USE

C. M. Morris and H. J. Prager
(Radio Corporation of America, Harrison, N. J.)

There is an increasing need for tubes of the receiving-tube type for applications outside the field of home entertainment, the most important being in industrial control and measurement. Some of the chief requirements over available radio types are stability of characteristics during operation, uniformity among tubes of one type, long life (10,000 hours), and mechanical sturdiness. Three tube types designed specifically for this service are described.

65. USE OF DIODE RECTIFIERS WITH ADJUSTABLE TRANSFORMERS FOR MOTOR SPEED CONTROL

W. N. Tuttle
(General Radio Company, Cambridge, Mass.)

Use of a full-wave diode rectifier with an adjustable transformer and choke for armature-voltage control of the speed of a d.c. shunt motor gives more than 50 per cent increase in the continuous-duty low-speed torque rating in comparison with operation on the usual type of thyratron rectifier. Use of a choke is feasible with diodes because the inductance required for commutation in the rectifier circuit is very much less than when thyratrons are employed and the output voltage is controlled by delaying the firing point of the tubes. Constructional details are given of a control for a ½-hp. motor providing reversal, dynamic braking, and speed control over a range of between 10:1 and 20:1. All equipment except the motor is contained in a case 12×9×4½ inches, small enough to be placed beside a machine for direct control by the operator.

66. SERVO-SYSTEM PERFORMANCE MEASUREMENT

C. F. White
(Naval Research Laboratory, Washington, D. C.)

Great advances in the design of servo systems have resulted from their analytical treatment in terms of circuit response to sinusoidally varying signals. A servo-system performance analyzer based on this principle is described. It includes among other features a signal generator and means for measuring both the amplitude and phase of the system response.

67. SPARK OSCILLATORS FOR ELECTRIC WELDING OF GLASS

J. P. Hocker
(Corning Glass Works, Corning, N. Y.)

In electric welding of glass, the objects are heated by current passed through the glass from contact electrodes or small flame brushes. Considerations of safety, economy, dependability, and performance have led to wide use of spark-oscillator power supplies. In designing these oscillators for specific applications, the chief conditions to be satisfied are: (1) The circuit should match the impedance of the load as well as possible, while this load decreases by several orders of magnitude during the operation. (2) At no time should the terminal voltage be large enough to spark over the surface, the limit itself being a function of glass temperature. A design procedure is described to attain these conditions.

Components and Supersonics

68. PHASE-CORRECTED DELAY LINES

M. J. Di Toro
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

The complexity and physical volume of electrical delay lines increases with the product of over-all bandwidth and delay time. In usual designs the over-all bandwidth is limited by phase, rather than amplitude, restrictions. A novel scheme of phase corrections is described for wired lines, which leads to a phase bandwidth greater than the amplitude bandwidth. The new phase-corrected line has a geometry amenable to continuous automatic fabrication. Statistical quality-control principles give promise of substantial reduction of rejects due to random manufacturing variations. Indicated applications are the improvement of multivibrator stability, reduction of television ghosts, and pulse-decoding systems.

69. ON THE THEORY OF THE DELAY-LINE-COUPLED TRAVELING-WAVE AMPLIFIER

H. G. Rudenberg
(Harvard University, Cambridge, Mass.)

The delay-line-coupled amplifier of Percival has a bandwidth, obtainable with a chain of standard electron tubes, which is one order of magnitude larger than that limiting a single-tube stage, although requiring a correspondingly larger number of tubes. Delay distortion and the cutoff frequency ω' are those of a similar passive delay line, each capacitance of which includes a tube capacitance. The gain of a lossless chain is $ng_m/\omega' C$ per stage of n tubes, each of transconductance g_m. Amplitude and delay distortions of such chains in single and cascaded stages will be derived, including the effects of reflections, losses, and resonances on traveling-wave operation.

70. LOSSES IN AIR-CORED INDUCTORS

R. E. FIELD
(General Radio Company, Cambridge, Mass.)

A thorough study of the losses in air-cored inductors shows that the total loss is due to three components. These are conductor resistance, eddy-current effects, and dielectric loss. If these losses are expressed as dissipation factors, the total is the simple sum of the components. Formulas are derived for the component losses, and detailed methods for minimizing the losses and arriving at an optimum coil design are given. Detailed information on the choice of the type of winding as well as optimum design constants is given for several typical examples.

71. A SIMPLIFIED DESIGN PROCEDURE FOR IRON-CORE TOROIDS

H. E. HARRIS
(Massachusetts Institute of Technology, Cambridge, Mass.)

Mathematical expressions for the inductance and dissipation factors of iron-core toroids are first derived. A simple and effective graphical method of design is then given. The method leads to a determination of proper core size and permeability for the desired frequency, number of turns and total wire area for optimum Q, and the strand size if Litz wire is indicated. The limitations due to hysteresis and distributed capacitance are considered. An example of the method is detailed and the result compared with measurements made on a sample, constructed according to design specifications.

72. COUPLING EFFECTS BETWEEN INFRARED RADIATION AND A SUPERSONIC FIELD

W. J. FRY AND F. J. FRY
(University of Illinois, Urbana, Ill.)

The study of two types of coupling effects between a modulated beam of infrared radiation and a supersonically excited gas has been continued. The two gas combinations, air water vapor and CO_2 water, reported upon previously, have now been studied in greater detail, and more accurate values for the various quantities have been obtained. The magnitudes of the coupling effects have been studied as a function of gas composition, modulation frequency of the infrared shutter, and temperature of the source of radiation. A detailed analysis of the operation of the instrument has been carried out and checked by experimental observations. The mechanisms which were proposed are consistent with the present measurements.

Television

73. A UNITARY TUNER-AMPLIFIER FOR TELEVISION RECEIVERS

E. L. CROSBY, JR.
(Bendix Radio, Baltimore, Md.)
G. W. CLEVENGER
(Bendix Aviation Corporation, Baltimore, Md.)
AND H. GOLDBERG
(National Bureau of Standards, Washington, D. C.)

The unit to be described includes a 12-channel tuner using push-pull triodes; a stagger-tuned amplifier, the tuner being a component of the staggered system; compact effective traps; and high-level sound pickoff. The gain and bandwidth of the tuner vary less than 5 db and 15 per cent. The input circuit dynamically terminates a 300-ohm line with a v.s.w.r. of less than 1.25. Gain to the picture detector is 100 db. Sound is completely limited for a signal of 30 microvolts.

74. A PICTURE-MODULATED R.F. GENERATOR FOR TELEVISION RECEIVER MEASUREMENTS

ALLAN EASTON
(Hazeltine Electronics Corporation, Little Neck, N. Y.)

Present techniques of measurement of television receivers make desirable standard signal generators, operating on at least one of the thirteen commercially allocated channels, and capable of being fully modulated by any standard RMA composite video signal. This paper describes a generator of this type.

75. THE APPLICATION OF PROJECTIVE GEOMETRY TO THE THEORY OF COLOR MIXTURE

F. J. BINGLEY
(Philco Corporation, Philadelphia, Pa.)

The paper describes a new method of theoretical analysis for use in solving color-mixture problems. The method described provides, on the one hand, a powerful means of theoretical analysis characterized by the clearness of perception of geometric analysis, and on the other, a convenient graphical tool for obtaining numerical results with rapidity. Application of the method in the field of color television is described. Some new properties of color mixture discovered by the use of this method are described and discussed.

76. REFLECTION OF TELEVISION SIGNALS FROM TALL BUILDINGS

ANDREW ALFORD AND G. J. ADAMS
(Andrew Alford Laboratories, Boston, Mass.)

The intensity and the distribution of television ghosts produced by reflection from tall buildings depend on the location and height of the transmitting antenna. A prediction of ghosts which may be expected from a transmitting antenna located at a given site and height requires a detailed knowledge of the reflections from tall buildings. Theory, microwave model studies, and whole-scale measurements of reflections from buildings at a frequency in the lower television band are presented.

77. FIELD-COVERAGE CONSIDERATIONS OF NEW YORK TELEVISION STATIONS

T. T. GOLDSMITH, JR., AND R. P. WAKEMAN
(DuMont Research Laboratories, Passaic, N. J.)

A comprehensive study of the performance characteristics of DuMont television station WABD, New York, 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.

Broadcasting and Recording

78. MODERN DESIGN FEATURES OF CBS STUDIO AUDIO FACILITIES

R. B. MONROE AND C. A. PALMQUIST
(Columbia Broadcasting System, Inc., New York, N. Y.)

The design of a recently completed broadcasting-studio audio-control console, with facilities capable of handling the origination of the largest and most elaborate radio productions, is described. This unit, comparable in size to a standard office desk, is entirely self-contained. Many new and novel features are included, and the performance is well within requirements set forth for a.m., f.m., and television audio facilities. Although designed primarily for broadcasting, the fundamental ideas and methods are applicable to other services.

79. METHODS OF CALIBRATING FREQUENCY RECORDS

R. C. MOYER
(RCA Victor Division, Indianapolis, Ind.)
D. R ANDREWS AND H. E. ROYS
(RCA Victor Division, Camden, N. J.)

When making response measurements of disk reproduction systems, it is desirable to use a record of known calibration. The reflected-light-pattern method of calibrating frequency records is widely used and generally accepted. Other calibration means have been investigated, one of which uses a variable-speed turntable to reproduce the recorded tones at a common reference frequency. This method is particularly suitable for evaluation of the low frequencies where the light-pattern method is difficult to apply because the recorded amplitude and not the velocity is constant, resulting in a light pattern of varying width. Another method has been developed which is believed to be of merit because it offers a means of calibrating disks when the light pattern is indistinct, as it sometimes is with records cut at 33⅓ r.p.m. A comparison with the results obtained while using the f.m. calibrator for measuring the amplitude of stylus motion

during recording is given. Factors which enter in reproduction, such as tip size, force, mechanical impedance of the pickup, and contact between the groove and stylus, are discussed.

80. DISTORTIONS IN MAGNETIC-TAPE RECORDING DUE TO THE CONFIGURATION OF THE BIAS FIELD

S. J. Begun
(The Brush Development Company, Cleveland, Ohio)

It has been found that the field pattern around the gap of a ring-type magnetic recording head is a function of the wavelength to be recorded. Experiments have indicated that the field strength decays more rapidly in a direction away from the gap for shorter than for longer wavelengths, a phenomenon which has been called the penetration effect. It can be shown that in d.c. biasing no objectionable distortions are introduced as a consequence of this penetration effect. In a.c. biasing, on the other hand, distortions of low frequencies can be expected, particularly if a thick magnetic recording medium is used.

81. INSTANTANEOUS AUDIENCE-MEASUREMENT SYSTEM (IAMS)

P. C. Goldmark, J. W. Christensen, Andrew Bark, and J. T. Wilner
(Columbia Broadcasting System, Inc., New York, N. Y.)

The paper deals with a new system of measuring radio audiences, employing transponder techniques. The lecture will be accompanied with demonstrations.

Electronics III

Tube Manufacture

82. ASTM COMMITTEE WORK—FACTORY TESTS ON CATHODE NICKEL

J. T. Acker
(Western Electric Company, New York, N. Y.)

The paper deals with the methods of testing radio-tube cathode materials in factory production, and especially with a comparison of several specific lots of materials of variable content. It is believed that this is the first time the electron-tube industry has made mass tests on a well-controlled engineering basis of cathode materials which vary in single component elements.

83. A STANDARD DIODE FOR RADIO-TUBE-CATHODE CORE-MATERIAL APPROVAL TESTS

R. L. McCormack
(Raytheon Manufacturing Company, Newton, Mass.)

A diode has been designed and used for testing various samples of cathode material in several plants and laboratories during the last two years. Several criteria have been used for evaluating the emissive power of the various materials tested. To simulate the usual space-charge-limited emission test commonly used on receiving tubes, a cathode-temperature versus emission characteristic has been taken on each test lot. Temperature-limited emission has been examined under both low-field, low-temperature conditions and normal-temperature, high-field conditions. Results indicate that this method has several important advantages over the present approved method.

84. EUROPEAN PRACTICES IN THE MANUFACTURE OF CATHODES

T. H. Briggs
(Superior Tube Company, Norristown, Pa.)

The paper deals with European practices in the manufacture of cathodes and their use in radio tubes. These data were obtained by a group under Government auspices in September and October of 1947. Information is included on the many details in the making of cathode nickel and its alloys and its formation into filaments and heater cathodes. Information also is included on cathode coatings and other processing details.

85. PROCESSING VACUUM-TUBE COMPONENTS

P. D. Williams
(Eitel-McCullough, Inc., San Bruno, Calif.)

In power tubes the life of a thoriated filament is very sensitive to the processing employed, and this is also true of the phenomenon of grid emission. The processes used for the different elements also are often interrelated in their effects, and their close spacings and higher temperatures complicate the problem. The metallurgist and chemist can be very helpful in tube-manufacturing procedures to attain the desired results. Some of the means of accomplishing these results are described.

86. CONTINUOUS EXHAUST MACHINE FOR ELECTRON-TUBE MANUFACTURE

L. G. Hector
(Sonotone Corporation, Elmsford, N. Y.)

This machine for miniature and subminiature receiving-tube manufacture differs basically from standard exhaust machines in that the table rotates continuously. This machine also differs from conventional forms in that each position is individually exhausted with an oil-diffusion pump. In the machine now in use one fore pump handles two diffusion pumps. These pumps travel with the rotating table; consequently, all connecting vacuum lines are extremely short and all central valve systems are completely eliminated. The speed of the machine is limited primarily by operator ability.

Measurements I

V.H.F., U.H.F., and S.H.F.

87. SWEPT-FREQUENCY 3-CENTIMETER IMPEDANCE INDICATOR

H. J. Riblet
(Submarine Signal Company, Boston, Mass.)

An indicating system is described which will present on a cathode-ray tube sufficient information to determine the magnitude and phase of the impedance of a load at a number of closely spaced frequencies over a 12 per cent frequency range centered in the 3-cm. band. The first model of the equipment measures reflection coefficients with an accuracy, for low standing-wave ratios, of ±4 degrees in phase and ±8 per cent in magnitude. A novel r.f. circuit, called a wave sampler, makes the accuracy of this system independent of frequency.

88. AN AUTOMATIC V.H.F. STANDING-WAVE-RATIO PLOTTING DEVICE

W. A. Fails, L. L. Mason, and K. S. Packard
(Airborne Instruments Laboratory, Inc., Mineola, N. Y.)

A method is described for automatically plotting on an oscilloscope the diagram of the standing-wave ratio versus frequency for any impedance. The equipment comprises a modulated power oscillator covering an octave in the v.h.f. range. This oscillator feeds the unknown impedance through a pad, and a directional coupler is inserted between the pad and the load; the output of this coupler is detected by means of two bolometers, which supply two audio-frequency voltages whose ratio is measured electronically and plotted on the oscilloscope of an indicator unit.

89. MICROWAVE IMPEDANCE BRIDGE

M. Chodorow, E. L. Ginzton, and J. F. Kane
(Stanford University, Stanford, Calif.)

A twelve-terminal waveguide circuit can be formed by joining six waveguides symmetrically at a common junction. Experiments show that this circuit balances and behaves like a conventional Wheatstone bridge. A theoretical derivation of the equivalent circuit of the structure verifies these experiments and shows that the structure satisfies exactly the same admittance relationship as the conventional bridge circuit. In measuring microwave impedances, it is possible to use only sliding shorts as the adjustable impedances, and no variable resistance standard is needed. The value of the unknown impedance is determined from a measurement of three physical lengths.

90. IMPEDANCE MEASUREMENTS BY MEANS OF DIRECTIONAL COUPLERS AND SUPPLEMENTARY VOLTAGE PROBE

B. Parzen
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

This paper describes a method of measuring impedance at very-high, ultra-high, and microwave frequencies. Two directional couplers in direction opposition and one voltage probe are placed in a short section of line. From measurements of the output of the couplers and probe, calculations can be made of the impedance at the end of the line. An experimental model of the equipment for use at low frequencies (50 to 500 Mc.) is described. Impedance values as obtained by means of this equipment are compared with those obtained by means of a slotted line. The advantages and disadvantages of this method over the slotted-line, three-voltmeter, and bridge methods are discussed.

91. A WAVEGUIDE BRIDGE FOR MEASURING GAIN AT 4000 MC.

A. L. Samuel
(University of Illinois, Urbana, Ill.)
and C. F. Crandell
(Southwestern Bell Telephone Company)

A bridge has been constructed for measuring the gain and phase delay of amplifiers in the vicinity of 4000 Mc. The equipment is described, and the methods employed to reduce the possible errors are discussed. The general method may be adapted for use in any desired frequency range.

Computers I

Systems

92. LARGE-SCALE COMPUTERS

R. L. Snyder
(University of Pennsylvania, Philadelphia, Pa.)

Following a brief historical summary of the large-scale computer field, a precise delineation of the characteristics which differentiate modern devices will be made. A critical discussion of existing and under-construction machines will be given, after which some new ideas for computer components and systems will be presented.

93. THE UNIVAC

J. W. Mauchly
(Eckert-Mauchly Computer Corporation, Philadelphia, Pa.)

The Univac, a new large-scale general-purpose computing machine, will be described. It is a decimal machine which handles alphabetical as well as numerical information. Support of the development has come from the U. S. Bureau of the Census via a contract with the National Bureau of Standards.

94. ENGINEERING DESIGN OF A LARGE-SCALE DIGITAL COMPUTER

J. R. Weiner, C. F. West, and J. E. DeTurk
(Raytheon Manufacturing Company, Waltham, Mass.)

A general-purpose scientific computer capable of performing hundreds of arithmetic operations per second is described in terms of its major components: high-speed memory, central control, arithmetic unit, and input-output devices. Detailed designs of a mercury delay-line memory unit, a shift register, a parallel adder, a selection circuit, and an input-output unit are presented.

95. A NETWORK ANALYZER FOR THE STUDY OF ELECTROMAGNETIC FIELDS

K. Spangenberg, G. Walters, and F. W. Schott
(Stanford University, Stanford, Calif.)

An electric network which has been constructed along principles enunciated by Kron for analog studies of electromagnetic fields will be described. Virtually every type of problem associated with field configurations having rotational symmetry can be solved with the analyzer to be presented.

Propagation

96. CONTINUOUS TROPOSPHERIC SOUNDING BY RADAR

A. W. Friend
(Radio Corporation of America, Princeton, N. J.)

Original tropospheric-sounding experiments, which were conducted between 1.614 and 17.3 Mc. (from 1935 to 1941) produced positive results which were often somewhat difficult to interpret by simple theory. Additional tests were made on 2.398 Mc., during 1942, under more nearly idealized conditions; and on 2800 Mc., during 1946 and 1947, with extremely high power and sharply defined vertical beams. Recordings indicate many echo traces of various types, with echoes from cirrus and alto-stratus above 30,000 feet. Certain "dot" and other irregular reflections on 2800 Mc. are attributed to dielectric-gradient effects which are manifest in different forms on lower frequencies. Mathematical analyses are presented in support of theoretical deductions.

97. A THEORY ON RADAR REFLECTIONS FROM THE LOWER ATMOSPHERE

W. E. Gordon
(University of Texas, Austin, Tex.)

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

98. NEW TECHNIQUES IN QUANTITATIVE RADAR ANALYSIS OF RAINSTORMS

David Atlas
(Air Matériel Command, Wilmington, Ohio)

Experiments establish qualitatively the relation between echo power and rain intensity, and theory provides a basis for a simplified radar rain-intensity computer. Required and actual accuracy of measuring techniques are discussed. Techniques are described for (1) use of gain-setting at threshold of echo visibility on a PPI scope as measure of echo power; (2) "pip-matching" on a PPI scope; (3) construction of two- and three-dimensional "storm contour maps" showing isohyets throughout detectable rain area; (4) automatic presentation of such two-dimensional "maps" on a PPI scope by use of a simple modification; and (5) manually or automatically correcting distortion effects in maps introduced by rain attenuation and determination of true attenuation values. Limited discussion is presented regarding absolute accuracies and possible application of mapping to avoid hazardous storms.

99. THE PROPAGATION OF RADIO WAVES THROUGH THE GROUND

Knox McIlwain
(Hazeltine Electronics Corporation, Little Neck, N. Y.)
and H. A. Wheeler
(Consulting Radio Physicist, Great Neck, N. Y.)

A theoretical and experimental study of the propagation of radio waves through ground has resolved certain inconsistencies in prior work. Tests covered depths to several hundred feet and frequencies from 0.6 to 1000 Mc. As expected, dry ground is better than wet. At the lower frequencies, ground behaves as a homogeneous, poorly conducting medium; at the higher, the rate of attenuation increases much more rapidly, indicating pockets of moisture separated by dry ground. A special technique has been used to test the horizontal propagation through substrata, which is especially useful to detect and trace dry layers sandwiched between wet layers. The results show the limitations of radio waves for deep geophysical prospecting, though they may be useful for related exploration.

100. DESIGN AND APPLICATION OF A MULTIPATH TRANSMISSION SIMULATOR

H. F. Meyer and A. H. Ross
(Coles Signal Laboratory, Red Bank, N. J.)

Theory and design of a laboratory instrument for simulating the effects of multipath transmission encountered on long-haul high-frequency radio circuits are presented. The instrument provides for adjustable time delay between paths up to 3 milliseconds and for a choice of voice frequency amplitude, frequency, or phase modulation. The effects of multipath transmission on voice, single and multichannel voice-frequency teletype, and facsimile radio communication circuits utilizing various methods of modulation and detection may be studied. Correlation between theoretical computation of multipath effects in experimental data obtained with the simulator is demonstrated.

Electronics IV

New Forms of Tubes

101. NEW DESIGN FOR A SECONDARY-EMISSION TRIGGER TUBE—NU TR-1032-J

C. F. Miller and W. McLean
(National Union Radio Corporation, Orange, N J.)

The NU TR-1032-J is a nine-pin miniature tube with a triode input section producing a primary electron beam. This beam impinges on a secondary-emission surface, and secondary electrons are collected by two different output elements which may be used either separately or as a unit. Typical tube characteristics, a basic circuit description, and a new circuit in which this tube may be used are described, as well as suggested uses which include the following: relaxation oscillator, multivibrator, pulse inverter, modulator, oscillator, and dynatron.

102. A SPIRAL-BEAM METHOD FOR THE AMPLITUDE MODULATION OF MAGNETRONS

J. S. Donal, Jr., and R. R. Bush
(Radio Corporation of America, Princeton, N. J.)

A new method is described for the amplitude modulation of magnetrons. Although the method has so far been applied only to the modulation of an existing 1-kilowatt c.w. magnetron at 850 Mc., scaling to higher frequencies should prove to be perfectly feasible. In principle, a beam of electrons spiralling in a longitudinal magnetic field varies the conductance presented by a resonant cavity coupled to the magnetron and so varies the power delivered to the load. The linearity of the system is reasonably good and the bandwidth is at least 20 Mc. The depth of voltage modulation realized is 85 to 90 per cent, while the frequency variation during the amplitude-modulation cycle is only ± 15 kc.

103. THE DYOTRON—A NEW MICROWAVE OSCILLATOR

E. D. McArthur
(General Electric Company Schenectady, N. Y.)

The dyotron tube is structurally similar to a triode but differs in the use of a moderately noncritical transit angle and in the necessity for an r.f. short circuit between cathode and grid. Since the grid and cathode have no r.f. potential difference, there is no need for an input circuit or a feedback circuit. Experimental data of two kinds are presented. One type of experiment is designed to test the validity of the basic principles. The other kind comprises performance data of the type which will interest the application engineer and includes typical data on tuning range, power output, and frequency stability. The tubes which are discussed are experimental models; commercial types are not yet available.

104. ELECTROSTATICALLY FOCUSED RADIAL-BEAM TUBE

A. M. Skellett
(National Union Radio Corporation, Orange, N. J.)

In the electrostatically focused radial-beam tube, a combination of fields produces a single radial electron beam which may be rotated by rotating the uniform component of the combined fields. Details are given of such a tube with twelve anodes and twelve associated control grids which is no larger than an ordinary radio receiving tube. The beam current is of the order of 1 milliampere, and the frequency of rotation is limited only by the inductance and capacitance of the elements of the tube. This tube is an inertialess distributor with applications to time-division multiplex, telemetering, remote control, and other high-speed switching functions.

105. A NEW TWO-TERMINAL HIGH-VOLTAGE RECTIFIER TUBE

G. W. Baker
(Chatham Electronics Corporation, Newark, N. J.)

A new two-terminal high-voltage rectifier tube designed for use in the voltage-multiplying stages of a radio-frequency power supply is described, and samples will be shown. This new tube employs a phenomenon not previously applied to rectifier tubes. It has no heater circuit like cold-cathode tubes, but it has the low-voltage drop and high inverse-peak-voltage rating of hot-cathode tubes.

Measurements II

106. SIMPLIFICATION OF THEORY OF SUPERSONIC INTERFEROMETRY

F. E. Fox
(Catholic University)
and J. L. Hunter
(John Carroll University, Cleveland, Ohio)

Cady's equivalent piezoelectric circuit theory has been extended by Van Dyke, Hubbard, and Fox to cases of coupled fluid columns as in the supersonic interferometer. Interferometric measurements of sound absorption in liquids check theoretical predictions fairly closely, but unfortunately involve extensive theoretical interpretation. Correlation by most other methods is notoriously poor. A simplification of the theory, and corroborative data, are offered, with the hope of strengthening the claim of the interferometer to reliability of measurement.

107. FREQUENCY MEASUREMENT BY SLIDING HARMONICS

J. K. Clapp
(General Radio Company, Cambridge, Mass.)

The method is most easily outlined by describing a particular application. A 950-kc. crystal oscillator is combined with a stable 50- to 60-kc. oscillator to produce a frequency adjustable from 1000 to 1010 kc. A harmonic generator, controlled by this source, produces harmonics in the range from 100 to 200 Mc. Any frequency in this range can then be matched by sliding the next lower harmonic toward higher frequencies to obtain zero beat. The frequency is determined by the harmonic number and the interpolation-oscillator setting. Advantages include relatively high accuracy (25 in 10^6 or better), simplicity of operation, and no wide-range interpolating or receiver circuits.

108. A GENERAL-PURPOSE OSCILLOGRAPH FOR PRECISION TIME MEASUREMENT

R. P. Abbenhouse
(Allen B. DuMont Laboratories Inc., Clifton, N. J.)

This paper describes an oscillograph designed for general-purpose precision time measurements on transient or controlled phenomena with special features provided for use in connection with television signals. Circuits are discussed, and a brief summary of performance specifications and general applications is given. Its use as a television instrument is emphasized, showing how picture and synchronizing components of the transmitted composite signal may be measured in terms of resolution, rise time, duration, and periodicity.

109. SOME CONSIDERATIONS IN EXTENDING THE FREQUENCY RANGE OF RADIO NOISE METERS

W. J. Bartik and C. J. Fowler
(University of Pennsylvania, Philadelphia Pa.)

In the frequency region above 20 Mc. the electrical disturbances previously defined as radio noise affect other than audio services. Thus, in this part of the spectrum, at least, an objective type of meter capable of accurately measuring certain characteristics of the noise is indicated. From a measure of these characteristics the noise should be classified so that the interference value for various services can be determined. Some of the factors involved in choosing the proper characteristics to measure, and in the design of a suitable noise meter, are summarized. The requirements are such that two noise meters, a laboratory standard, and a portable field instrument may be necessary.

110. SOME CONSIDERATIONS IN THE DESIGN OF PRECISION TELEMETERING EQUIPMENTS

R. Whittle
(Federal Telecommunication Laboratories, Inc., Nutley, N. J.)

Problems encountered in precision telemetering systems are discussed, with particular attention to remote indication of position. Methods of transmitting information are compared for necessary bandwidth, inherent accuracy, effects of noise on accuracy, stability, and simplicity. The co-ordinate system is important in presenting position information, rectangular being somewhat simpler to transmit by radio. In radar systems, for instance, information is obtained in polar form and must be converted if rectangular co-ordinates are transmitted. Time sharing to reduce intermodulation distortion when multiple signals are telemetered is discussed. Phase comparison for telemetering continuous rotation through angles greater than 360 degrees is considered and a precise method described. Problems involving accuracies of 1 degree or better are discussed.

Computers II

Components

111. MEGACYCLE STEPPING COUNTER

C. B. Leslie
(Naval Ordnance Laboratory, Washington, D. C.)

The paper describes the development and construction of a modified ring- or stepping-type counter capable of operating at a megacycle rate. The device is one outcome of work on a large-scale high-speed digital computer at the Naval Ordnance Laboratory.

112. RECTIFIER NETWORKS FOR MULTIPOSITION SWITCHING

N. Rochester
(Sylvania Electric Products Inc., Flushing, N. Y.)
and D. R. Brown
(Massachusetts Institute of Technology, Cambridge, Mass.)

A new type of multiposition switch utilizing crystal rectifiers is used in electronic digital computers requiring switching times of less than one microsecond and in other applications demanding extreme compactness. Several varieties of this switch are analyzed, and applications and practical limitations are discussed.

113. MERCURY DELAY-LINE MEMORY USING A PULSE RATE OF SEVERAL MEGACYCLES

I. L. Auerbach, J. P. Eckert, Jr., R. F. Shaw, and C. B. Sheppard
(Eckert-Mauchly Computer Corporation, Philadelphia, Pa.)

A delay-line memory organ for an electronic computer has been constructed for operation at pulse-repetition rates of several megacycles per second. The high pulse rate makes possible the storage of considerably more information in a given space than was possible with previous types of memory organs. Recirculation circuits are described, as well as means for introducing and extracting information.

114. METHODS FOR VISUAL OBSERVATION OF PATTERNS RECORDED ON MAGNETIC MEDIA

S. N. Alexander, L. Marton, and I. L. Cooter
(National Bureau of Standards, Washington, D. C.)

Two methods to obtain additional information about the recording characteristics of magnetic media are described. One develops the magnetic field patterns by using a microscopic iron powder in a light oil surrounding the sample; the other employs an electron-optical technique for revealing the magnetized regions through the distortion they introduce into an electron beam. Examples are given in connection with electronic digital-computing-machine techniques.

115. SELECTIVE ALTERATION OF DIGITAL DATA IN A MAGNETIC DRUM COMPUTER MEMORY

A. A. Cohen and W. R. Keye
(Engineering Research Associates, Inc., St. Paul, Minn.)

Information coded in terms of binary digits (1's and 0's) may be recorded on magnetic tapes bonded to a continuously rotating drum. Such an experimental storage system is described, in which individual digits may be selectively altered. Recorded patterns containing 150 digits per linear inch are scanned at the rate of 200,000 digits per second.

Microwaves

116. CAVITY RESONATORS FOR HALF-MEGAVOLT OPERATION

A. E. Harrison
(Princeton University, Princeton, N. J.)

Losses in cavity resonators for high-voltage applications such as linear accelerators and high-power tubes have not received much attention, although various forms of electron loading have been investigated at low power levels. A technique was developed for determining the losses in a resonator at high power levels by measuring the Q under transient conditions with pulse power applied. Results indicate that losses from "multipactor" action of secondary electrons and field emission can be controlled. Approximately 500,000 volts peak was obtained across a three-quarter-inch gap. Curves giving data on dimensions of typical 3000-Mc. resonators are included.

117. ANALYSIS AND PERFORMANCE OF WAVEGUIDE HYBRID RINGS FOR MICROWAVES

H. T. Budenbom
(Bell Telephone Laboratories, Inc., Whippany, N. J.)

This paper presents an analytical treatment of waveguide hybrid rings for microwaves, considered as re-entrant transmission lines. The resulting lines are transformed into equivalent tee or lattice network sections, and determinantal methods are applied in analyzing these equivalent network assemblies for their transmission properties. Experimental data obtained on a carefully constructed sample of a three-arm and a four-arm ring are presented, and the good agreement between theory and experiment is noted.

118. FREQUENCY STABILIZATION WITH MICROWAVE SPECTRAL LINES

W. D. Hershberger and L. E. Norton
(Radio Corporation of America, Princeton, N. J.)

Absorption lines of gases at reduced pressure exhibit Q's of 100,000 in the 24,000-Mc. range, and the center frequency is unaffected by pressure and temperature. Stabilization of a "K"-band klystron has been effected, using the 23,870.1-Mc. line of ammonia contained in a short section of matched waveguide, both at the center frequency of the line itself and at frequencies removed from the line frequency by a controlled intermediate frequency. Indications are that the frequency stability compares favorably with that of quartz crystals, and applications to other microwave frequencies and a clock are shown.

119. SYNTHESIS OF DISSIPATIVE MICROWAVE NETWORKS FOR BROAD-BAND MATCHING

H. J. Carlin
(Polytechnic Institute of Brooklyn, Brooklyn, N. Y.)

Interpolation in the complex plane is employed to handle microwave network functions. This yields an approximating rational function over a specified bandwidth, and leads to a lumped-circuit approximation for the microwave structure, which is used as a basis for the synthesis of matching networks. In various problems involving dissipative devices, the poles of the rational approximating function may satisfy special conditions. In such cases the ideal lumped matching network has a simple realizable form, and may be transformed into a suitable microwave structure. Applications of this method and experimental results are given for the synthesis of broad-band coaxial attenuators.

120. ANALYSIS OF A MICROWAVE ABSOLUTE ATTENUATION STANDARD

A. B. Giordano
(Polytechnic Institute of Brooklyn, Brooklyn, N.Y.)

The analysis of a microwave absolute standard of attenuation will be presented. The system is comprised of a coaxial-line launcher and a coaxial-line receiver separated by a below-cutoff cylindrical waveguide section. The launching and receiving sections are coupled by means of exponentially decaying field components of the TM_{0m} modes. A method of matching the field components at the input and output discontinuity planes will be described. The method leads to the determination of the reactive attenuation and input impedance of the system as the length of the waveguide section is varied. A practical model will be described.

121. 10-CENTIMETER POWER-MEASURING EQUIPMENT

Theodore Miller
(Westinghouse Electric Corporation, East Pittsburgh, Pa.)

In part 1 on low-power measurements (100 microwatts to 5 milliwatts), the salient features of a power-measuring cavity, using Littelfuse bolometers, designed to cover the wavelength range from 9.0 to 10.5 centimeters with a power reflection of less than 1 per cent, are shown. The methods used to broad-band this cavity by means of resonant diaphragms are described. In part 2 on high-power measurements (100 watts to 1 kilowatt), a direct-reading water load using a continuously circulating water column matched into a waveguide system is described. The use of this load as a variable attenuator, and calibration methods, are discussed.

Receivers

122. THE APPLICATION OF NOISE THEORY TO THE DESIGN OF RECEIVERS

W. A. Harris
(Radio Corporation of America, Harrison, N. J.)

The mechanism by which noise is produced in an electron tube is discussed, and the relation between induced grid noise and plate noise is illustrated. An equivalent circuit with noise generators arranged to simulate the noise sources is then analyzed to determine the optimum noise figure attainable under various conditions. An appropriate figure of merit for tube noise is seen to be the frequency for which $R_{eq}g_{in}$ is unity. The frequencies corresponding to chosen values for the noise figure are presented for several receiving-tube types. The paper concludes with a discussion of the circuit conditions which must be met in order to obtain noise figures approximating the theoretical values.

123. THE DESIGN OF INPUT CIRCUITS FOR LOW NOISE FIGURE

M. T. Lebenbaum
(Airborne Instruments Laboratory, Inc., Mineola, N. Y.)

This paper extends the treatment of receiver input-circuit design for minimum noise figure to the case of the transitionally coupled double-tuned circuit with secondary loading. This type of input circuit affords considerable improvement over the simpler matching networks. The results of the analysis, including both "active" and "passive" tube loading, are presented in nomographic form. With the aid of the nomogram, it is possible to determine easily and rapidly the bandwidth of the input circuit that will give minimum noise figure. Simple design equations are given that may be used to determine the parameters of the double-tuned circuit required to achieve this bandwidth.

124. FREQUENCY CONVERTERS

W. H. Lewis
(Ordnance Research Laboratory, State College, Pa.)

This paper deals with multielectrode tubes having signals applied on two grids and having a plate load tuned to the difference frequency of the two signals. Presuming the converter tube to have high internal impedance, mathematical expressions are derived relating the mixer output to the tube parameters and load impedance. With appropriate assumptions, the relations are expressed in terms of power series and a Fourier expansion is made. The coefficients of interest are evaluated and it is shown that an exact expression for conversion gain may be obtained without requiring an analytic expression for the tube characteristics.

125. AN AUTOMATIC-TRACKING DIRECTION-FINDER RECEIVING SYSTEM FOR METEOROLOGICAL USE

William Todd
(Evans Signal Laboratory, Belmar, N. J.)

Radio Set AN/CRD-1 is an automatic-tracking meteorological radio direction finder which is a part of a system for determining the wind velocities as well as the radiosonde data through the atmosphere to about 100,000 feet by tracking a balloon-borne radio transmitter operating on a carrier frequency of 1680 Mc. Principles of operation, features of design, and application of many novel electronic principles in this equipment to yield an accuracy better than 0.05 degree in both elevation and azimuth are discussed.

Active Circuits

126. REACTANCE-TUBE CIRCUIT ANALYSIS

R. C. Maninger
(U. S. Navy Electronics Laboratory, San Diego, Calif.)

Exact expressions are derived for the equivalent series resistance and equivalent series reactance of the so-called "reactance"-tube circuits. From these expressions, conditions are derived under which it is shown that the circuits can behave as negative resistances, negative inductances, or negative capacitances. It is also shown that the Q of a reactance-tube circuit has a maximum theoretical value of $\sqrt{\mu}/2$ where μ is the amplification factor of the tube.

127. ELECTRONICALLY CONTROLLED REACTANCE

J. N. Van Scoyoc and J. L. Murphy
(Armour Research Foundation, Chicago, Ill.)

Equations are developed for feedback amplifier circuits whose input reactance can be varied electronically by an applied voltage signal. The signal is so applied as to vary the gain which is normally near unity within the feedback loop. One circuit tested was a two-stage amplifier, and the other a variation of the cathode-follower circuit. Experimental results are shown and applications discussed.

128. STABLE REGULATED POWER SUPPLIES

R. R. Buss
(Northwestern University, Evanston, Ill.)

High-order stability of a regulated power supply can be obtained by using for control action the nonlinear relationship between either the light output or electron emission and heating power of a filament. Combining either of the basic regulating actions with the degenerative control action of the conventional regulator gives instantaneous response and low internal impedance. Regeneration permits further control of the internal impedance. Operation of the regulator is analyzed and correlated with experimental data.

129. THE PHOTOFORMER

D. E. Sunstein
(Philco Corporation, Philadelphia, Pa.)

A simple general nonlinear element will be described, employing feedback principles with a cathode-ray tube, photocell, and a shaped mask. The law of output versus input provided is accurately expressed by the geometric shape of the mask, and may be readily changed at will. The device has application in computers, modulators, tone generators, demodulators, signal generators, and any circuit where a predescribed relationship shall exist between an input signal and output signal. Demonstrations and slides will illustrate the versatility of the apparatus.

130. MODE SEPARATION IN OSCILLATORS WITH TWO COAXIAL-LINE RESONATORS

H. J. Reich
(Yale University, New Haven, Conn.)

The resonance frequencies of a capacitance-terminated coaxial-line resonator, determined graphically, explain why separation of the various modes is possible by the use of two resonators. Mode separation is favored by using resonators with large difference in products of characteristic impedance by terminating capacitance. Excessive difference in CZ_0 products may, however, occasionally lead to coincidence of two modes. The graphical method may readily be used in the determination of theoretical tuning curves. Predicted curves agree in their general aspects with measured curves for a lighthouse-tube oscillator. Agreement is improved by considering tube elements as extensions of coaxial lines.

Industrial Engineering Notes[1]

New Developments By Bureau Of Standards

The January issue of "The Technical News Bulletin," a monthly publication of the National Bureau of Standards, carries a description of a new constant voltage power supply for use in certain electrolytic determination and separation processes which was developed by the Electronic Instrumentation Laboratory of the Bureau. Copies may be obtained from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C., at 10 cents each.

The February issue of this same publication gives an account of a new procedure for measuring the gain of hearing aids. The new procedure, which was developed by the sound laboratory of the National Bureau of Standards, is said to offer manufacturers and commercial laboratories a useful and economical method for maintaining adequate quality control of hearing aids. The equipment may be constructed at extremely low cost as compared to an expensive sound-insulated, echoless room. Hence control of the gain performance of hearing aids should now be readily available even to smaller manufacturers, the Bureau believes.

Simulated Blizzards Test Equipment

Man made blizzards on a continuous basis, equalling the intensity of storms experienced in the most remote Arctic regions, are being "manufactured" by engineers in the climatic test chambers of the Signal Corps at Fort Monmouth, N. J. This simulated condition under which military equipment is being tested is believed by the Army to be the first accomplishment of its kind.

Recent F.C.C. Nominations

Subject to Senate approval, President Truman recently named Wayne Coy, radio engineer of the *Washington Post*, to succeed Charles R. Denny as F.C.C. Chairman, and George E. Sterling (A'27–M'28–SM'43) to fill the unexpired term of Commissioner E. K. Jett. Mr. Coy's term expires June 30, 1951, and Mr. Sterling's term expires June, 1950.

F.C.C. To Extend F.M. Licenses

In two separate petitions, the National Association of Broadcasters and the Frequency Modulation Association asked the F.C.C. to extend the license periods of f.m. stations from one year to three. Both petitions cited f.m.'s growth and need for stability as justifications.

[1] The data on which these NOTES are based were selected, by permission, from "Industry Reports," published by the Radio Manufacturers' Association, issues of December 19, 1947, and January 2, 1948, and from "News," published by the Radio Manufacturers' Association of Canada, issue of December 20, 1947. The helpful attitude of these organizations in this matter is here gladly acknowledged.

Radio Advance Cited

The annual report of the F.C.C. for the 1947 fiscal year ending June 30, 1947, detailed the activities of the Commission during the year and noted the highlights of the radio industry's advances in that period. A.m. licensed broadcast stations had passed the 1000 mark at the end of the fiscal year, and both television and f.m. broadcasting services had doubled in number of stations authorized. The latter, including stations and operators, rose to nearly 550,000 authorizations.

Copies of the detailed F.C.C. annual report may be obtained from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C., at 25 cents each.

Proposed Amendment Of F.C.C. Rules

With the view of incorporating new allocations in the 152 to 162 Mc. band into its rules and to remove from the remote pickup rules the 30 to 40 Mc. frequencies, the F.C.C. proposed to amend its regulations governing experimental auxiliary broadcast services. Copies of the Commission's proposed changes (mimeograph No. 12957) may be obtained from the Secretary of the Federal Communications Commission, Washington 25, D. C.

Radio Primer On Sale

A companion publication to "An ABC of the F.C.C.," is "Radio—A Public Primer" which is now being sold by the Superintendent of Documents, Government Printing Office, Washington 25, D. C., for 10 cents a copy. It traces the development of radio, explains its operation, and reviews broadcast and other types of radio services.

Ship Radar License Form

Recently the F.C.C. approved a new ship radar license form for use in the Ship Services and adopted changes in its regulation to cover the new form. Copies of the form (F.C. C. Form 501-B) may be obtained from the Secretary of the Federal Communications Commission, Washington 25, D. C.

Additional Circuits For Television

In connection with a 1948 expansion program costing $76,130,000, the American Telephone & Telegraph Company and certain Bell System associated companies have been authorized by the F.C.C. to supplement existing facilities. For television, the program proposed to provide two additional circuits in the New York-Washington coaxial cable, two between Washington and Charlotte, two between New York and Albany, two between Philadelphia and Chicago, and two between Chicago and St. Louis. This would permit, according to the F.C.C. grant, television programs to originate or be received at Baltimore, Richmond, Pittsburgh, and Cleveland, in addition to the other cities named. Boston may be tied in by means of the experimental microwave circuits now existing between that city and New York. Until such time as the circuits are required for commercial use, they will be available for gaining experience in operating long distance television circuits and for training personnel along the routes involved.

Ship Radar Regulation

In December, 1947, the F.C.C. adopted amendments to its rules governing operator license requirements of ship radar stations. Under the new amendments unlicensed personnel may perform the normal operation on board ship of radar stations licensed in the Ship Service. They may not, however, make any adjustments or perform any servicing or maintenance that may affect the operation of the station. Copies of the F.C.C. order (No. 12342) may be obtained from the Secretary of the Federal Communications Commission, Washington 25, D.C.

3,110 Wartime Contracts Renegotiated

Late in December the Chief Signal Officer announced that the total gross refunds of excessive profits recovered by the Signal Corps are estimated at $480,000,000, with a net recovery of $265,000,000 after deducting the estimated credit granted contractors for income and excess profits taxes that had been paid on the gross recovery. To accomplish this the Signal Corps expended an estimated $1,275,000, which represents 0.27 of 1 per cent of the gross refunds and 0.48 of 1 per cent of the net refunds.

F.M. License Renewals Staggered

Under a new F.C.C. proposal f.m. licenses (including non-commercial educational) would be renewed for a period of one year to expire on the first of February, April, June, August, October, or December, depending upon the frequency assigned to each station. Outstanding f.m. licenses are not affected by the action.

Canadian School Equipment

The new Canadian RMA brochure on "School Sound Systems" was printed recently and approximately 3000 copies were circulated to Canadian schools which contain six rooms and over, and where interest in this type of equipment might be expected. The School Equipment Committee, under the chairmanship of Mr. F. W. Radcliffe, reviewed the new School Sound Recording and Playback Equipment brochure recently published by the American RMA, with a view of determining whether or not this information should also be prepared and circularized to Canadian school authorities.

New F.M. And Television Station Grants

Conditional grants for twenty-one new f.m. stations and a construction permit for a new television station at Atlanta, Ga., were issued toward the end of 1947 by the F.C.C.

F.C. C. records showed a total of 374 f.m. stations in operation at the end of 1947. New stations on the air are: Birmingham, Ala. (WSGN-FM); Bakersfield, Calif. (KERN-FM); Miami Beach, Fla. (WKAT-FM); Chicago, Ill. (WENR-FM); Detroit, Mich. (WXYZ-FM); Clayton, Mo. (KXLW-FM); Pottsville, Pa. (WPAM-FM); Los Angeles, Calif. (KECA-FM); Edinburg, Texas (KURVA-FM); Canton, Ohio (WHBC-FM); Herrin, Ill. (WJPF-FM); Syracuse, N. Y. (WNDR-FM); Las Vegas, Nev. (KENO-FM); Huntington, W. Va., (WPLH-FM); Marysville, Calif. (KMYC-FM); Manchester, N. H. (WMUR-FM), and Los Angeles, Calif. (KKLA).

November Excise Collection

The following figures show collections in November on radio sets, phonographs, and their component parts to be approximately $55,000 below the October collections; November, 1947, $5,458,021.54; October, 1947, $5,513,134.48. In November, 1946, excise collections on these same articles amounted to $4,870,807.15.

Canadian Government Regulations

On November 18, 1947, new government regulations went into effect covering both an increase in excise tax from 10 to 25 per cent, and new import regulations. This resulted in a number of problems requiring action both by individual member companies and by the Association.

The effect of the new import embargo and other restrictions on the Canadian radio industry depend to a considerable extent on the success of the efforts of the Canadian Furniture Manufacturers' Association to have the embargo on imported veneers lifted. The direct importation of face veneer for the first nine months of 1947 amounted to a total of 21,894,354 square feet, having a total value of $722,681. This importation is at the rate of 29,000,000 square feet per year and the current rate of production of the two Canadian companies manufacturing face veneer is 18,800,000 square feet per year, or not more than 39 per cent of the total Canadian consumption of face veneer. Although the Canadian manufacturers plan to increase their capacity by over 50 per cent it still would mean that 56 per cent of the veneers which normally have been imported will be the measure of a shortage in Canadian supply, unless some relief is granted.

The Emergency Exchange Conservation Act indicates that no permit shall be issued for the import of goods listed in Schedule "I" (embargo section) unless, in the opinion of the Minister, exceptional hardship would result. The Emergency Import Control Division have indicated that Radio receivers and chassis will be permitted entry when required for special engineering purposes only. Also, it is understood that some companies have been granted special import permits for cabinets required to complete the manufacture of receivers, when the receivers are complete, except for cabinets on order in the United States. Establishment of such hardship cases must be submitted to the Special Import Division.

Canadian Production

Canadian RMA member company radio receiver sales and inventories reports for November, 1947 indicated a new industry record with production of 103,329 units compared with the previous record of 94,409 units in October, and sales of 102,300 radio receivers having a total retail value of $7,703,854 as compared with the previous record monthly sales of 86,991 units at $6,591,001 in October. Total production of RMA member companies (not including exports) for the first eleven months of 1947 was 833,349 units and manufacturers' Canadian sales in the same period total 721,370 receivers having an over all total retail value of $49,525,147.

Exports of Canadian made radio receivers for the first eleven months of 1947 totaled 51,469 with a value of $1,540,967.

RMA Meeting in New York

Many new activities were considered at the meetings of the Marine Section and the Piezoelectric Quartz Crystal Section, RMA Transmitter Division, under Chairman C. E. Maass and G. E. Wright, respectively. in New York on December 15 and 16, 1947 Meetings of the Aviation Transmitter Broadcast, General Communications and Transmitter Tube Sections were held prior to the RMA midwinter conclave, January 20 to 22, 1948.

Canadian RMA Broadcast Relations Committee

The newly formed Broadcasting Relations Committee, under the chairmanship of A. B. Hunt (A'43–SM'43) held its first meeting in Montreal on December 5, 1947, and its first report was placed before the board at the January 7 meeting. Mr. Hunt's committee includes F. R. Deakins, W. M. Angus (A'41), W. Dixon, R. A. Hackbusch (A'26–M'30–F'37) and M. M. Elliott. It was set up to deal with all matters of mutual interest to the radio manufacturers and broadcaster groups and to afford liaison with the Canadian Association of Broadcasters, Canadian Broadcasting Corporation and the Department of Transport.

Canadian Service Committee

The RMA Service Committee met at London, Ontario, January 23. The Committee has made considerable progress in the preparation of further bulletins for Canadian service technicians and a number of other important bulletins are under preparation.

Canadian Engineering Committee

Under the chairmanship of S. Sillitoe (A'35) the RMA Engineering Committee met in Montreal on December 4, 1947, and prepared further recommendations for the proposed new CSA Specification C22.2 No. 1 "Power-Operated Radio Devices" (third edition). Supporting data for the requested 5 milliampere leakage current was reviewed and prepared for submission to the CSA.

The Engineering Committee has unanimously approved the Mc. marking for f.m. dials.

A subcommittee, under the chairmanship of A. B. Oxley (A'25–M'33–SM'43) is making a special survey of oscillator radiations from superheterodyne receivers.

1948 Activities Of Canadian RMA

The joint Canadian-American RMA Directors' meeting will be held in Canada in 1948. The location is the Royal York Hotel, Toronto, and the dates are April 8 and 9.

The Canadian I.R.E. Convention will be held at the Royal York Hotel, Toronto, on Friday and Saturday, April 30, and May 1, 1948. It is planned to hold a number of the RMA division and committee meetings in conjunction with the I.R.E. Convention to facilitate attendance by RMA member company representatives.

The Canadian RMA Annual Meeting will be held at the Royal York Hotel on Tuesday, June 15, 1948.

RMA Meetings

The following RMA engineering meetings have been held:

January 7—Committee on Audio Facilities
January 9—Subcommittee on Tube Sockets
January 9—Subcommittee on Vacuum Capacitors
January 13—Committee on Cathode-Ray Tubes
January 13—Transmitter Tube Committee
January 14—General Communications Committee
January 15—Broadcast Transmitter Committee
January 16—Committee on Aviation
January 22—Committee on Traffic.

Calendar of COMING EVENTS

I.R.E. National Convention
March 22–25, 1948

Chicago I.R.E. Conference
April 17, 1948

Cincinnati Spring Meeting
April 24, 1948

Syracuse RMA-I.R.E. Spring Meeting
April 26–28, 1948

Canadian I.R.E. Convention
April 30 and May 1, 1948

I.R.E.-URSI Meeting
May 3–5, 1948

New England Radio Engineering Meeting
May 22, 1948

1948 West Coast Convention of the I.R.E.
September 30-October 2, 1948

Books

Radar Aids to Navigation, edited by John S. Hall, L. A. Turner, and R. M. Whitmer

Published (1947) by McGraw-Hill Book Co., 330 West 42 Street, New York 18, N. Y. 382 pages+7-page index+xiii pages. 191 figures. 6×9 inches. Price $7.50.

The book entitled "Radar Aids to Navigation" is the second in the Radiation Laboratory Series. It was written by thirty-three authors and edited by L. A. Turner, J. S. Hall, and R. M. Whitmer. The preface of this book states that it was intended primarily to describe the advantages and limitations of radar equipment when applied to problems of navigation and pilotage.

In the foreword, written by Dr. L. A. Du Bridge, the secondary purpose of the book is revealed by inference from the statement that (speaking of the entire series) "these volumes stand as a monument to this Group," the staff of the Radiation Laboratory.

The preface states that the book is written in a nontechnical form. Actually a certain number of algebraic equations are present in the volume and, therefore, the writing cannot be described as completely nontechnical. The greatest percentage of the material it contains can be readily understood by persons having the equivalent of a high school education.

The book contains four sections. These sections are the "Introduction," in two chapters; "Airborne Radar," in four chapters; "Ground-Based Radar," in three chapters; and "Shipborne Radar," in two chapters. Data are given on the various radar navigational devices developed during the war (principally by the Radiation Laboratories). In addition, it describes a number of navigational developments not produced by the Radiation Laboratory, but produced by other organizations. Some of these devices are of a type developed prior to the past war; thus, in Chapter 2, there is found a description of the four-course radio range, aircraft direction finders, ground and shore direction finders, and celestial navigation.

As a portion of the description of the various radar devices, pertinent data such as peak power, beam width, and other important characteristics are given; thus, the publication serves as a valuable equipment handbook. The book contains a large number of pictorial presentations. In these illustrations, the many pictures of the radar scopes are particularly noteworthy, since pictures of this character are not easily reproducible.

This book should be of particular interest to those who deal with navigational problems and who were not closely associated with the wartime developments as, in the course of a few evenings, it enables the reader to obtain a broad general knowledge of the field.

Reading over the description of the various pieces of apparatus, the critical reader will find that they are sometimes incomplete. He will also find that the authors occasionally use terms and abbreviations which are not defined, and unless he has had previous connection with the field, or can make reference to other text, he will be unable to understand their meaning. These omissions, no doubt are indications of the haste in which the volume was prepared. Reference is made to other volumes in the series and they may make up for the omissions in "Radar Aids to Navigation."

The reader may wonder why a description is included of such pre-wartime navigational devices as the four-course radio range. It might be presumed that this material was included in order to make the volume complete, but if this were the case some of the important wartime navigational devices such as the SCS-51, instrument landing system, the Oboe navigation system and the YG beacon, presumably should have been included.

While the preface states that emphasis is placed more on what can now be done with radar, than on what should be possible in the future, the impression will be gained that the purpose of the volume is about equally divided between describing the wartime navigational radar developments and showing how and why they should be applied to civilian tasks. In attempting to show the application of radar to civilian usage, the authors discuss some problems which they have not had the opportunity to study in great detail and, therefore, portions of their conclusions may be subject to questioning by experts in these fields. This speculation on postwar usage may serve to detract from the otherwise very creditable accomplishment.

There is a great need for a book such as "Radar Aids to Navigation," and the present volume goes a long way toward filling this need; but it is regrettable that the authors did not spend a greater effort in describing the work which they have done, thereby leaving to posterity a valuable technical history and creating for themselves a truly great monument of a great work.

Peter C. Sandretto
International Telephone and Telegraph Corp.
New York 4, N. Y.

Electronic Transformers and Circuits, by Reuben Lee

Published (1947) by John Wiley & Sons, 440 Fourth Avenue, New York 16. N. Y. 260 pages+12-page index+4-page bibliography+6-page appendix+vii pages. 209 figures. $9\frac{1}{4}\times 6$ inches. Price $4.50.

This book is a detailed, meticulous exposition of the *design of transformers for electronic circuits*. In fact, that should have been the title, since the present one might be misconstrued to refer to electronic impedance changing devices.

The first two chapters review elementary transformer theory and contain a very complete practical discussion of transformer construction, with useful design charts. The plug for Fosterite seems a trifle optimistic, since war experience showed that the only thoroughly satisfactory small transformers for tropical use were hermetically-sealed oil-filled cases. There follows a two-chapter exposition of the design of transformers for use with rectifiers, with examples. This first section of the book is excellent. Unfortunately, not so good an account can be given of the next four chapters. These contain much vacuum-tube circuit theory which is necessarily incomplete. This space could much better have served for discussions of the special design problems involved in saturable reactors, magnetic amplifiers, peaking transformers, and so forth. The final chapter on Pulse Transformers regains the satisfactory level of the first four.

The author states that one purpose of the book is "to furnish electronic equipment engineers with an understanding of the effect of transformer characteristics on electronic circuits." This purpose the book substantially fulfills. His other purpose is "to provide a reference book on the design of transformers for electronic circuits." The reviewer is not a transformer designer so the design techniques expounded seem good to him. From his experience with designers in general, he would anticipate that other transformer designers might differ sharply with Mr. Lee's methods.

For a new volume the number of typographical errors is remarkably small. The main detail criticism of the book is the occasional appearance of a completely bewildering sentence: "A transformer having an open-circuit secondary has twice the voltage and gives the same response at twice the 'low-end' frequency of a line matching transformer of the same turns ratio," doubtless has, or once had, a meaning but it is not obvious. The recurrence of such solecisms was frequent enough to be annoying.

On the whole the book is a workmanlike job, filled with valuable design data and charts, and is a valuable addition to the literature on transformers.

Knox McIlwain
Hazeltine Electronics Corp.
Little Neck, L. I.

Standards Errata

The third sentence of paragraph 2.14 of the I.R.E. "Standards on Radio Receivers, Methods of Testing Frequency-Modulation Broadcast Receivers—1947," recently distributed to all voting members of the Institute, should read: "The capacitance in farads is equal to 75×10^{-6} divided by the resistance in ohms."

Sections

Chairman		Secretary
P. H. Herndon c/o Dept. in charge of Federal Communication 411 Federal Annex Atlanta, Ga.	ATLANTA March 19	M. S. Alexander 2289 Memorial Dr., S.E. Atlanta, Ga.
F. W. Fischer 714 Beechfield Ave. Baltimore 29, Md.	BALTIMORE	E. W. Chapin 2805 Shirley Ave. Baltimore 14, Md.
John Petkovsek 565 Walnut Beaumont, Texas	Beaumont— Port Arthur	C. E. Laughlin 1292 Liberty Beaumont, Texas
W. H. Radford Massachusetts Institute of Technology Cambridge, Mass.	BOSTON	A. G. Bousquet General Radio Co. 275 Massachusetts Ave. Cambridge 39, Mass.
A. T. Consentino San Martin 379 Buenos Aires, Argentina	BUENOS AIRES	N. C. Cutler San Martin 379 Buenos Aires, Argentina
R. G. Rowe 8237 Witkop Avenue Niagara Falls, N. Y.	BUFFALO-NIAGARA March 17	R. F. Blinzler 558 Crescent Ave. Buffalo 14, N. Y.
G. P. Hixenbaugh Radio Station WMT Cedar Rapids, Iowa	CEDAR RAPIDS	W. W. Farley 1920 Fourth Ave. S.E. Cedar Rapids, Iowa
Karl Kramer Jensen Radio Mfg. Co. 6601 S. Laramie St. Chicago 38, Ill.	CHICAGO March 19	D. G. Haines Hytron Radio and Electronics Corp. 4000 W. North Ave. Chicago 39, Ill.
J. F. Jordan Baldwin Piano Co. 1801 Gilbert Ave. Cincinnati, Ohio	CINCINNATI March 16	F. Wissel Crosley Corporation 1329 Arlington St. Cincinnati, Ohio
W. G. Hutton R.R. 3 Brecksville, Ohio	CLEVELAND March 25	H. D. Seielstad 1678 Chesterland Ave. Lakewood 7, Ohio
C. J. Emmons 158 E. Como Ave. Columbus 2, Ohio	COLUMBUS April 9	L. B. Lamp 846 Berkeley Rd. Columbus 5, Ohio
L. A. Reilly 989 Roosevelt Ave. Springfield, Mass.	CONNECTICUT VALLEY March 18	H. L. Krauss Dunham Laboratory Yale University New Haven, Conn.
Robert Broding 2921 Kingston Dallas, Texas	DALLAS-FT. WORTH	A. S. LeVelle 308 S. Akard St. Dallas 2, Texas
E. L. Adams Miami Valley Broadcasting Corp. Dayton 1, Ohio	DAYTON March 18	George Rappaport 132 E. Court Harshman Homes Dayton 3, Ohio
A. Friedenthal 5396 Oregon Detroit 4, Mich.	DETROIT March 19	N. C. Fisk 3005 W. Chicago Ave. Detroit 6, Mich.
N. J. Reitz Sylvania Electric Products, Inc. Emporium, Pa.	EMPORIUM	A. W. Peterson Sylvania Electric Products, Inc. Emporium, Pa.
F. M. Austin 3103 Amherst St. Houston, Texas	HOUSTON	C. V. Clarke, Jr. Box 907 Pasadena, Texas
R. E. McCormick 3466 Carrollton Ave. Indianapolis, Ind.	INDIANAPOLIS	M. G. Beier 3930 Guilford Ave. Indianapolis 5, Ind.
C. L. Omer Midwest Eng. Devel. Co. Inc. 3543 Broadway Kansas City 2, Mo.	KANSAS CITY	Mrs. G. L. Curtis 6003 El Monte Mission, Kansas
R. C. Dearle Dept. of Physics University of Western Ontario London, Ont., Canada	LONDON, ONTARIO	E. H. Tull 14 Erie Ave. London, Ont., Canada
Walter Kenworth 1427 Lafayette St. San Gabriel, Calif.	LOS ANGELES March 16	R. A. Monfort L. A. Times 202 W. First St. Los Angeles 12, Calif.
O. W. Towner Radio Station WHAS Third & Liberty Louisville, Ky.	LOUISVILLE	D. C. Summerford Radio Station WHAS Third & Liberty Louisville, Ky.
E. T. Sherwood Globe-Union Inc. Milwaukee 1, Wis.	MILWAUKEE	J. J. Kircher 2450 S. 35th St. Milwaukee 7, Wis.
R. R. Desaulniers Canadian Marconi Co. 211 St. Sacrement St. Montreal, P.Q., Canada	MONTREAL, QUEBEC March 17	R. P. Matthews Federal Electric Mfg. Co. 9600 St. Lawrence Blvd. Montreal 14, P.Q., Canada
J. E. Shepherd 111 Courtenay Rd. Hempstead, L. I., N. Y.	NEW YORK April 7	I. G. Easton General Radio Co. 90 West Street New York 6, N. Y.
L. R. Quarles University of Virginia Charlottesville, Va.	NORTH CAROLINA-VIRGINIA	J. T. Orth 4101 Fort Ave. Lynchburg, Va.
K. A. Mackinnon Box 542 Ottawa, Ont. Canada	OTTAWA, ONTARIO March 18	D. A. G. Waldock National Defense Headquarters New Army Building Ottawa, Ont., Canada
P. M. Craig 342 Hewitt Rd. Wyncote, Pa.	PHILADELPHIA April 1	J. T. Brothers Philco Radio and Television Tioga and C Sts. Philadelphia 34, Pa.
E. M. Williams Electrical Engineering Dept. Carnegie Institute of Tech. Pittsburgh 13, Pa.	PITTSBURGH April 12	E. W. Marlowe 560 S. Trenton Ave. Wilkinburgh PO Pittsburgh 21, Pa.
O. A. Steele 1506 S.W. Montgomery St. Portland 1, Ore.	PORTLAND	F. E. Miller 3122 S.E. 73 Ave. Portland 6, Ore.
N. W. Mather Dept. of Elec. Engineering Princeton University Princeton, N. J.	PRINCETON	A. E. Harrison Dept. of Elec. Engineering Princeton University Princeton, N. J.
A. E. Newlon Stromberg-Carlson Co. Rochester 3, N. Y.	ROCHESTER March 18	J. A. Rodgers Huntington Hills Rochester, N. Y.
E. S. Naschke 1073-57 St. Sacramento 16, Calif.	SACRAMENTO	
G. M. Cummings 7200 Delta Ave. Richmond Height 17, Mo.	ST. LOUIS	N. J. Zehr Radio Station KWK Hotel Chase St. Louis 8, Mo.
C. N. Tirrell U. S. Navy Electronics Lab. San Diego 52, Calif.	SAN DIEGO April 6	S. H. Sessions U. S. Navy Electronics Lab. San Diego 52, Calif.
L. E. Reukema Elec. Eng. Department University of California Berkeley, Calif.	SAN FRANCISCO	W. R. Hewlett 395 Page Mill Rd. Palo Alto, Calif.
W. R. Hill University of Washington Seattle 5, Wash.	SEATTLE April 8	W. R. Triplett 3840—44 Ave. S.W. Seattle 6, Wash.
C. A. Priest 314 Hurlburt Rd. Syracuse, N. Y.	SYRACUSE	R. E. Moe General Electric Co. Syracuse, N. Y.
C. A. Norris J. R. Longstaffe Ltd. 11 King St., W. Toronto, Ont., Canada	TORONTO, ONTARIO	C. G. Lloyd 212 King St., W. Toronto, Ont., Canada
O. H. Schuck 4711 Dupont Ave. S. Minneapolis 9, Minn.	TWIN CITIES	B. E. Montgomery Engineering Department Northwest Airlines Saint Paul, Minn.
G. P. Adair 1833 "M" St. N.W. Washington, D. C.	WASHINGTON April 12	H. W. Wells Dept. of Terrestrial Magnetism Carnegie Inst. of Washington Washington, D. C.
J. C. Starks Box 307 Sunbury, Pa.	WILLIAMSPORT April 2	R. G. Petts Sylvania Electric Products, Inc. 1004 Cherry St. Montoursville, Pa.

SUBSECTIONS

Chairman		Secretary
P. C. Smith, 179 Ido Avenue, Akron, Ohio	Akron (Cleveland Subsection)	J. S. Hill, 51 W. State St., Akron, Ohio
J. D. Schantz, Farnsworth Television and Radio Company, 3700 E. Pontiac St., Fort Wayne, Ind.	Fort Wayne (Chicago Subsection)	S. J. Harris, Farnsworth Television and Radio Co., 3702 E. Pontiac, Fort Wayne 1, Ind.
F. A. O. Banks, 81 Troy St., Kitchener, Ont., Canada	Hamilton (Toronto Subsection)	E. Ruse, 195 Ferguson Ave., S. Hamilton, Ont., Canada
A. M. Glover, RCA Victor Div., Lancaster, Pa.	Lancaster (Philadelphia Subsection)	C. E. Burnett, RCA Victor Div., Lancaster, Pa.
E. J. Isbister, 115 Lee Rd., Garden City, L. I., N. Y.	Long Island (New York Subsection)	F. Q. Gemmill, Sperry Gyroscope, Great Neck, L. I., N. Y.
A. D. Emurian, HDQRS. Signal Corps Engineering Lab., Bradley Beach, N. J.	Monmouth (New York Subsection)	Ralph Cole, Watson Laboratories, Red Bank, N. J.
J. B. Minter, Box 1, Boonton, N. J.	Northern N. J. (New York Subsection)	A. W. Parkes, Jr., 47 Cobb Rd., Mountain Lakes, N. J.
A. R. Kahn, Electro-Voice, Inc., Buchanan, Mich.	South Bend (Chicago Subsection) February 19	A. M. Wiggins, Electro-Voice, Inc., Buchanan, Mich.
W. M. Stringfellow, Radio Station WSPD, 136 Huron Street, Toledo 4, Ohio	Toledo (Detroit Subsection)	M. W. Keck, 2231 Oak Grove Place, Toledo 12, Ohio
R. M. Wainwright, Elec. Eng. Department, University of Illinois, Urbana, Illinois	Urbana (Chicago Subsection)	M. H. Crothers, Elec. Eng. Department, University of Illinois, Urbana, Illinois
W. A. Cole, 323 Broadway Ave., Winnipeg, Manit., Canada	Winnipeg (Toronto Subsection)	C. E. Trembley, Canadian Marconi Co., Main Street, Winnipeg, Manit., Canada

I.R.E. People

George E. Sterling

George E. Sterling

President Truman nominated George E. Sterling (A'27–M'28–SM'43) Federal Communications chief engineer, to succeed Commissioner E. K. Jett, who resigned on December 31, 1947, to become vice-president and radio director of the Baltimore Sunpapers.

Mr. Sterling was born at Peaks Islands, Portland, Me., on June 21, 1894. He attended Johns Hopkins University and Baltimore City College. In 1908 he established his first amateur station at his home in Maine, and in 1913 he obtained his amateur license; one of the first in that state. He has been continuously associated with radio since that date, except for a brief period during World War I. In 1916 he was on the Mexican border with Company "M" of the Second Maine Infantry, and later, overseas with the 103rd Infantry, 26th Division. Transferring to the United States Signal Corps. he served 19 months in the American Expeditionary Forces in France and was a radio instructor in the Signal Corps schools. On completion of Officers' Training School at Langres, France, he was commissioned a Second Lieutenant, Signal Corps Reserve, and assisted in organizing and operating the first radio intelligence section of the Signal Corps in France, which located enemy radio stations and intercepted their messages. For this work he received a citation from the Chief Signal Officer of the AEF for "especially excellent and meritorious service." In 1923 Mr. Sterling entered the Federal service as a radio inspector in the Bureau of Navigation, Department of Commerce.

He is the author of "The Radio Manual," which is recognized and used extensively as a standard textbook on radio communications equipment and procedure by radio schools and for Government training purposes and as a reference book by colleges and universities.

Mr. Sterling served as a delegate of the Provisional International Civil Aviation Organization at the Demonstrations of Radio Aids to Air Navigation by the United Kingdom at London, from September 7 to October 5, 1946, and subsequently by the government at Indianapolis, from October 9 to 18, 1946. He was chairman of the United States delegation to the engineering conference looking toward the third NARBA meeting, which convened in Havana in November of last year.

Orville M. Dunning

Orville M. Dunning (A'34–SM'44), chief engineer of the Military Products Division Hazeltine Electronics Corporation, Little Neck, L. I., was awarded a Certificate of Commendation by the United States Navy for his achievements during World War II. The certificate was accompanied by a citation reading in part, "This award is made for your outstanding ability and unremitting effort . . . in designing airborne radar identification equipment as a generic series of standard units. The production design was so successfully co-ordinated that several manufacturers were able to produce, operating simultaneously, tens of thousands of such equipments."

David M. Davis

David M. Davis

David McClure Davis (M'43–SM'43), drowned on August 7, 1947, after saving his daughter Rebecca. He was picnicking on a small island near Ft. Walton, Florida, together with his wife, Mary, and their two daughters, Rebecca and Sally. On the return trip to the mainland the motor stalled, and then, as it started suddenly, the boat shipped water and capsized.

Davis was born in Laurel, Miss., on September 23, 1909, and graduated from Princeton University in 1931 with a degree in engineering. In 1936 he obtained a law degree from George Washington University. He was associated with the General Electric Company for some time and in 1942 joined the Zenith Radio Corporation as patent counsel. He was a member of the American Patent Law Association, and he was admitted to practice in Illinois and the District of Columbia and before Court of Customs and Patent Appeals, the United States Court of Claims, and the United States Supreme Court.

Robert G. Rowe

CHAIRMAN, BUFFALO-NIAGARA SECTION

R. G. Rowe (A'43–M'46) was born in North Tonawanda, N. Y. on December 13, 1915. He became interested in radio at the age of twelve and secured an amateur radio license at the age of fifteen years. He was graduated from the University of Michigan in 1938 with a B. A. degree, majoring in chemistry and minoring in economics. While attending college he was active in the Signal Corps unit of the Reserve Officers Training Corps and was president of the Michigan chapter of Sigma Phi Epsilon fraternity.

Upon graduation he became associated with the Rowe Paint & Varnish Co., Inc., a paint-manufacturing concern headed by his father and located at Niagara Falls, N. Y. In 1939 he was made vice-president of the aforementioned company, where his duties included research in the formulation of special protective coatings for local chemical industries. Having an unsatiated interest in the field of radio and applied electronics, in 1941 he accepted a position with the research department of the Carborundum Company of Niagara Falls, where he was employed to head a group working on high-frequency generators for ultrasonics and dielectric and induction heating. He now holds the position of senior research engineer in charge of the electrical and electronics section of the Carborundum Company research department.

During the war he was responsible for the design of vacuum-tube-operated devices to speed up the testing of the increased volume of abrasive and refractory materials produced by his company, his group having developed and constructed five different types of testing equipment used for this purpose. He is assignee to his company of three United States and foreign patents and twelve United States and foreign patent applications. The patent work required by his particular inventions, being initially alien to the regular scope of his company's patent department, gave him the opportunity to work with this group at some length, and thereby enabled him to personally process some six applications released to him. In the past two years he has written eight articles for radio periodicals.

In his spare time, Mr. Rowe, who holds commercial radiotelegraph and radiotelephone licenses, has set up the Mobile Radio Service to lease and install radiotelephone communication equipment for local taxicab and other mobile services. He also acts as technical advisor for a local antenna manufacturer, and sporadically operates amateur radio station W2FMF.

The Weston Electrical Instrument Corporation

WESTON OPENS NEW ENGINEERING AND ADMINISTRATION BUILDING

The three-story, T-shaped building which the Weston Electrical Instrument Corporation has recently built at Newark, N.J., enlarges the Weston plant to 380,000 square feet of floor area in 19 buildings and allows for expansion of engineering and administration facilities.

Preparing the Oral Version of a Technical Paper*

WILLIAM J. TEMPLE†, ASSOCIATE, I.R.E.

Summary—**The presentation of technical information in print and by word of mouth before professional groups should be regarded as different techniques. Some practical suggestions for the preparation of material for oral presentation are offered.**

RECENT PAPERS and editorials in the PROCEEDINGS OF THE I.R.E.[1,2] have strongly advised that, in presenting technical material before professional groups, an author should never merely read aloud the paper he has prepared for publication in print. It is the purpose of this article to make a few practical suggestions for the preparation of an oral version of such a paper. Most of these suggestions are derived from standard textbooks.[3,4]

If you will recall your own experiences as a listener in a meeting and as a reader at your own desk, you will see that there are very good reasons for you, as an author, to adopt two different methods of composing your material to suit these two different situations.

At your own desk, with a printed article before you, you can peruse it at your own pace and take as much time as you need to absorb it. You can read and re-read a crucial paragraph. You can take the time to puzzle out the meaning of a difficult or complicated sentence. If the writer uses an unfamiliar word, you can look it up. If you are interrupted, or if your attention wanders, or if you go to sleep, you can resume your reading later; going back to the beginning of the article, if need be, to pick up the thread of development. As a listener, you can do none of these things.

These reflections indicate that the oral version of a paper should differ from its printed version in two important ways. Designed for aural rather than visual reception, it should be shorter, and it should be simpler in structure. And, if you are going to avoid the soporific monotony with which almost everyone except a professional actor or a radio announcer reads aloud, you must prepare to present your material by *talking*, not reading aloud or reciting from memory. Speaking is more natural, more vivid, and more direct than writing or reading, and these advantages are worth more than the additional effort required for thorough preparation. The suggestions which follow are intended to give you practical hints on abbreviating and simplifying your material, and preparing an oral presentation of it.

In the first place, you must beware of the temptation to include everything you have to say in your oral version. Speaking is much slower than silent reading. In print, an author can sometimes afford to express himself exhaustively. In an oral presentation, neither the time limit nor the listeners' patience will permit anything but a selective treatment. Furthermore, an article, even in a technical journal, reaches a larger and presumably more general audience. In delivering a talk, the speaker must address himself to the group of individual listeners immediately confronting him, and he may touch lightly on matters which he would spell out carefully in writing even for technically informed readers.

In remembering the necessity for keeping within a time limit, you must remember also that the listener's capacity for absorbing detail is not so great as the reader's. The desire for full and complete treatment in a limited time may lead you into the mistake of using a compressed, compact style which is suitable only for reading and study with concentration. If you include too much detail for the sake of accuracy, you may find that you have sacrificed clarity. A graph shows more at a glance than the table of data from which it was constructed. A block diagram is more legible than a photograph of the apparatus.

The second suggestion is simplicity of structure. Clarity is largely the result of simplicity. The listener will go away from the meeting remembering what you have said if you have analyzed your material thoroughly and presented it in two or three or four (the fewer the better) clearly labeled main parts. If you can arrange your main points in a way that makes sense, label them plainly, and notify your listeners when you leave one and begin another, you make it easy for them to follow you. You also make it easier for them to remember what you have said, because you have provided pegs on which to hang the details.

Arrange your details under your main headings in orderly sequence. Common types of sequence are those of time, space, cause and effect, and special topical arrangements. Your main topics should follow one kind of sequence, but there is no rule against using different types under different main headings. Perhaps it should be unnecessary to add that you should not jump back and forth from one main topic to another in presenting your material. The more thorough and logical your analysis and arrangement, the less likely you are to skip around in delivering your talk.

Having set the limits of your talk and selected your main divisions and arranged your material tentatively in your mind or on paper, you are ready to consider your introduction and conclusion, and to begin to think about the actual words you plan to use in talking to your audience.

The function of an introduction is to get the attention of your listeners and to give them some reason for continuing to listen to you. If your name and your subject have appeared in the printed program of the meeting, a part of this function has been performed for you, and you may assume that at least some of the people present are there because they want to hear what you have to say. Nevertheless, it is important in preparing your talk for you to ask yourself this question: "Why should *these* people listen to *me* discuss *this* subject at *this* time?" Write out the answer to this question. It will focus your attention on the four important factors in every public-speaking situation: the audience, the speaker, the subject, and the occasion. If there is a good reason for your audience to listen to you, it will be found in one or more of these factors. Plan in your introductory remarks to tell them why they need to know what you have to say, and support your statement as you would any main point, illustrating the need by an incident or example, reinforcing your statement by additional examples or facts, and pointing out the direct relation of the subject to their professional interests. This part of your speech should be short (probably a tenth or less of its length), but it is important. Do not omit it.

The function of the conclusion is to summarize or recapitulate the main ideas of your talk, to draw attention to important conclusions, and to reinforce by restatement or application whatever impression you wish to leave in your listeners' minds.

A generalized skeleton outline of a report based on this plan might look something like the following:

Introduction

I. (After addressing the chair and the audience.) Opening statement designed to get attention (reference to your subject or the immediate circumstances of your talk; rhetorical question; startling statement or quotation; humorous story, *if apt*)
   - A. Support for statement
     - 1. Detail
     - 2. Detail
   - B. Further support
     - 1, 2, etc. (details)

II. Restatement

III. Statement telling why your listeners need your information
   - A. Support (illustration)
     - 1, 2, etc. (details)
   - B. Further support (additional facts or example)

IV. Statement relating subject directly to present audience
   - A, B, etc. (supporting statements)

V. Summary statement or restatement

VI. Preliminary summary of subject (enumeration of main parts or other clear indication of direction in which you intend to lead your listeners)

Body

I. Statement of first main part of subject
   - A. Support
     - 1. Detail
     - 2, 3, etc. (details)
   - B, C, etc. Further development of first main part

II, III, etc. Second, third, etc., main parts

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* Decimal classification: R040. Original manuscript received by the Institute, August 11, 1947.

† Brooklyn College, Brooklyn, N. Y.

[1] W. L. Everitt, "The presentation of technical developments before professional societies," PROC. I.R.E., vol. 33, pp. 423–425; July, 1945.

[2] Arthur C. Downes, "Proper presentation of papers before technical meetings," PROC. I.R.E., vol. 35, p. 235; March, 1947.

[3] J. M. O'Neill (editor), "Foundations of Speech," Prentice-Hall, Inc., New York, N. Y., 1941.

[4] Alan H. Monroe, "Principles of Speech (Brief Edition)," Scott, Foresman and Company, Chicago, Ill., 1945.

Conclusion

I. Final summary
   A. Recapitulation of main points
   B. Important conclusions
   C. Reinforcement (perhaps by referring again to listeners' need for your information).

When you have fitted your material to such an outline you may feel that the principal part of your work is finished, but there are two further steps in preparation which will add greatly to your own comfort and confidence on the platform, and, incidentally, to the comfort and edification of your listeners. The first of these is to clothe your outline with words, and the second is to practice delivering your talk aloud.

In choosing the actual words you will use in delivering your talk, you may have to experiment to find the method that is best suited to you. Some speakers prefer to write out a full draft so that they will not have to depend on the inspiration of the moment for striking, accurate, or thoughtful phraseology. Others, realizing that very few people can write as spontaneously and vigorously as they talk, combine the last two steps in preparation by taking the detailed outline to a private place where they can go through the material aloud, as if talking to a friend.

Whether you write it out first or not, practice aloud is the best way to fix the sequence of your ideas in your mind and to try out variations of wording. It is at this stage that you should time yourself and revise your plan accordingly.

The final stage of preparation for delivery is to put your outline or manuscript in your pocket and go through your talk without looking at the paper. Rehearse it, as you will deliver it, on your feet. If you have access to a recording device, use it; any office dictating machine will serve the purpose. In listening to the record you may learn something profitable about your vocal powers and limitations and about your distinctness of utterance.

The best kind of final rehearsal is a tryout performance before a group as much as possible like your ultimate audience. If you can arrange to talk about your subject in a seminar or in a conference of your colleagues, you will get a great deal of value from the experience.

These suggestions may be carried out in various ways. You may decide to prepare first the version of your paper destined for print, abbreviating it and simplifying it for oral delivery. Or, you may start with the oral version and rewrite it for publication, as Everitt suggests, so as to take advantage of the criticism and discussion precipitated by your presentation. Whichever you do, remember that, as Downes points out, "The presentation of a well-prepared abbreviated version will not in any way detract from the value to those in the audience more familiar with the general subject of the paper, since they can study it in detail on publication, and will hold the interest of the entire audience." And that "the papers most vividly remembered are those presented as though the author were talking to a few friends." Don't read before them—talk to them.

High-Power Ionosphere-Measuring Equipment*

P. G. SULZER†, ASSOCIATE, I.R.E.

Summary—**This paper describes a transmitting and receiving installation that provides high power output and excellent amplitude and timing resolution for ionosphere measurement. A wide range of operating frequencies, pulse widths, and pulse-repetition frequencies is available to permit the study of various ionosphere phenomena.**

Introduction

The pulse method of ionosphere investigation[1] uses a series of short-duration radio signals which are transmitted, reflected, and received, much as in radar. The pulses are longer, however, as are the distances and elapsed times. Frequencies in the ordinary communications range are used, since the purpose of the work is to investigate ionosphere characteristics at those frequencies.

Equipment Now in Use

Most of the equipment being used at present employs comparatively low power. This is particularly true of installations used in routine measurements for predicting communications conditions.[2] A peak power of $\frac{1}{2}$ kw. is typical.

For present experimental work, however, a transmitter of at least 50-kw. output is desirable. To utilize the capabilities of this power, a highly versatile receiving setup is necessary.

The New Equipment

The transmitting equipment described here has a frequency range from 1 to 16 Mc., with a peak power output of 100 kw. from 1 to 8 Mc., decreasing to 50 kw. at 16 Mc.

The receiving equipment is tunable from 0.54 to 20 Mc., and is capable of passing 100-microsecond pulses with little distortion. There are three values of full-scale range available: 200, 500, and 1500 km. Range markers are provided every 20 km. An attenuator is used to compare the amplitudes of received signals.

The installation is divided into two parts: the transmitter, and the receiving equipment. A photograph of

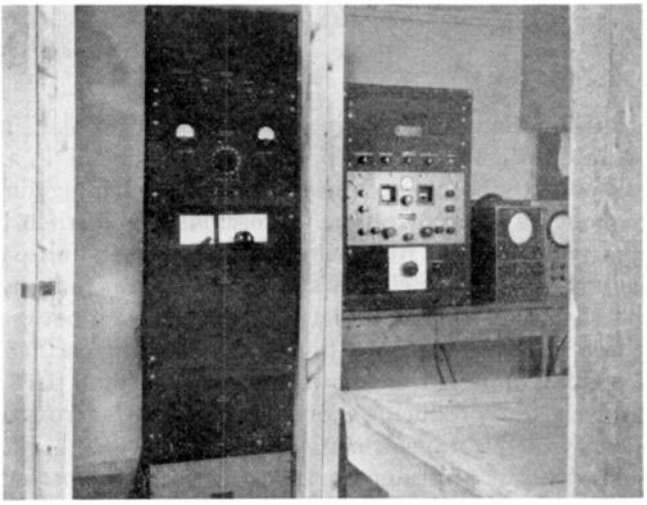

Fig. 1—Photograph of the complete equipment.

* Decimal classification: R248.13. Original manuscript received by the Institute, July 1, 1947.

† Pennsylvania State College, State College, Pa.

[1] G. Breit, and M. A. Tuve, "A Test for the existence of the conducting layer," *Phys. Rev.*, vol. 28, pp. 554–575; September, 1926.

[2] T. R. Gilliland, "Field equipment for ionospheric measurements," *Jour. Res. Nat. Bur. Stand.*, vol. 26, pp. 377–381; May, 1941.

the complete equipment appears in Fig. 1, the transmitter being mounted on the rack at the left. The receiving equipment is shown on the table at the right of Fig. 1.

Transmitter

The complete transmitter is mounted in a standard 6-foot relay rack, and is composed of the following units, mounted from top to bottom: pulse generator, high-voltage power supply, power oscillator, and pulse modulator.

Pulse Generator

The pulse generator develops positive pulses of 1000 volts amplitude that are used to drive the modulator. A schematic diagram of the unit appears in Fig. 2. The repetition frequency is variable from 15 to 60 per second; pulse durations from 20 to 200 microseconds are available.

The functions of this unit are controlled by a blocking oscillator, V_3. When S_2 is open, the cathodes of V_3 are returned to the 6.3-volt heater supply, effecting line-frequency synchronization; the oscillator may then be operated at the line frequency, 60 c.p.s., or submultiples thereof. When S_2 is closed, synchronization is removed, and the oscillator can be set to any frequency between 15 and 60 c.p.s. The operation of the blocking oscillator is normal, plate current flowing for only a small part of the cycle, and producing short negative pulses across the primary of the transformer T_4.

The output pulses of the blocking oscillator perform two functions: (1) They trigger the sweep and range marker generator in the receiving equipment (described later), and (2) they drive the trigger circuit V_4. This is a conventional single-sided flip-flop circuit, with section B normally cut off. During the active part of the cycle, B conducts, producing a negative pulse whose duration depends mainly on the time constant of C_8-R_{10}. By changing R_{10}, it is possible to vary the pulse duration from 20 to 200 microseconds.

The negative pulse is applied to the control grid of V_5, a limiting amplifier. In the absence of the pulse, this tube draws heavy plate current and, because of a high load resistor, has low plate voltage. Plate current is cut off by the negative pulse, and the plate voltage rises to the full power-supply value, 1000 volts.

The output is therefore a positive pulse of about 1000 volts amplitude. A direct-coupled cathode follower, V_6, is used to provide low output impedance for the unit.

Pulse Modulator

The pulse modulator provides a +10,000-volt pulse for the plate circuit of the power oscillator. The schematic diagram appears in Fig. 3. This unit consists essentially of a power amplifier using two type-5D21 tetrodes in parallel. When inactive, these tubes have sufficient control-grid bias for plate-current cutoff. Pulses from the previous unit drive the control grids positive, so that heavy plate current is obtained during the pulse interval. The plate voltage of the modulator tubes decreases to about 1000 volts when they are conducting, which means that a −9000 volt pulse is applied to the primary of T_7, a large pulse transformer. The purpose of this transformer is to reverse the polarity of the pulse and to provide, by means of taps, various output voltages up to 14,000.

The design of the transformer should be of some inter-

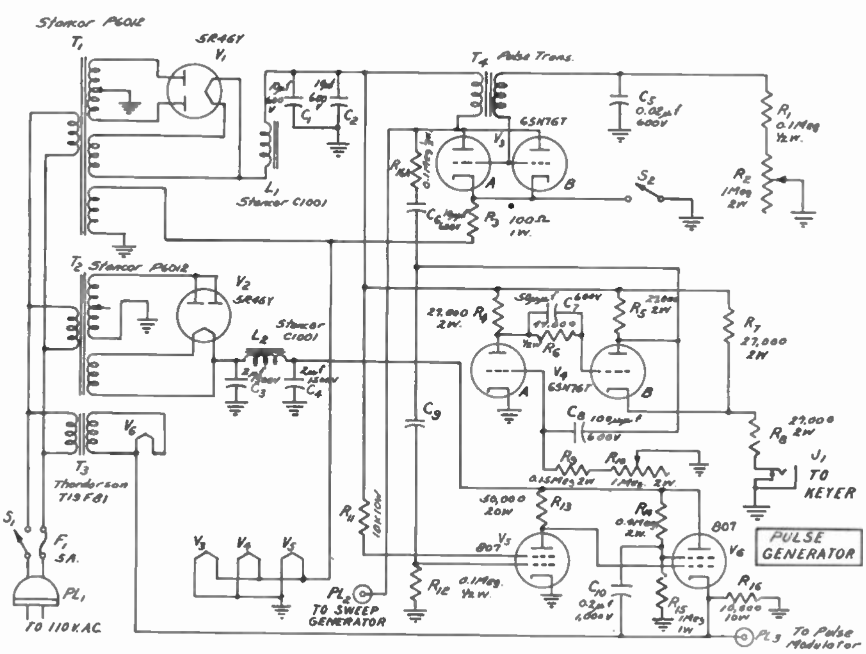

Fig. 2—Schematic diagram of the pulse generator.

est here, since these pulses are about 100 times as long as those used in radar. It was necessary to keep the magnetization current below 1 ampere to prevent excessive loading of the modulator tubes, which would have distorted the pulse shape. Flat-top pulses were required in this equipment to keep the frequency of the power oscillator constant during the pulse. When a rectangular pulse of amplitude E is applied to a series R-L circuit, which is the equivalent of the transformer primary, the initial rate of change of current, E/L, may be considered constant during the pulse interval t. The magnetization current is given by

$$Im = \frac{E}{L} t \quad \text{or} \quad L = \frac{E}{Im} t.$$

For the values used here, $E = 9000$ volts, $Im = 1.0$ ampere, $t = 100$ microseconds, and $L = 0.9$ henry. The transformer was designed for a primary inductance of 1 henry at 1.0 ampere d.c. Primary and secondary windings were split and interleaved to keep leakage inductance low.

Power Oscillator

The power oscillator converts the d.c. pulse from the modulator into r.f. energy. A schematic diagram of this device appears in Fig. 4. The high-frequency portion of the circuit is conventional, consisting of two 715B tetrodes connected in parallel as triodes. The plate circuit is tuned; feedback is provided by an untuned, inductively coupled grid coil.

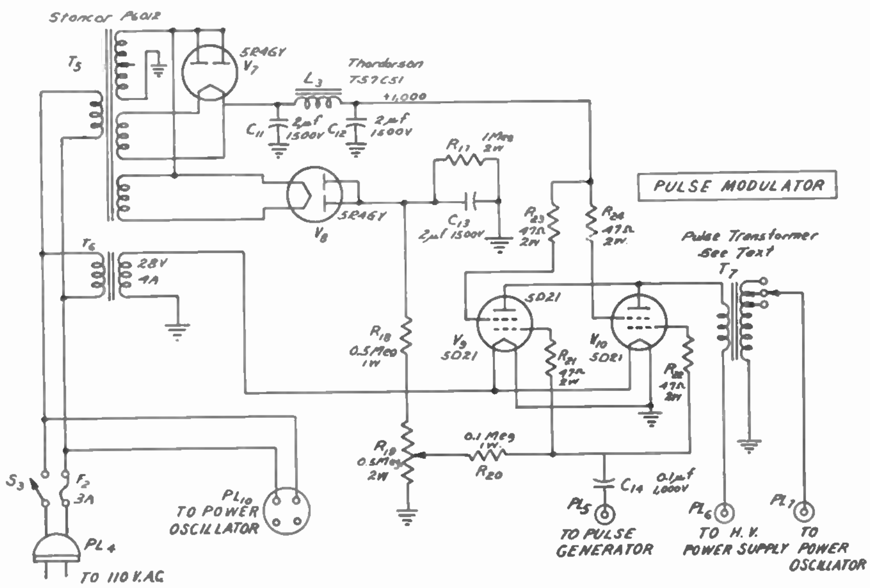

Fig. 3—Schematic diagram of the pulse modulator.

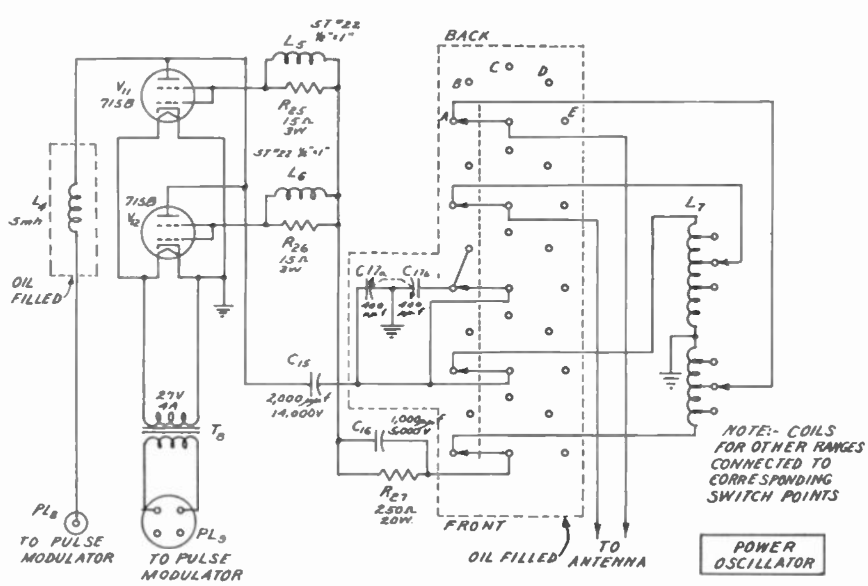

Fig. 4—Schematic diagram of the power oscillator.

The operating frequency is variable from 1 to 16 Mc. by means of the band switch S_8 and the tuning capacitor C_{17}. These two components are mounted in oil to prevent breakdown. If the tuning capacitor were

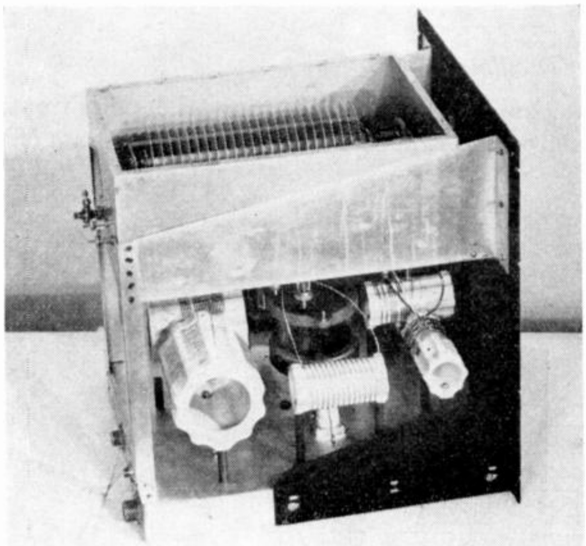

Fig. 5—Left-hand view of the power oscillator.

mounted in air, a spacing of more than 1 inch would be necessary; as it is, the spacing is only 0.2 inch. Leads were brought through the bottom of the oil tank by means of soldered metallized "pyrex" insulators. A commercial grade of mineral oil was found to be satisfactory, even at 16 Mc. The oil used is "Voltesso 36," to which was added 0.2 per cent "Paranox 441," an oxidation inhibitor. It was also necessary to mount the r.f. choke L_4 in oil to prevent sparking between sections. Left-hand and right-hand views of the power oscillator appear in Figs. 5 and 6, respectively.

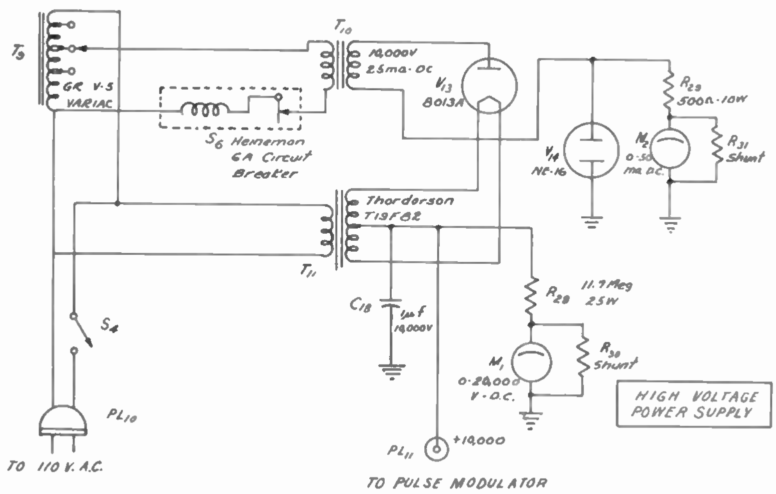

Fig. 7—Schematic diagram of high-voltage power supply.

High-Voltage Power Supply

The high-voltage power supply furnishes 10,000 volts d.c. at 25 ma. for the modulator. A schematic diagram appears in Fig. 7; it can be seen that the circuit is a half-wave rectifier using an 8013A diode. Since the load current is low, sufficient filtering is provided by a single 1-μfd. capacitor.

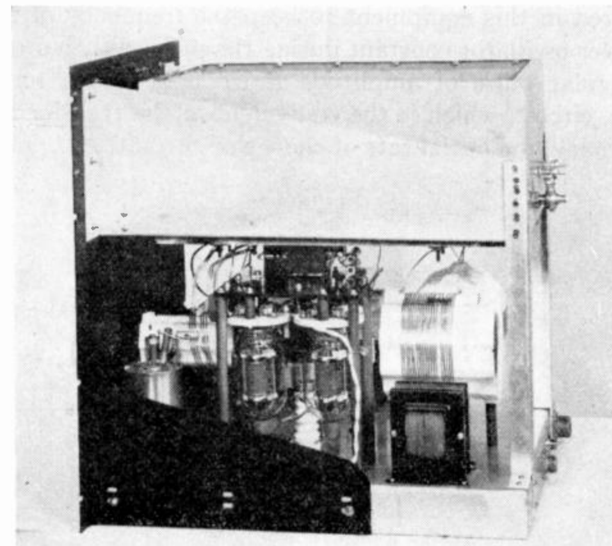

Fig. 6—Right-hand view of the power oscillator.

The d.c. milliammeter M_2 is protected by a neon tube V_{14}, in the event that excessive load current is drawn at low output voltages. A circuit-breaker, S_6, protects the high-voltage transformer T_{10}.

It was necessary to mount this transformer in oil to prevent corona, which caused excessive noise in the receiver.

Receiving Equipment

The receiving equipment is mounted in a table-top relay rack and consists of the following units, mounted from top to bottom: sweep and range marker generator, radio receiver, and attenuator unit. The indicator is a Dumont type-208 oscilloscope, which is mounted to the right of the small rack.

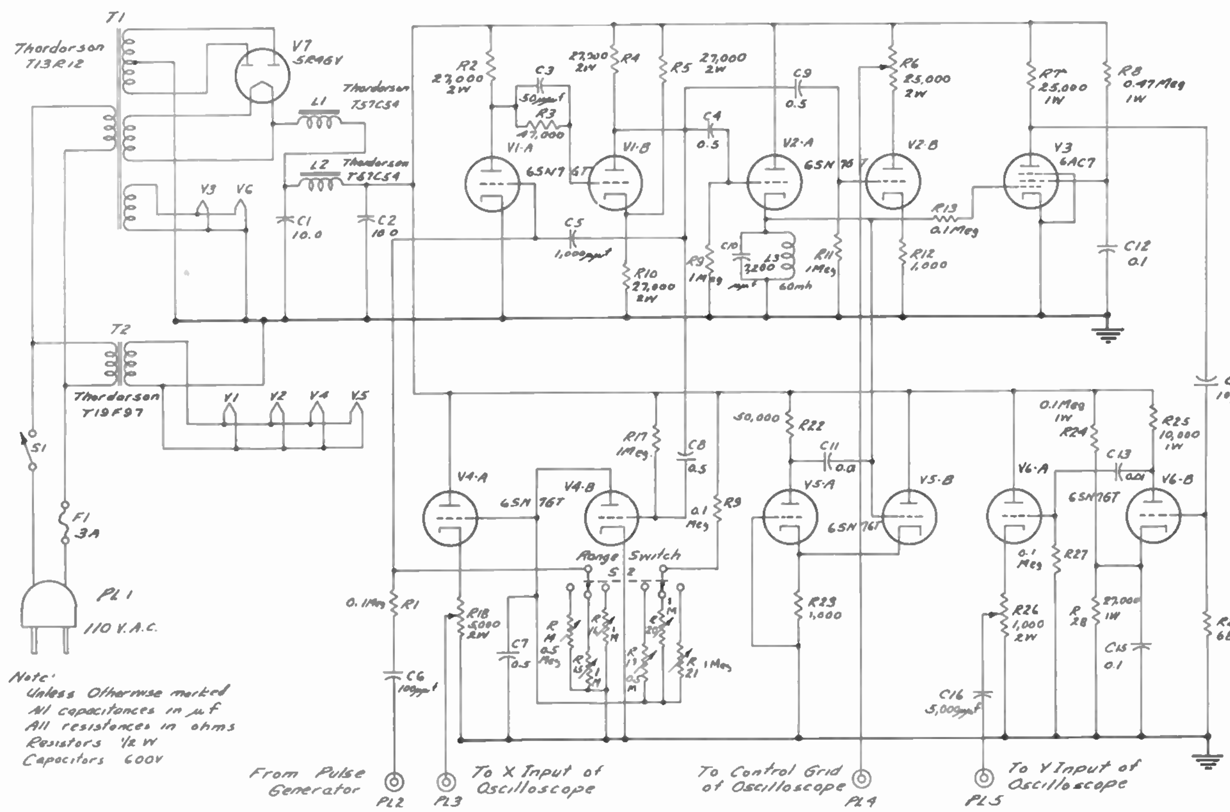

Fig. 8—Schematic diagram of the sweep and range-marker generator.

Sweep and Range-Marker Generator

The sweep and range-marker generator provides the sweep voltage, range-marker pulses, and intensifying pulses for the indicator. The schematic diagram of the unit appears in Fig. 8.

A negative sweep-initiating pulse from the transmitter pulse generator is applied to the grid of V_{1A}, which is the normally conducting tube of a single-sided flip-flop circuit. A negative rectangular pulse is obtained at the plate of V_{1B}. The leading edge of this pulse is coincident in time with the sweep-initiating pulse; its duration is determined by the setting of the range switch. This pulse drives three circuits: the range-marker generator, the sweep generator, and the intensifying-pulse generator.

The range-marker generator is controlled by an *L-C* circuit connected in the cathode of V_{2A}. This tube is cut off by the negative rectangular pulse, shocking the *L-C* circuit into oscillation. The amplitude of the oscillation is kept constant by the negative resistance appearing between the grid of V_{5B} and ground. This negative resistance shunts the tuned circuit and decreases the damping. The magnitude of this negative resistance is controlled by the resistor in the cathode of V_{5B}, which is adjusted for good frequency stability. Oscillations cease when V_{2A} is again allowed to conduct.

The oscillator frequency is adjusted to 7500 c.p.s. Voltage from the *L-C* circuit is applied to the grid of the limiter V_3, the output of which is differentiated by the *R-C* circuit in the grid of V_{6B}. This tube is biased to cutoff by the network connected between B+, cathode, and ground. Thus, the positive differential pulses are amplified by V_{6B} and impressed on the grid of V_{6A}, a cathode follower. The output of V_{6A}, which is applied to the Y axis of the oscilloscope, is comprised of a series of short-duration pulses which are positive in sign and separated by 1/7500 of a second. Assuming an average group velocity of 300,000 km./sec. for the r.f. signals being timed, the delay between marker pulses is equivalent to a distance of 40 km. or, for the vertical-incidence case being considered, an equivalent height of 20 km.

Sweep voltage is generated by allowing the capacitor connected between the plate of V_{4B} and ground to charge through a high resistance connected to B+. Charging starts when V_{4B} is cut off by the negative pulse from the flip-flop circuit. The capacitor is rapidly discharged when the pulse ceases. The resulting almost-linear increase in plate voltage of V_{4B} is applied through V_{4A}, a cathode follower, to the X axis of the oscilloscope. It should be noted that the start of the sweep coincides with the start of the transmitted pulse, since this coincides with the sweep-actuating pulse from the pulse generator.

The intensifying-pulse generator, V_{2B}, is cut off by the negative rectangular pulse from the flip-flop circuit, and the resulting positive pulse, applied to the control grid of the cathode-ray tube, is used to increase the intensity of the oscilloscope during the sweep.

Radio Receiver

The receiver is a Hammarlund Super-Pro which has been rendered suitable for pulse reception by the following modifications:

In the original receiver, the r.f. and first-detector grid returns were made through a.v.c. isolation networks, d.c. isolation from the tuned circuits being obtained by grid coupling capacitors. These *R-C* networks would cause blocking and loss of receiver sensitivity through rectification of the transmitted pulse. Consequently, the grid resistors were removed and the coupling capacitors were shorted out.

It was necessary to remove the a.v.c. and grid-bias control of the i.f. stages for the same reason. Cathode-bias networks were provided for these stages, as well as for the r.f. stages.

The i.f. pass characteristic was broadened to 20 kc. at 3 db down by placing suitable damping resistors across the transformers. This was required to more nearly preserve the pulse shape in the receiver. This is important in this equipment, since pulse amplitude is to be a measured quantity.

The a.v.c. second detector was modified to provide a low-impedance direct-coupled cathode-follower output.

Cathode bias gain control was incorporated in the r.f. and i.f. stages. This is provided by the attenuator unit.

Attenuator Unit

The attenuator unit, Fig. 9, is essentially a calibrated voltage divider which applies a positive potential to the

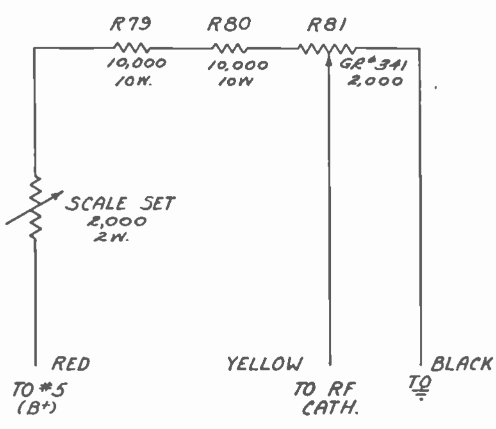

Fig. 9—Schematic diagram of the attenuator unit.

cathodes of the r.f. and first-i.f. amplifier tubes. These tubes have remote-cutoff characteristics, with the result that the applied cathode bias is nearly a logarithmic function of the receiver gain. Therefore, the attenuator scale, calibrated in decibels, is almost linear. This permits reading within 1 db up to the maximum attenuation, which is 100 db.

Results

Fig. 10 is a photograph of oscilloscope patterns showing the affect of power increase on *F*-layer reflections. Fig. 10(a) was taken with a peak power of 1 kw., while Fig. 10(b) was taken immediately afterward with 50 kw. The increase in the number of multiple reflections is readily apparent, as is the improved detail. These oscillograms were made on 6.425 Mc.

(a)

(b)

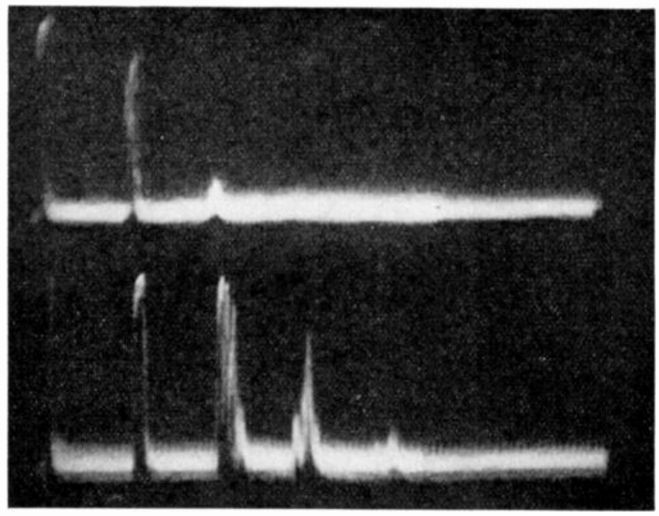

Fig. 10—*F*-layer reflections. (a) 1 kw. (b) 50 kw.

Other effects appear with high power. Fig. 11 shows *F*-layer reflections on the same frequency, plus a transient *E*-layer reflection. The latter-type reflections apparently result from patches of ionization that exist for only a few seconds in the *E* region.

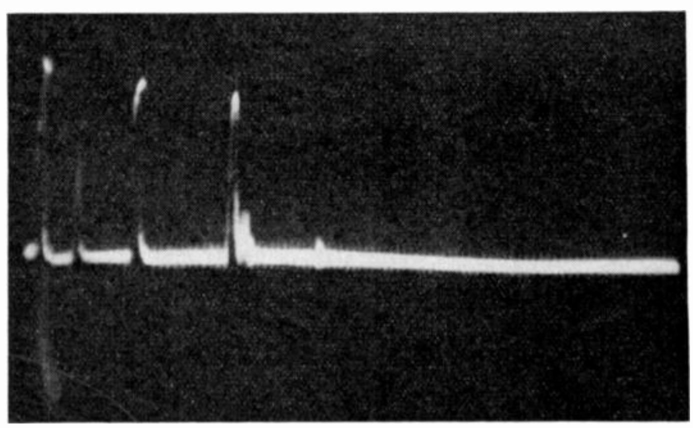

Fig. 11—*F*-layer reflections and transient *E*-layer reflections

Acknowledgment

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Home Projection Television

Part I. Cathode-Ray Tube and Optical System*

H. RINIA†, J. DE GIER†, AND P. M. VAN ALPHEN†

Summary—**A comparison of direct-viewing with projection tubes leads to the conclusion that the latter are better suited to provide a picture of adequate dimensions. The characteristics of a 2.5-inch cathode-ray tube for television projection in an average living-room are described. Some details of the tube are as follows: a very small spot size, achieved by unusually close tolerances in neck and gun dimensions; a narrow neck, reducing the energy required for the magnetic focusing and deflection to a value about equal to that for direct-viewing tubes; a face plate ground to meet the optical requirements of the projection system; and a metal backing for the screen so that high reflectivity and good electrical conductivity aid in the achievement of an adequate brightness.**

For projecting the image on the viewing screen, a modified Schmidt-mirror system was adopted. The different possible modifications and their advantages and disadvantages are discussed. A simple and flexible method for preparing aspherical correction plates is described; it consists in molding the correctors on a glass plate from a gelatine solution and drying them afterwards. The design and performance of the projection system are discussed briefly.

Introduction

The television receivers that appeared on the market during the years 1936 to 1940 were almost exclusively of the direct-viewing type. The dimensions of the picture were relatively small, since practical considerations limit the size of the tubes. The larger sizes of direct-viewing tubes are expensive, heavy, and awkward to manipulate. The machines necessary for sealing in, exhausting, and other processing are bulky; the space necessary for storage of envelopes and finished tubes is large, and the transportation costs are also high. With projected pictures, however, the tube can be much smaller, and machines for handling the production of the tubes need not be much larger than those in use for receiving tubes. The consumption of material also is markedly lower. Therefore, the cost of a projection tube suitable for attaining picture dimensions in the range of 12×16 inches is much lower than that of a direct-viewing tube for the same picture size. This not only tends to reduce the initial price of a projection set, but also reduces the tube replacement cost to only a fraction of that for a direct-viewing set with comparable picture dimensions.

The first Philips projection television tube, completed in 1937, has been described elsewhere.[1] Extensive development in several directions has taken place since that time.

In this paper a new 2.5-inch projection tube is described. The maximum dimensions of the picture on the face of the tube are 1.42×1.89 inches, the aspect ratio being 3:4. In order to project a large and sharp picture, the optical system used with such a tube must be characterized by high optical efficiency and good resolution. A modified Schmidt-mirror system, utilizing an aspherical correction plate produced in a simple manner from ordinary gelatine, can fulfil these requirements. In this way a bright, sharp, flat television image of 12×16 inches has been successfully produced. The small tube makes it possible to use a mirror having a focal length of only 4 inches, with the result that the optical components and the entire projection path can easily be contained in a medium-sized cabinet.

I. The Projection Television Tube

1. Theoretical Considerations

It must be realized that, in order to resolve the same number of lines, the diameter of the luminous spot on a 2.5-inch tube must be only one-fourth as large as for a 10-inch tube. With the maximum available picture dimensions of 1.42×1.89 inches, the maximum tolerable spot size for 525 scanning lines amounts to 1.42/525 =0.0027 inch, or 0.0068 centimeter. A rather rough indication of the size of the spot that can be achieved with the tube operated under the conditions given below can be obtained from a consideration of the two most important factors determining that size:

(1) The effect of cathode-emission-current density, which is necessarily finite, and in which the emitted electrons have a thermal distribution of velocity.

(2) The effect resulting from the astigmatism and the curvature of the image field caused by the deflection system.

Owing to the first factor alone, without limiting apertures in the beam, the spot for the so-called "ideal gun" would have a half-value width[2]

$$d_1 = \frac{2}{\alpha}\left[\frac{\Phi_T}{\Phi}\,\frac{I}{\pi j_0}\ln 2\right]^{1/2}$$

where α is the semiangle subtended at the screen by the cross section of the beam at the focusing coil (the exit "aperture angle" of the beam); $\Phi_T = kT/e$ where T is the absolute temperature of the cathode, k is Boltzmann's constant, and e is the charge of the electron; Φ is the potential of the screen with respect to the cathode; I is the beam current; and j_0 is the average current density taken over the effective cathode area.

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† N. V. Philips' Gloeilampenfabrieken, Eindhoven, the Netherlands.

[1] M. Wolf, "The enlarged projection of television pictures," *Philips Tech. Rev.*, vol. 2, pp. 249–253; August, 1937.

[2] G. A. Morton, "Electron guns for television application," *Rev. Mod. Phys.*, vol. 18, pp. 362–378; July, 1946.

With regard to the spot distortion caused by the deflection system, it can be shown[3] that it is possible to design deflection coils in such a way that the astigmatism disappears. However, the curvature of the image field cannot be influenced to any practical extent and will always remain. The resulting increase of the diameter of the spot will be dependent on the position on the screen. Under the most favorable focusing conditions the maximum value d_2 of this increase occurring in the center and at the edges of the screen is given by the relation:

$$d_2 = \frac{1}{2}\,\alpha y^2\left(\frac{1}{l}+\frac{1}{L}-\frac{1}{\rho}\right)$$

where y is the distance from the edge to the center of the picture on the screen, l is the length of the deflection coils in the direction of the beam, L is the distance between the center of deflection and the screen, and ρ is the radius of the curvature of the screen (the sign of ρ being positive when the center of curvature and the center of deflection are at the same side of the screen). The best compromise is to design the tube in such a way that α satisfies the equation $d_1=d_2$. With $\Phi_T=0.1$ volt, $\Phi=25{,}000$ volts, $I=100$ microamperes $=10^{-4}$ amperes, $j_0=0.2$ amperes per square centimeter, $y=2$ centimeters, $l=5$ centimeters, $L=9.5$ centimeters, and $\rho=10$ centimeters, one finds that, for $\alpha=0.01$ radian, $d_1=d_2=0.0041$ centimeter.

It is clear, therefore, that even at the edges of the screen, where the spot diameter will be $d_1\sqrt{2}=0.0058$ centimeter, the lines are still separated.

It can be shown that other factors, such as space-charge repulsion in the beam, spherical aberration of the focusing lens, and the remaining errors caused by the deflection system, will, in our case, play a comparatively unimportant role.

By substituting in the above formulas data obtained from a satisfactory commercial 10-inch direct-viewing tube, one finds a spot half-value width that is roughly four times larger than for this tube, provided that α was chosen equally advantageously. These considerations therefore suggest that the 2.5-inch projection tube can be made to give adequate resolution for 525 lines, and it does so in practice.

2. Construction of the Tube

Fig. 1 gives an outline drawing of the tube. The region of the neck around which the coils are placed has a maximum outside diameter of 21.5 millimeters. This small diameter was chosen in order to keep the required deflection and focusing energy low. The diameter of the neck at the electron gun is somewhat smaller still, and this portion is treated to obtain a very accurate internal diameter. The tolerance for this dimension is only a few microns, while that for the rest of the neck may amount to 1 millimeter. Careful finishing of the gun components for roundness and freedom from burr, for example, is also required. The high-voltage part of the gun is inserted in the accurately calibrated part of the neck. The low-voltage part, which is mounted on a glass stem, also fits very accurately in the calibrated neck, ensuring good centering of the electron beam.

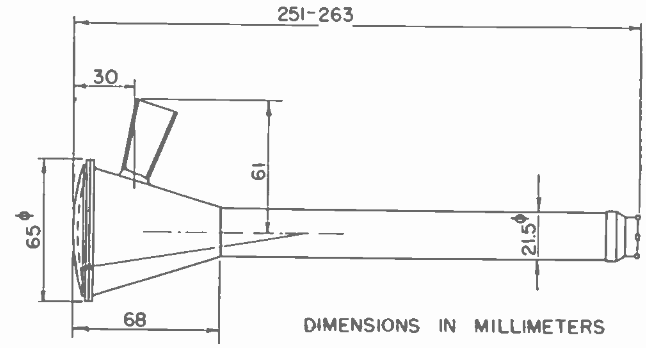

Fig. 1—Outline drawing of the tube, showing dimensions in millimeters.

The high-voltage anode terminal consists of a button in a glass cup that is sealed on the cone. It is connected with the high-voltage electrode in the neck through an aquadag layer inside the tube. The glass cup serves to lengthen the external leakage path from the high-voltage contact to the coils, for example, which as a rule are grounded. Moreover, the outside of the cone and part of the neck are covered with a conductive coating that can be grounded. This outer coating, together with the conductive coating inside the tube, forms a capacitor of some 300 μμfd. capacitance that is used for the final smoothing of the high voltage applied to the tube.

As the tube is intended for use in a Schmidt-mirror system having a wide aperture and consequently a very limited optical depth of focus, the curvature of the face must be accurately defined. For this reason the face plate is carefully ground and polished to the required radius before being sealed in place.

3. Electrical and Phototechnical Data

Fig. 2 shows the characteristic $I_s=f(V_g)$ of the tube. The accelerating voltage is 25 kv. The average current under representative operating conditions may be 60 to 100 μa.

Focusing and deflection are both magnetic. The number of ampere turns needed for focusing, when using a shielded coil with an 11-millimeter air gap, is approximately 600. With a distance of 96 millimeters between the center of deflection and the screen, the deflection for a 25-kv. beam is given by $0.018\,lH$ centimeter, where l is the length of the coil in centimeters and H is the field strength in oersteds. With a coil 5 centimeters long, a deflection of 2.5 centimeters in one direction is obtained in practice at a field strength of 30 oersteds.

[3] To be published by J. Haantjes in *Philips Research Reports.* This paper will also include a derivation of the relation given below for d_1.

The exact shade of white that is preferred for the light emitted from the fluorescent screen is found to be a fairly subjective matter. However, since a somewhat bluish white is usually preferred, the phosphors for the tube were so chosen that the color temperature is in the neighborhood of 6500°K.

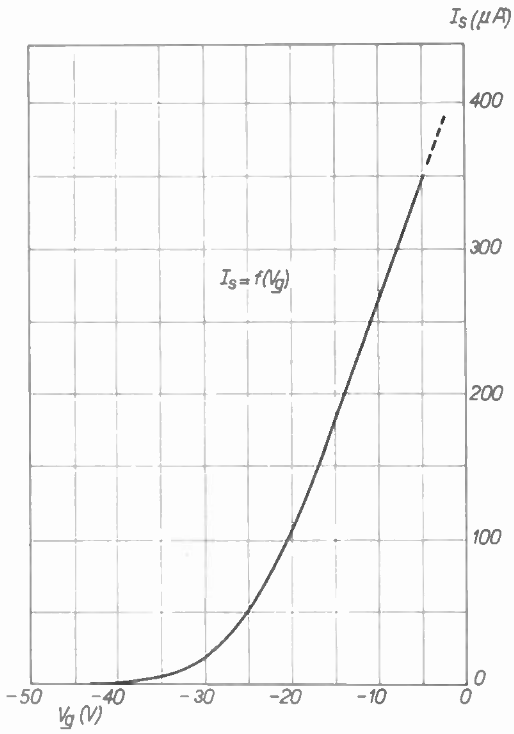

Fig. 2—Beam-current characteristic.

4. *Metal Backing of the Screen*

The backing of the fluorescent screen by an extremely thin layer of conductive and highly reflecting metal has led to significant improvement in some of the performance characteristics of high-voltage, high-intensity cathode-ray tubes. References to such a layer and the advantages derivable from its use have appeared in the patent literature for a long time. A thin, opaque, highly reflecting metal layer not only increases the useful luminous output by adding to the light emitted in the forward direction a considerable portion of the light emitted in a backward direction, but also improves the over-all contrast, as internal reflections and irradiations are no longer transmitted through the screen.

Another advantage of the metal backing is that the good electric conductivity of the layer ensures more stable operation of the screen. Without such a layer, phosphors that were known to yield high efficiencies in the 5- to 10-kv. range could not be used in high-voltage tubes because poor secondary emission allowed the screen potential to drop far below the anode voltage. However, the metal backing serves as a high-tension lead so close to the phosphor grains that the charge is easily conducted away. For this reason such phosphors show, with metal backing, a far greater gain in efficiency than is due to the optical reflectivity alone. The metal backing is also virtually impervious to the massive negative ions which may be present in the beam, and thus effectively performs the function of an ion trap.

If one attempts to deposit a layer by evaporation upon the bare grains, its thickness will be very irregular and the conducting and reflecting qualities will be poor. Moreover, the particles of metal can pass freely through interstices between the grains and be deposited on the glass itself, causing absorption and backward reflection. Various technologists have therefore been working independently toward improved methods for depositing the layer. One method consists in filling up the spaces in the grainy phosphor layer with a suitable inert material, using only enough of it to obtain a continuous, smooth surface covering the grains just to their tips. Another method is the use of a thin membrane stretched over the tops of the grains. Upon the smooth surface so obtained a thin layer of metal can then be evaporated. The filler or membrane can afterwards be removed either by combustion or by evaporation.

Aluminum is the metal generally used, because of its high optical reflectivity, and because its low atomic weight permits easy penetration by the electron beam. The evaporation of the aluminum is a simple procedure. A layer 0.15 to 0.50 microns thick is opaque to light and absorbs only a small portion of the electron energy, as the depth of penetration of 25-kv. electrons in aluminum is about 15 microns. Fig. 3 shows an example of the light gain obtained. Aluminum has the advantage that its oxide is colorless, so that the reflectivity of the layer is not affected during the firing of the membrane or the heating of the tube during the evacuation process. It is believed that the oxide also tends to reduce the possibility of evaporation of the metal under intense electron bombardment.

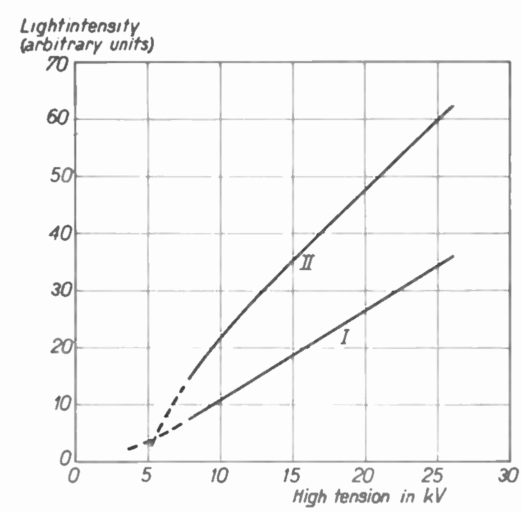

Fig. 3—Comparison between the light output of a metal-backed and an unbacked projection-tube screen. (I) unbacked; (II) metal-backed.

5. *Discoloration by X rays*

After long use at an accelerating voltage of about 15 kv. or more, an area of discoloration in the form of the scanning frame appears in glass of the type ordinarily used for tube faces. This phenomenon was found to result from the action of soft X rays generated by the electron bombardment of the phosphor. Some discoloration also occurs in the glass cone, as the X rays are scattered in all directions. By examining a cross section cut from the tube face, one can see that the density of the color centers in the glass decreases in the direction away from the phosphor side. The color centers thus formed will in time cause a 5 to 10 per cent light absorption. The color of the transmitted fluorescent light also changes somewhat, as the absorption is not uniform throughout the spectrum.

Favorable results were obtained with a tube face of a special glass that can be sealed to the cone, and with which discoloration does not occur to an appreciable degree.

II. The Projection System

1. *Modification of the Schmidt System*

For television projection either lenses or mirrors can be used, but concave mirrors offer superior advantages.[4] The Schmidt-mirror system gives especially good results because of its large numerical aperture.

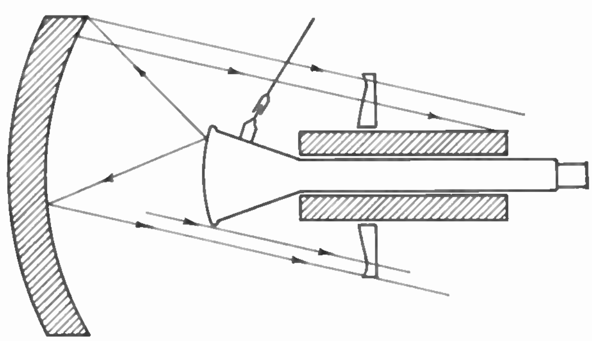

Fig. 4—The Schmidt system with perforated correction plate. All the accessories of the cathode-ray tube lie in the light path. The neck of the tube intercepts some of the light from the edges of the picture.

The Schmidt system in its original form cannot be used with the finite throw distance required for projection television. Another problem arises because the curved fluorescent screen of the cathode-ray tube must be introduced between the correction plate and the mirror, while the tube is ordinarily longer than the available space between these elements. One very obvious solution is to make a hole in the center of the correction plate and let the neck of the tube extend through it.[5,6] However, this solution is often not acceptable, because not only the whole tube, but also all the wiring, connections, supports, and coils are in the optical path and intercept a portion of the light. It is especially unacceptable for a small tube such as the one described here, as it merits consideration only when the tube face is larger than the cross section of the coils and other accessories. Even then it suffers from the disadvantage that the beams from the edges and corners of the picture are inclined with respect to the tube, and thus may be partially intercepted by the focusing coils and the tube neck (Fig. 4). This effect is diminished by making the slope of these beams small, i.e., by giving the optical system a longer focal distance for a given fluorescent screen diameter; however, to maintain a given optical speed, this in turn necessitates a mirror of larger diameter. This solution therefore leads toward large, awkward mirror systems.

Another possible solution is to make a hole in the concave mirror, to introduce the face of the tube through this hole, and to place a plane mirror between the concave mirror and the correction plate. All the connections, coils, and supports for the tube are now outside the light path, where ample space is available for them. But a difficulty remains in choosing the dimensions of the plane mirror. If it is made too small, the middle of the tube face radiates no light toward the edges of the spherical mirror and the correction plate. On the other hand, if it is made too large, much of the light reflected by the concave mirror is cut off. It is obvious that the dimensions are restricted within rather narrow limits by these conditions. Additional difficulties are encountered when such considerations are applied to the edges and corners of the picture (see Fig. 5). The cone of light required to utilize the full aperture of the Schmidt optical system in projecting the images of these areas is not provided by a small plane mirror, and the difficult choice is presented of either making the plane mirror larger and cutting off more light from the center of the picture, or allowing the brilliance of the picture to decrease rapidly toward the edges. The latter phenomenon is called vignetting, or window shut-off. Only when the requirements for uniform illumination of the picture are not made too severe can this arrangement of the optical system be considered. In addition, this method increases the space required for the projection system.

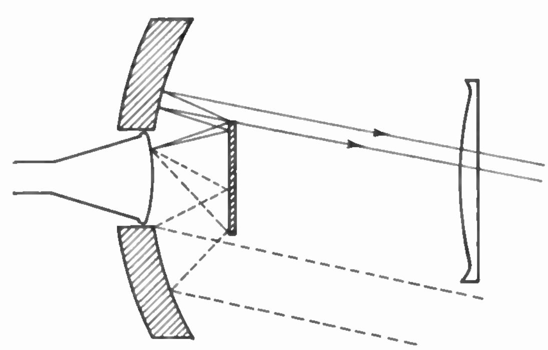

Fig. 5—Mirror system with perforated concave mirror and auxiliary plane mirror. Vignetting occurs because only narrow cones of light are transmitted from the edges of the picture.

[4] K. Pestrecov, "Television optics," *Electronic Ind.*, vol. 4, pp. 80–83; August, 1945.

[5] J. G. Maloff and D. W. Epstein, "Reflective optics in projection television," *Proc. Nat. Electronics Conf.*, pp. 190–206; Chicago, Ill., October, 1944.

[6] J. G. Maloff and D. W. Epstein, "Reflective optics in projection television," *Electronics*, vol. 17, pp. 98–105; December, 1944.

The arrangement we have utilized also employs a plane mirror.[7] However, it is not situated between the tube and the concave mirror, but between the concave mirror and the correction plate (see Fig. 6); and it is placed in an oblique position so that the Schmidt system is "folded," and occupies only half the space of the conventional arrangement. In the plane mirror is a hole only large enough to permit the tube face to be inserted through it. Behind the mirror there is ample room for coils, connections, and supports for the tube. Due to the fact that the hole in the mirror practically coincides with the tube face, vignetting of the edges of the picture does not occur. The light loss due to the hole in the mirror and to interception by the tube face is practically the same for the beams from the edges and corners as for the beams from the center of the picture. There is no interference from the coils and the neck of the tube, since these are behind the plane mirror.

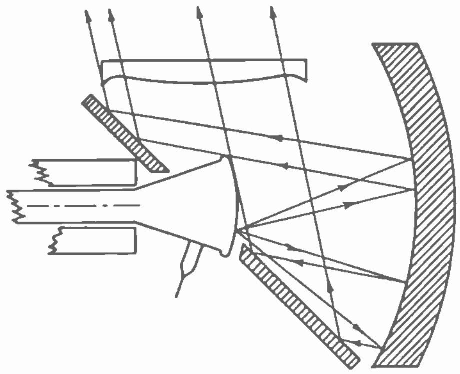

Fig. 6—Mirror system with 45-degree plane mirror. Wide-angle beams are transmitted even for the edges of the picture.

Because the light path in the Schmidt system is folded it is possible to mount the projection tube with the optical system in a small space, and to enclose the whole in a dust-proof box. All the mirrors then remain clean, as only the outside of the correction plate is exposed to dust; but, since this is a glass plate, its cleaning offers no difficulty. Furthermore, this plate can be rigidly fixed in position so that there is no danger of disturbing the adjustment of the optical system when it is touched.

2. *Preparation of the Aspherical Correction Plate*

The practicability of utilizing the Schmidt optical system in projection television receivers depends upon whether or not it is possible to make accurate correction plates by a simple and inexpensive process.

It is possible to counteract the spherical aberration of a concave mirror to a considerable extent by means of a glass compensator which has spherical rather than aspherical surfaces, but which must be rather thick.[8] A corrector of this type can be accurately made from optical glass by familiar methods, but due to the great thickness it becomes heavy and expensive. Furthermore, in order to avoid any strong chromatic aberration, the correction system must consist of two parts, each made of a different kind of glass. Also, the highest speed of the optical system is not easily obtained with this type of corrector.

A more practicable type of correction plate results when an aspherical contour of the surface is used. The minimum thickness of the plate and the minimum slope of the surface are obtained when the correction for spherical aberration is combined with a plano-convex lens. The contour of the plate must be computed to provide optimum performance at the projection distance desired for the modified Schmidt system. A possible way of making such an aspherical plate is by molding it from transparent plastics. A mold of the desired form is made, and some plastic such as polystyrene or perspex is pressed in it. The mold must be very accurate in shape, and must have a surface of optical quality. Moreover, it must be resistant to the required pressing and heating treatments, so the choice of material from which it can be made is limited. The mold is usually made and polished entirely by hand, and is given the correct final shape by local retouching. A separate mold is required for each shape of correction plate that is to be made.

We have followed an entirely different line.[9] The starting point was the observation that a gelatine gel retains its initial smoothness of surface after drying. This is contrary to the behavior of most other substances, which upon drying generally take on a more or less wrinkled surface.

To utilize this principle in making aspherical surfaces, the procedure is as follows: A mold is turned on a precision lathe, the shape of the surface being made the negative of that of the aspherical correction plate. In radial directions the mold has the same dimensions as the correction plate, but the variations in depth of contour are exaggerated by some chosen factor; for in-

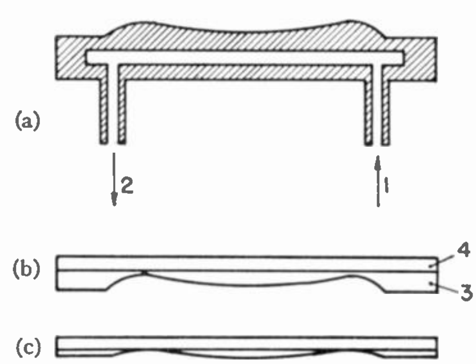

Fig. 7—Preparation of aspherical correction plates. (a) The mold with inlet *1* and outlet *2* for warm or cool water. (b) Glass plate *4* with the wet gelatine gel *3*. (c) The same plate as (b) after drying.

[7] French Patent No. 875,672, June, 1942.
[8] N. V. Philips, Dutch Patent No. 54,918, October, 1940.

[9] H. Rinia and P. M. van Alphen, "A new method of producing aspherical optical surfaces," *Proc. Kon. Ned. Akad. v. Wetenschappen*, vol. 49, pp. 146–149; 1946; Dutch Patent No. 57,677, October, 1939.

stance, ten times. If the actual correction plate has a total variation in thickness of 0.5 millimeter, this will become 5 millimeters in the mold (Fig. 7). The mold is gently heated, a 10 per cent solution of gelatine in water is poured on it, and a glass plate is then placed over it. The excess gelatine is pressed out and a thin layer remains between the glass plate and the mold. The whole is then cooled until the gelatine solution has solidified, and the glass plate is raised from the mold. Owing to the strong adhesion of gelatine to glass, the gelatine gel is loosened from the mold and sticks to the glass. The plate is then dried. The water evaporates and the layer of gelatine shrinks, but due to its adhesion to the glass the shrinkage can take place only in the direction of its thickness (just as with a photographic plate). Thus the glass plate is left with only a thin layer of gelatine, the shape of whose surface is a tenfold reduction of that of the mold; that is, it has exactly the required shape of the aspherical plate.

This method offers a number of advantages, as follows:

(1) Due to the fact that the final surface is a tenfold reduction of the shape of the mold, the mold itself need not be so accurately finished.

(2) The mold needs to be only moderately heated and cooled. There is no pressing at all, and distortion does not occur.

(3) The correction plate consists mostly of glass, so the possibility of distortion subsequent to molding is negligible. But the gelatine is also very hard and resistant to scratches, and may safely be cleaned with a soft cloth. The completed plate very much resembles an unexposed, fixed lantern slide.

(4) By means of a single mold, correction plates of different shapes can be made. By merely varying the concentration of the gelatine solution, correction plates are obtained which have different "optical power." In this way it is very easy to make correction plates for different projection distances with the same mold, and to correct for changes in thickness of the glass used for the tube face.

3. *Details of Design*

The correct adjustment of the optical system requires placing the center of the correction plate exactly at the center of curvature of the concave mirror. To facilitate this adjustment, the center of the plate is indicated by the point of a V-shaped mark impressed by the mold. The point of the inverted image reflected by the concave mirror must be brought into coincidence with the point on the plate so that the two reversed V's form a cross. Using this criterion, the mirror is brought to the position at which its center of curvature is just in the center of the correction plate.

The dimensions of the television receiver cabinet depend on the choice of focal length of the optical system. In a Schmidt-mirror system the focal length is of the same order of magnitude as the diameter of the correction plate. But the correction plate must have a larger diameter than the fluorescent screen of the tube, since otherwise no reflected light could pass through this plate. It is reasonable to assume that the diameter of the correction plate must be at least twice the diameter of the tube face.

The small projection tube, with a screen diameter of only 2.5 inches, makes it possible to use a focal length of 4 inches. The distance between the correction plate and the viewing screen is then only 30 inches for a 12×16-inch picture. This optical path length can be easily included in a cabinet of medium size containing only one auxiliary plane mirror.

The components of the projection television optical system described here are shown in Fig. 8. The system

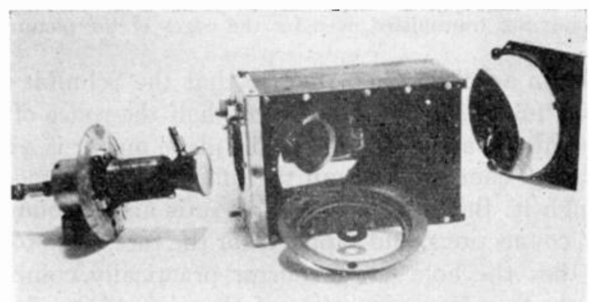

Fig. 8—The optical components and the cathode-ray tube.

has a numerical aperture (the sine of the semiapex angle of the cone of gathered light) of 0.64. The optical efficiency of the mirror is the square of the numerical aperture; in this case, 41 per cent. Masking, absorption, and reflection losses reduce the over-all efficiency to approximately one-half of this value. In an actual television receiver cabinet one additional plane mirror is ordinarily used, thus reducing the over-all optical efficiency to approximately 17 per cent. On actual pictures in such a receiver utilizing a directional viewing screen with an "amplification factor" of approximately four, measured values of highlight brightness lie between 15 and 20 footlamberts. Assuming an over-all optical efficiency of 17 per cent in the complete television receiver, a linear magnification of 8.5 times, and a viewing screen "amplification factor" of four, highlight brightness values of 15 and 20 footlamberts at the viewing screen correspond to tube-face brightness values of 1600 and 2100 footlamberts, respectively. Additional effective viewing screen brightness may be obtained by the use of screens with different directivity characteristics.

The sharpness of the images produced by this optical system with a gelatine correction plate is extremely good, and it can easily render a definition of 525 lines. The durability of the gelatine plate is comparable with that of a photographic negative or a lantern slide.

Part II. Pulse-Type High-Voltage Supply*

G. J. SIEZEN† AND F. KERKHOF†

Summary—**The performance of rectifier circuits of the voltage-multiplier type, energized by voltage pulses occurring in an inductive load in the plate circuit of a sawtooth-driven, biased beam-power tube, is briefly analyzed, and formulas are given for the internal resistance of circuits of this type comprising any number of stages. A method for substantially reducing the internal resistance by means of automatic bias control of the driver tube is described. Various factors determining the optimum number of rectifier stages for a given output voltage are discussed. An exceedingly compact high-voltage supply having automatic voltage control and furnishing 25 kilovolts for projection-type television tubes is described. This unit employs a voltage-tripling circuit with miniature rectifier tubes, the cathodes of which are indirectly heated by pulsed energy. A newly developed low-loss magnetic ferrite material has been successfully applied in the high-voltage supply.**

Introduction

THE DEVELOPMENT of low-cost television circuits is characterized by two definite trends. The first is a reduction of the physical size and is universally followed. The second, less obvious but equally important, is a reduction of the d.c. power consumption of the unit concerned. Some reduction usually occurs as a direct consequence of any substantial size reduction; further reduction is sought due to the fact that d.c. power, cheap as it may be so long as the total demand of the receiver can be met by one standard-type power supply, becomes expensive as soon as an increase of the demands necessitates the use of an additional power supply, as usually is the case with television receivers.

Applying the above considerations to the high-voltage supply unit, it soon becomes evident that a very special technique is needed if the voltage required exceeds a few kilovolts. This is caused by the fact that the output power required generally does not exceed a few watts, although the voltages may be in the order of 10 kv. for direct-viewing tubes and 25 kv. for projection tubes.

Experience has shown that conventional power supplies, operated from the a.c. line voltage, are bulky and expensive because of the fact that their size and weight becomes a function of the output voltage, rather than of the power; furthermore, they are dangerous, since they can furnish currents considerably in excess of the normal tube requirements of a few hundred microamperes.

The solution obviously must be sought in the direction of higher operating frequencies for the high-voltage unit, and various circuits in which this objective is pursued have been proposed.

One of these, known as the fly-back type, generates the high voltage as a by-product of horizontal scanning. Although this would appear to be the most economical method, it has, at least at present, the serious disadvantages that the output voltage is dependent on the horizontal sweep amplitude, which causes undesirable effects when the horizontal synchronization is lost; and that the fly-back time is increased by the additional loading.

In another system, known as the r.f. type, the rectifier circuit is energized by a separate high-frequency oscillator through a specially designed high-quality band-pass filter.[1,2] In this case, operating frequencies ranging from 300 to 1200 kc. are used, and it is claimed that, with a suitable setting, the internal resistance can be kept sufficiently small. However, the size and power efficiency still leave much to be desired, and it is difficult to shield other parts of the television receiver from the nonsynchronous r.f. interference produced by this kind of high-voltage supply.

This paper deals in some detail with a third type of high-voltage supply, known as the pulse type. It differs from the fly-back type in that a separate pulse generator is employed, operating at a frequency which is considerably lower than the horizontal sweep frequency. It is shown that, on account of the greater flexibility with regard to the choice of operating frequency, this method presents some marked advantages which lead to remarkable compactness, low cost, and good power efficiency.

1. Operating Principle and Advantages

The operation of the pulse-type high-voltage supply is based on the periodic interruptions of a current i_m through an inductance L. Assuming that this inductance is shunted by a stray capacitance C_p, the peak voltage V_m of the transient oscillation caused by the interruption can be deduced from the energy equation

$$\tfrac{1}{2}Li_m^2 = \tfrac{1}{2}C_pV_m^2$$

which expresses that the energy stored in L at the moment of interruption equals the energy stored in C_p during the first peak of the ensuing transient oscillations. This yields

$$V_m = i_m\sqrt{\frac{L}{C_p}}\,. \tag{1}$$

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† N. V. Philips' Gloeilampenfabrieken, Eindhoven, the Netherlands.

[1] R. S. Mautner and O. H. Schade, "Television high voltage r.f. supplies," *RCA Rev.*, vol. 8, pp. 43–81; March, 1947.

[2] O. H. Schade, "Radio-frequency operated high-voltage supplies for cathode-ray tubes," Proc. I.R.E., vol. 31, pp. 158–163; April, 1943.

Hence, if $i_m=120$ ma., $L=0.5$ henry, and $C_p=50$ $\mu\mu$fd., we find from (1):

$$V_m = 12 \text{ kv.}$$

This numerical example immediately shows that, with a comparatively small inductance value, and currents that can be furnished by a medium-sized power output tube, peak voltages in the order of 10 kv. can easily be generated across the coil. As shown in the basic diagram of Fig. 1, the interruptions can be effected elec-

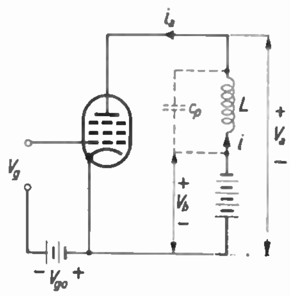

Fig. 1—Schematic circuit diagram of the pulse-voltage generator.

tronically by periodically driving the tube from maximum current to beyond cutoff. Although in principle the plate-current wave shape previous to the interruption is immaterial, a partial sawtooth shape as shown in Fig. 2 has some advantages from the standpoint of screen-grid overload and power efficiency.

In this case the plate voltage drop $L(di_a/dt)$ can, during the time that plate current flows, be kept constant at a value sufficient to draw the maximum current i_m to the plate. The most convenient grid-voltage wave shape is a sawtooth form, as shown in Fig. 2, which, after the interruptions, must drive the grid sufficiently negative to keep the tube cut off in spite of the high positive peaks occurring on its plate. For this and other practical reasons, the plate-current pulse ratio is preferably chosen larger than 0.25. The driver tube obviously should be a pentode or beam-power tube with sufficient maximum emission and adequate anode insulation. If $f_i=1/T$ is the interruption frequency, ΔV_a the permissible plate voltage drop, and α the plate-current pulse ratio as defined by Fig. 2, we find:

$$L\frac{di_a}{dt} = \frac{Li_m}{\alpha T} = \Delta V_a.$$

Hence,

$$f_i = \frac{\alpha \Delta V_a}{Li_m}. \tag{2}$$

If one uses the values $L=0.5$ henry and $i_m=120$ ma., as in the foregoing numerical example, together with $\Delta V_a=300$ volts and $\alpha=0.25$, the interruption frequency f_i given by (2) is

$$f_i = 1250 \text{ c.p.s.}$$

With $C_p=50$ $\mu\mu$fd., the frequency f_0 of the transient oscillations would be approximately 30 kc.

As Fig. 2 shows, the oscillation will immediately cease at the beginning of each plate-current pulse, as a result of the heavy damping caused by the tube.

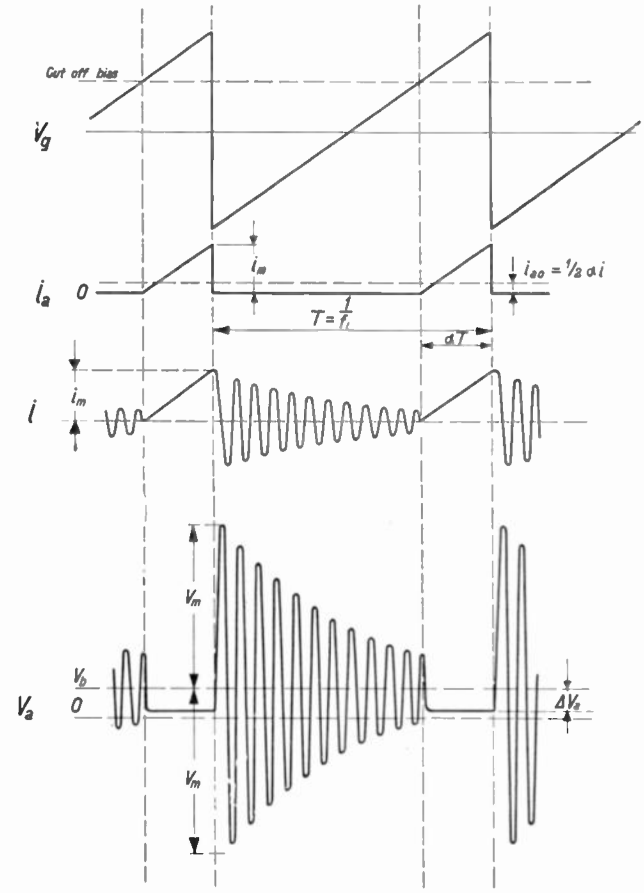

Fig. 2—Voltage and current wave shapes pertaining to Fig. 1.

If a single or multistage rectifier circuit is driven by the transient oscillations across the coil, some energy will be taken from the plate circuit during a small fraction of the first positive half-cycle and the first negative half-period. It is shown in the next section that this fraction, even under heavy external load conditions of the rectifier circuit, will be in the order of 10 per cent of the oscillation period. Consequently, the duration of the rectifier current pulses will be in the order of a few microseconds.

The above discussion indicates the following advantages of the pulse method:

(a) The interruption frequency can be high enough to permit a substantial reduction in the size of the filter capacitors of the rectifier circuit.

(b) The frequency of the transient oscillations is low enough to avoid interference with other parts of the television receiver.

(c) The inductance L is so small that it can easily be realized with a small shell-type core, thus eliminating stray magnetic fields. The unit therefore can be placed conveniently near the cathode-ray tube without causing magnetic disturbances of the electron beam.

(d) As a result of the transient character of the voltages, the insulation requirements on the a.c. side will be less severe than would be the case if the oscillations were continuous.

(e) The output voltage is a function of i_m, and can therefore be easily adjusted by varying the bias of the driver tube.

(f) Automatic control of this bias provides a convenient way of reducing the internal resistance of the supply. It will be shown that, as a result, satisfactory operation at a low power input can be obtained.

(g) Generation of the input power by a tube automatically limits the possible power output, so that the high-voltage supply can be short-circuited without harm.

2. Optimum Number of Rectifier Stages

Although theoretically the value of V_m is not limited and any voltage might be generated by using a single rectifier circuit, for practical reasons it is advisable to take several limiting considerations into account.

In the first place, the maximum anode peak voltages of the driver tube for pulse operation, assuming that this is a tube of moderate size like the 807 or 6GB6G, are limited to approximately 6 kv. The anode voltage, it is true, can be reduced by tapping the anode to the coil, but for a sufficient coupling factor it is found that the tapping ratio must not be too high. Furthermore, the maximum inverse-peak voltages on the rectifiers must be limited to practicable values.

An increase of V_m also unfavorably affects the physical size of the inductance L. It can be deduced from (1) that

$$V_m = 10^{-4} B \sqrt{\frac{q(l+\mu d)}{0.4\pi\mu C_p}} \tag{3}$$

in which B is the peak value of the magnetic flux (gauss), q the cross section of the core (square centimeters), l the average length of the magnetic path (centimeters), d the air gap (centimeters), and μ the permeability of the core material. From (3) it follows that, for a given core material, assuming that the air gap is optimum, the volume of the coil assembly increases with the square of V_m. In fact, the increase is still more rapid than this, as C_p also increases with increasing dimensions of the core.

For the above reasons it is advisable to use voltage multiplication beyond a certain value of output voltage. It is shown in Section 4 that this also offers a slightly lower internal resistance and a better power efficiency, although these consequences are of only secondary importance when automatic voltage control is applied.

Finally, the determination of the optimum number of voltage-multiplier stages is to some extent influenced by the cost of extra rectifiers and capacitors, and by the increase of the stray capacitance C_p they would cause. It is found to be a good practical compromise to choose the multiplication factor n so that V_m is limited to approximately 10 kv.

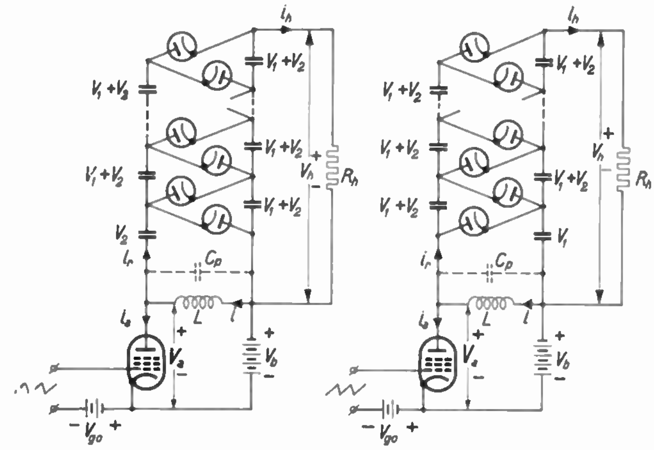

Fig. 3—Schematic diagram of multistage rectifier circuits for even and odd multiplication factors.

The generalized circuits for even and odd multiplication factors are given in Fig. 3. The difference between the two circuits is caused by the necessity of grounding one side of the load. Their performance is analyzed in the next section.

3. Analysis of the Generalized Circuit

It is assumed that a steady-state condition of the circuit has been reached, for a given value of the external load R_h, in which the voltage distribution on the capacitors of the rectifier circuit is that shown in Fig. 3. The capacitance of the capacitors is supposed to be so large that the voltages V_1 and V_2 are substantially constant during the interruption cycle.

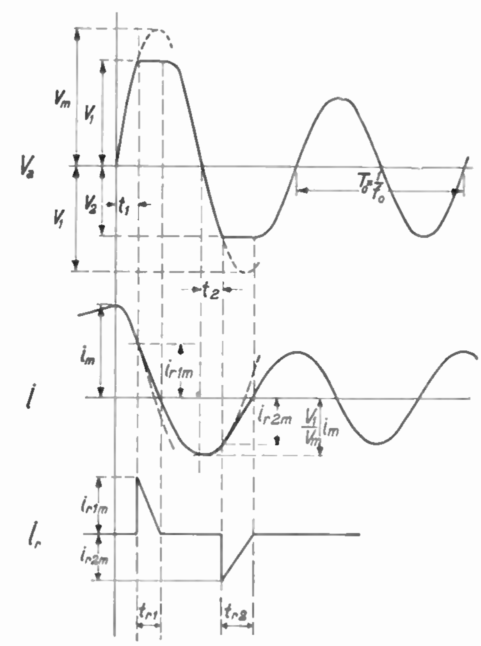

Fig. 4—Detail of voltage and current wave shapes pertaining to Fig. 3.

Furthermore, it is assumed that the internal resistance of the rectifier tubes can be neglected, and that the damping introduced in the resonant circuit formed by L and C_p has no appreciable effect on the first transient oscillation period.

Under these conditions the voltage V_a across the coil, the current i through the coil, and the current i_r supplied by the resonant circuit to the rectifier circuit will, immediately after the interruption, have the general character shown in Fig. 4.

The rectifier circuits will be energized during the periods when V_a tends to increase above V_1 or drop below V_2. During these rectification periods V is constant, and therefore i will follow the tangent of the preceding oscillation curve until $i=0$. No current is supplied to C_p as long as V is constant; during the rectification periods i_r is therefore equal to i, and consequently the current supplied to the rectifier circuit will have the sawtooth pulse form illustrated in Fig. 4.

The operating characteristics of the generalized circuit can be obtained by calculating V_1, V_2, and V_h in terms of a function of R_h.

It is assumed that n_1 rectifiers become conductive during the first, and n_2 rectifiers during the second rectification period. The output voltage will then be

$$V_h = n_1 V_1 + n_2 V_2 \tag{4}$$

with $n_1 = n_2 = n/2$ if the multiplication factor n is even, and $n_1 = (n+1)/2$, $n_2 = (n-1)/2$ if the multiplication factor n is odd.

A relation between V_1 and V_2 can be found from the following energy equations:

$$\tfrac{1}{2}C_p(V_m^2 - V_1^2) = n_1 V_1 i_h T = n_1 V_1 V_h (T/R_h) \tag{5}$$

$$\tfrac{1}{2}C_p(V_1^2 - V_2^2) = n_2 V_2 i_h T = n_2 V_2 V_h (T/R_h). \tag{6}$$

These equations state that the energy loss of the resonant circuit during the first and the second rectification period, respectively, equals the energy which must be supplied per interruption period T to the filter capacitors through the n_1 and n_2 rectifiers.

Setting

$$x = n^2 \frac{T}{R_h C_p} \tag{7}$$

and

$$\frac{V_h}{nV_m} = \frac{n_1}{n}\frac{V_1}{V_m} + \frac{n_2}{n}\frac{V_2}{V_m} = f(x) \tag{8}$$

as the regulation characteristic, it is found from (5) that

$$\frac{V_1}{V_m} = -\frac{n_1}{n} x f(x) + \sqrt{1 + \left(\frac{n_1}{n}\right)^2 x^2 f^2(x)}, \tag{9}$$

and from (5) and (6),

$$\frac{V_2}{V_m} = \sqrt{1 - 2xf^2(x)}. \tag{10}$$

Substitution of these results in (8) then yields the following expression for $f(x)$:

$$f(x) = \sqrt{\frac{P}{Q} + \left\{\frac{P^2}{Q^2} - \frac{\left(\frac{n_1^2 - n_2^2}{n^2}\right)^2}{Q}\right\}^{1/2}} \tag{11}$$

in which

$$P = \frac{n_1^2 - n_2^2}{n^2}\left(1 + 2\frac{n_1^2 - n_2^2}{n^2}x\right) + 2\left(\frac{n_2}{n}\right)^2\left\{1 + \left(\frac{n_1}{n}\right)^2 x\right\}^2 \tag{12}$$

$$Q = \left(1 + 2\frac{n_1^2 - n_2^2}{n^2}x\right)^2 + 8\left(\frac{n_2}{n}\right)^2 x\left\{1 + \left(\frac{n_1}{n}\right)^2 x\right\}^2. \tag{13}$$

For n even, (11) can be simplified to

$$f(x) = \frac{4 + x}{\sqrt{16 + 2x(4 + x)^2}}, \tag{14}$$

which is independent of n. For n odd, however, (11) will contain n as a parameter, and a slightly different regulation characteristic will be found for different values of n, as shown by Fig. 5.

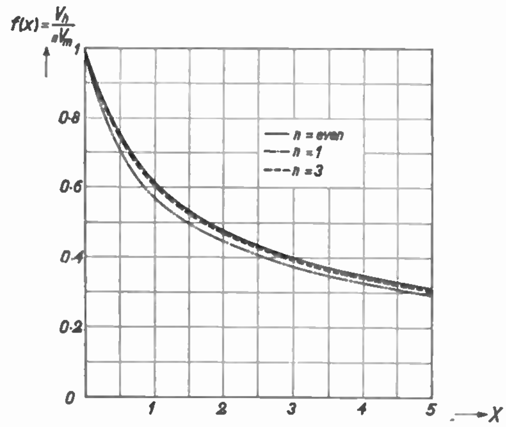

Fig. 5—Output voltage as a function of the load parameter $x = n^2 T/R_h C_p$.

From these generalized regulation characteristics, the usual characteristics giving the output voltage as a function of the output current can easily be derived.

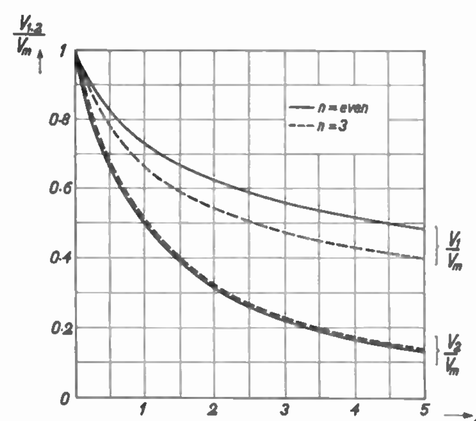

Fig. 6—The peak voltages V_1 and V_2 as a function of the load parameter x.

With $f(x)$ given by (11), V_1/V_m and V_2/V_m can be calculated from (9) and (10). The result is shown in Fig. 6 for n even, and for the case $n=3$ which is used in Section 5. It will be noticed that V_2/V_m falls about twice as fast as V_1/V_m with increasing load, a circumstance which will be used in the next section. From Fig. 4 the peak currents supplied by the resonant circuit to the rectifier circuit can be calculated as follows:

$$\frac{i_{r1m}}{i_m} = \cos\omega_0 t_1 = \cos\left(\arc\sin\frac{V_1}{V_m}\right) = \sqrt{1-\left(\frac{V_1}{V_m}\right)^2} \tag{15}$$

$$\frac{i_{r2m}}{i_m} = \frac{V_1}{V_m}\cos\omega_0 t_2 = \frac{V_1}{V_m}\cos\left(\arc\sin\frac{V_2}{V_1}\right) = \frac{V_1}{V_m}\sqrt{1-\left(\frac{V_2}{V_1}\right)^2} = \sqrt{\left(\frac{V_1}{V_m}\right)^2-\left(\frac{V_2}{V_m}\right)^2}. \tag{16}$$

As V_1/V_m and V_2/V_m are known from (9) and (10), the above equations yield the total rectifier peak currents i_{r1m} and i_{r2m} as a function of x. The result is given in Fig. 7, for n even, for $n=1$, and for $n=3$. The peak currents of the individual rectifiers can be found by dividing i_{r1m} and i_{r2m} by the number of rectifiers concerned, n_1 or n_2, respectively.

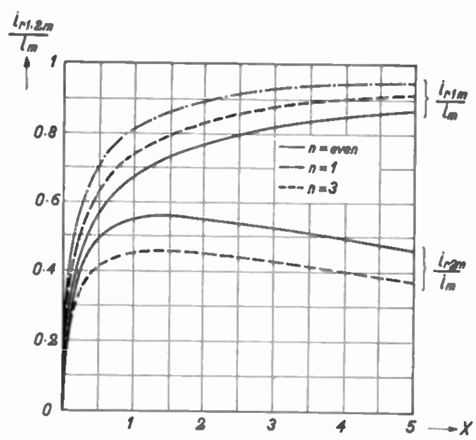

Fig. 7—Rectifier peak currents as a function of the load parameter x.

The durations t_{r1} and t_{r2} of the rectifier-current pulses can be found from

$$\frac{t_{r1}}{T_0} = \frac{1}{2\pi}\frac{i_{r1m}}{i_m}\frac{V_m}{V_1} = \frac{1}{2\pi}\sqrt{\left(\frac{V_m}{V_1}\right)^2 - 1} \tag{17}$$

$$\frac{t_{r2}}{T_0} = \frac{1}{2\pi}\frac{i_{r2m}}{i_m}\frac{V_m}{V_2} = \frac{1}{2\pi}\sqrt{\left(\frac{V_1}{V_2}\right)^2 - 1} \tag{18}$$

which follow from the condition that during the rectification periods the current i, which equals i_r, follows the tangent of the preceding oscillation curve.

Since V_1/V_m and V_2/V_m are known, the rectification-pulse durations as compared to the period T_0 of the transient oscillation can be expressed in terms of x from (17) and (18). The result is shown in Fig. 8 for the case of n even, $n=1$, and $n=3$.

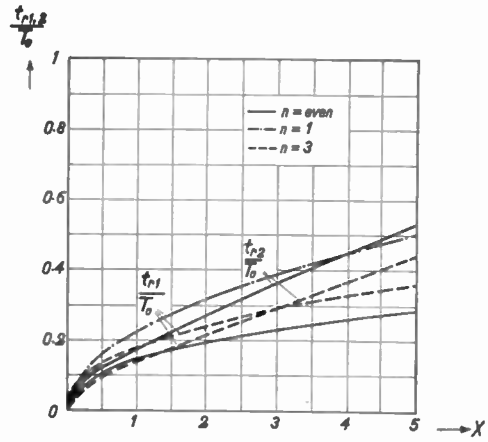

Fig. 8—Duration of the rectifier pulses as a function of the load parameter x.

Upon comparing Fig. 8 with Fig. 5, it is noticed that even for a heavy load, corresponding to a relative drop of the output voltage to 70 per cent, the rectifier-pulse durations will be in the order of 10 per cent of the transient oscillation period, as stated in Section 1.

4. Reduction of the Internal Resistance

The internal resistance of the supply can be found from

$$R_i = -\frac{dV_h}{di_h} = -n^2\frac{T}{C_p}\frac{f'(x)}{f(x)+xf'(x)}. \tag{19}$$

R_i is a function of x which in its general form may be derived from (11), (12), and (13). For small values of x the expression for the internal resistance takes the following general form:

$$R_{i0} = \frac{1}{2}\frac{V_b}{\Delta V_a}\left(\frac{n_1^2+n_1n_2+n_2^2}{n^2}\right)\frac{V_{h0}^2}{W_b} = \frac{1}{2}k\frac{V_b}{\Delta V_a}\frac{V_{h0}^2}{W_b} \tag{20}$$

in which V_{h0} is the output voltage at no load, and W_b the total power input to the circuit.

This relation shows that, for a given type of a circuit, a given relative anode-voltage drop of the driver tube, and a given output voltage, the internal resistance will be inversely proportional to the input power.

For n even we find that $k=\frac{3}{4}$, so that R_{i0} will be independent of the number of rectifier stages.

For n odd, however, we have $k=(3n^2+1)/4n^2$, and in that case the factor k in (20) will vary between $k=1$ for $n=1$ to $k=\frac{3}{4}$ for $n=\infty$.

A numerical example will show that, if no special measures are taken, a relatively large input power will be required to obtain a practicable value of the internal resistance.

If $V_{h0}=25$ kv., $n=3$ (voltage tripling), $V_b=350$ volts, and $\Delta V_a=280$ volts, the power required to give an internal resistance $R_{i0}=5$ megohms, as calculated from (20), will be $W_b=60$ watts, corresponding to a current consumption of $i_b=172$ ma.

Considering the fact that the output power required for a 25-kv. projection television tube is in the order of two or three watts, it is obvious that the foregoing result is hardly a practical one.

To combine good power efficiency with a sufficiently low internal resistance, it is, therefore, necessary to use automatic output-voltage control of some kind.

Automatic control of the driver-tube bias by means of a control voltage derived from the voltage peaks across the resonant circuit has proved to be a cheap and efficient method of obtaining the desired result. The control voltage can be generated by a small diode, connected to a separate winding on the inductor L and fed to the ground side of the grid resistor of the driver tube. For efficient control it has been found advantageous to choose the polarity of the control winding so that the control diode responds to the negative peak voltage V_2, which, according to Fig. 6, drops about twice as fast as V_1 with increasing load, and to use a suitable delay voltage. A practical execution of this principle is given in the following section.

With the above method of automatic voltage control, the value of i_m becomes a function of the output current. The most favorable design will, therefore, be that in which i_m equals the anode current of the driver tube at $V_g=0$, when the output current i_h reaches the maximum value required. Because the internal resistance of the circuit proper, as given by (20), is immaterial when automatic voltage control is applied, it will be possible to obtain a considerable reduction of the input power under maximum load conditions.

It has been found that with the foregoing method the regulation characteristic can be made substantially flat within the desired control range; the output voltage falls very rapidly beyond this range, which is a desirable feature since such a regulation characteristic affords protection against short-circuits.

5. *Practical Design*

The practical design of a high-voltage supply for projection television tubes is described below. In this typical case an output voltage of 25 kilovolts and a maximum output current of 150 microamperes were required, with an internal resistance not exceeding 5 megohms.

The circuit shown in Fig. 9 comprises a conventional blocking oscillator with the triode of the 6SR7 generating a 1000-c.p.s. sawtooth voltage that drives the grid of a 6BG6G output beam power tube. The anode of this driver tube is tapped to the inductor S_1, the top end of which is connected to the rectifier circuit.

Voltage tripling has been chosen to limit the peak voltage across the coil to approximately 8.5 kv. Three indirectly heated oxide-coated-cathode rectifier tubes, which have been developed for pulse operation, are used. The tubes are 1½ inches long by ½ inch in diameter, and have an inverse peak voltage rating in the order of 10 kv. The saturation current of these diodes is approximately 200 ma. The heating power of 0.5 watt per diode is derived from three windings S_2, S_3, and S_4, of a few turns each, coupled with S_1.

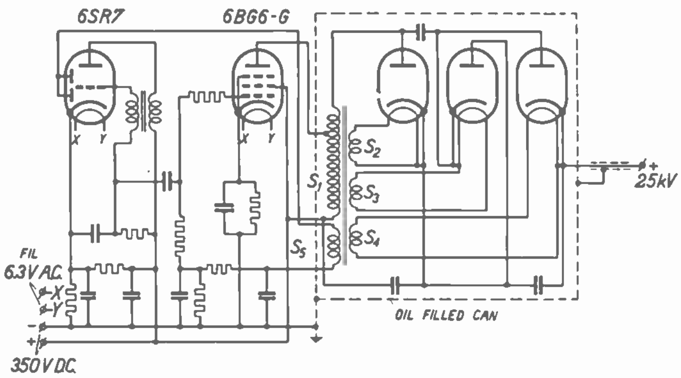

Fig. 9—Schematic diagram of a pulse-type 25-kv. supply.

Another separate winding S_5 generates the automatic control voltage which is rectified by the diodes incorporated in the 6SR7 sawtooth generator. The cathode bias of this tube acts as a delay voltage for the control voltage that is applied to the ground side of the grid resistor of the driver tube through a filter network.

The rectifier heaters cause some extra damping of the resonant circuit in the 6GB6G anode circuit; this, however, reduces the amplitude of the first few peaks of the transient oscillations only slightly, principally affecting the rate of decay of these oscillations. In fact, it can be shown that with $f_0=30{,}000$ c.p.s. the peak voltage will not be more than 3 per cent lower than the value found from (1) in Section 1, if the Q of the coil is reduced to 30.

In order to obtain sufficient heating power for the rectifier diodes at a low total power input, the losses of the resonant circuit proper have been kept low by using a shell-type core of a new magnetic material, known as "Ferroxcube" No. 3, which has a permeability of 800 and a maximum induction of approximately 2000 gauss. This material can be molded in any form, and was found to be highly suitable for the above purpose. Further data on this material have been published elsewhere.[3]

The magnetic circuit consists of a center core, two disks, and an outer ring. A small air gap is left on each side of the center core to reduce the maximum inductance to a tolerable value. The leads to the coil are passed through slots in the disks. The entire core and coil assembly has a volume of approximately 3 cubic inches.

It is evident that such small components cannot be operated in air. The high-voltage coil, rectifier diodes,

[3] J. L. Snoek, "Non-metallic magnetic material for high frequencies," *Philips Tech. Rev.*, vol. 8, pp. 353–360; December, 1946.

and filter capacitors are therefore enclosed in a metal can filled with transformer oil. The assembly is vacuum-impregnated and subsequently sealed.

The 25-kv. high-voltage lead is a polyvinyl-chloride insulated cable, covered with a flexible conductive coating and terminated with a molded connector to the anode terminal of the cathode-ray tube. A 1-megohm protective resistor is molded into this connector, and this resistor, together with a metallic coating on the outside of the cathode-ray tube, also acts as a final ripple-smoothing element.

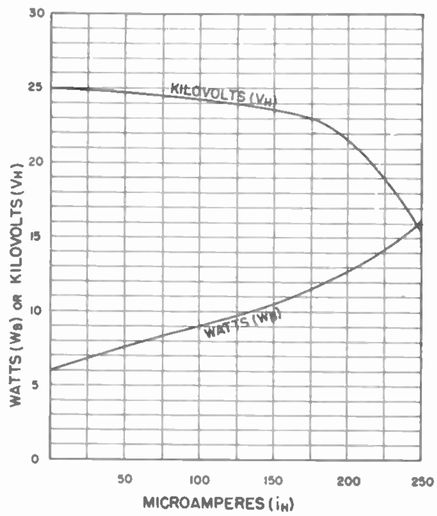

Fig. 10—Regulation characteristic and power-consumption characteristic of the circuit of Fig. 9.

The unit is operated from the existing 350-volt supply of a television receiver. Fig. 10 gives the output voltage V_h and the power input from the "B" supply, W_b, as a function of the output current. It will be noted that the power input compares very favorably with the value found in Section 4 for the equivalent circuit without regulation. Beyond 150 microamperes the output voltage falls rapidly, because the automatic voltage control loses its effect. Through this characteristic the unit is fully protected against short-circuits of the output terminals.

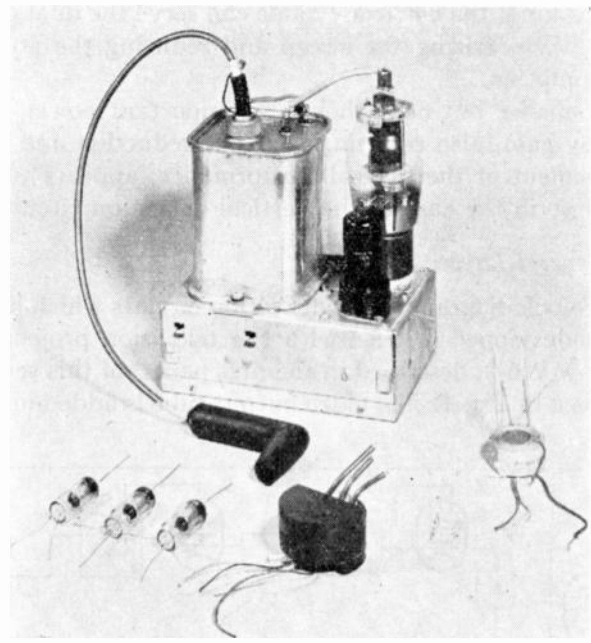

Fig. 11—View of the assembled 25-kv. supply unit with driver circuit. In front are some of the small parts used in the high-voltage circuit.

Fig. 11 shows the completed unit together with a high-voltage coil, a core and coil assembly, and three of the miniature rectifier diodes. The unit occupies a total mounting area in a television receiver of only 25 square inches, and is approximately 7 inches high.

Part III. Deflection Circuits*

J. HAANTJES† AND F. KERKHOF†

Summary—**High-efficiency magnetic deflection circuits which are equally adaptable to projection and direct-viewing television receivers, and a method of obtaining perfect interlace by utilizing the first serration in the vertical synchronizing signal, are described. The horizontal output stage comprises a power-output tube and an "efficiency diode." It is shown that the latter can be used in such a way that it effectively improves the power economy, suppresses spurious oscillations, and improves the sweep linearity. The vertical output stage is coupled to the deflection coils by means of a transformer, which, for reasons of power and space economy, is allowed to introduce considerable distortion of the sawtooth current wave shape. This distortion is compensated by a phase-correcting network in the grid circuit of the output tube. The deflection coils are wound in a flat layer with tapered cross section, and are afterwards bent to shape. A newly developed low-loss magnetic ferrite material has been advantageously utilized in the horizontal-deflection circuit.**

Introduction

The deflection circuits described in this paper are another development in the direction of combined power and space economy for low-cost television circuits. The need for power economy is particularly evident in the case of the horizontal sweep circuits, where the load of the output tube is almost entirely inductive, and the problem arises of coping with the energy stored in this inductance at the end of each stroke so that spurious oscillations during the next stroke are eliminated. Various methods of dissipating this energy have been suggested, but these must be

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considered unsatisfactory from the viewpoint of power efficiency.

It has been proposed to use "efficiency-diode" circuits to solve this problem, but it appears that hitherto this method has found little practical application, probably as a result of linearity difficulties. It is shown in this paper that, with a modification of the known circuits of this kind, the efficiency diode can serve the dual purpose of linearizing the sweep and reducing the power consumption.

A smaller but nevertheless an important power efficiency gain, also resulting in a size reduction and improvement of the over-all performance, appears to be possible in the case of the vertical deflection circuit.

1. General Layout

A block diagram of the deflection circuits which have been developed for use with the television projection tube MW6-2, described in the first paper[1] of this series, is given in Fig. 1. The video output tube is added in the

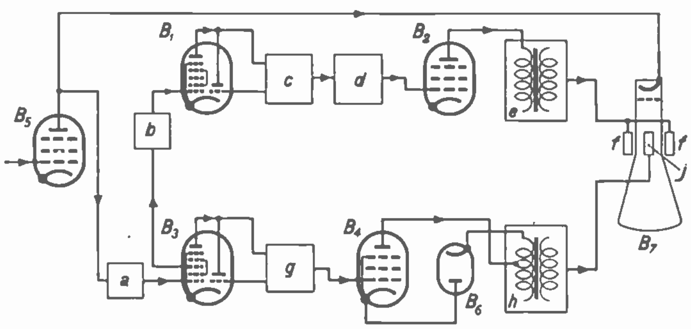

Fig. 1—Block diagram of the deflection circuits.

diagram to show how the complete signal is fed into the time-base unit. Apart from the video output tube, the time base proper is built by using four amplifier tubes and one diode. Two of these tubes are triode heptodes. The two heptode parts are used for the separation of the synchronization signals from the video signal, and for the separation of the vertical synchronization signal from the composite synchronization signal. The two triode parts are used to generate the sawtooth waves for the horizontal and vertical deflection.

The horizontal sawtooth wave is fed to the control grid of the output tube B_4. A so-called "efficiency diode" B_6 is added to this arrangement in order to attain good linearity and a very low power consumption.

The vertical sawtooth wave is fed through a phase-correcting network d to the output tube B_2. This tube feeds the output transformer for the vertical deflection.

The time base as given in the block diagram has been constructed with tubes available on the European market. The two triode heptodes are of the ECH21 type. It appears that an American version of the time base is possible if the two triode heptodes are replaced by a double triode, such as the 6SC7, and two pentagrids, which may be of the 6SA7 type. The triodes are used in generating the two sawtooth waves, whereas the two pentagrids are used in the same way as described in this paper for the separation of the synchronization pulses. The efficiency diode is of a special design and has as yet no equivalent on the American market.

2. The Separation of the Synchronization Signals

It is common practice to separate the horizontal and vertical synchronization signals in a television receiver by feeding the mixture of the signals into two circuits, a differentiating network and an integrating network. The signal at the output of the differentiating network synchronizes the horizontal-deflection circuits. Synchronization of the vertical-deflection circuits is achieved at the moment that the output of the integrating network exceeds a certain threshold value. However, it has seemed in practice that this method of synchronizing for the vertical deflection often gives rise to improper interlacing. One of the reasons is that a receiver almost always contains certain couplings between the two deflection circuits; for example, via the supply voltages. The moment of synchronization of the vertical blocking oscillator may, in this event, be dependent on the phase of the horizontal sweep, giving rise to a pairing of the odd and the even rasters.

The time base described here uses a different method of synchronization for the vertical sawtooth oscillator.

In all television standards established up to the present, several pulses one-half line time apart occur in the vertical synchronization signal. Fig. 2(a) shows the last few horizontal synchronizing pulses and the beginning of the vertical synchronizing pulse. The first of the short pulses occurring in the vertical synchronization signal always comes one-half line time after the beginning of the latter. In our method, this first pulse is separated from the mixture and is used as the synchronizing pulse for the vertical sawtooth oscillator.

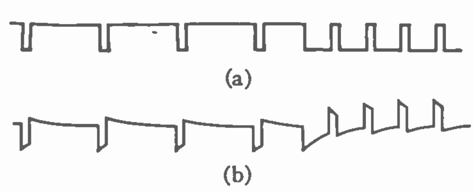

Fig. 2—(a) Shape of the synchronization signal near the beginning of a vertical synchronization signal, after polarity is reversed. (b) Shape of the voltage on the resistor if the voltage of (a) is applied to a series connection of a capacitor and a resistor, the time constant of this combination being about equal to one-half the duration of a line.

To attain this separation, the complete video signal is fed to the first grid of the heptode part of tube B_3 with such a polarity that the synchronization signals form the most-positive part of the signal. The signal has

[1] H. Rinia, J. de Gier, and P. M. van Alphen, "Part I—cathode-ray tube and optical system," Proc. I.R.E., this issue, pp. 395–401.

such an amplitude that only the synchronizing signals are amplified. The anode and the third grid serve as output electrodes of the heptode. The signal on the anode, after it has been differentiated, synchronizes the horizontal sawtooth generator. The third grid of the heptode is set at a relatively low positive potential. It is chosen as an output electrode because it shows a very small capacitance to the anode, which prevents coupling from the horizontal sawtooth generator to the signal on the third grid. The signal on this grid will show the wave shape given in Fig. 2(a). It is fed through a differentiating network with a time constant in the order of one-half line time to the first grid of the heptode of the tube B_1. After this differentiation the signal will show the wave shape of Fig. 2(b). During intervals between two vertical synchronization signals, the first grid of B_1 is cut off by means of a negative potential. The bias is so adjusted that a current can flow only during the pulses in the vertical synchronization signal which, as shown in Fig. 2(b), form the most positive part of the signal.

The screen grids of the heptode of B_1 are fed through a high resistance which is by-passed by a capacitor of a relatively small value (see Fig. 3). Accordingly, if current starts to flow in the tube, the potential of the screen grid will fall very rapidly. The third grid of this tube is coupled to the screen grids through a capacitor, thus causing the third grid to undergo the same voltage drop. The result of this is that only during the first pulse will current flow to the anode. The pulses which come next in the same synchronization signal will not appear in the anode current, as the third grid will suppress any further current to the anode.

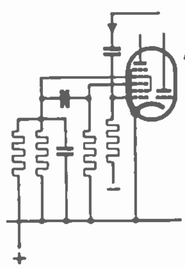

Fig. 3—Circuit for separating the vertical synchronizing signal from the composite synchronizing signals.

The single pulse in the anode current is used as a synchronizing signal for the vertical blocking oscillator. As the moment of synchronization of the vertical sawtooth generator is very exactly determined, a highly accurate interlace is obtained in this way.

The two sawtooth-wave generators used in this time base are blocking oscillators that generate the sawtooth waves with the aid of the two triode parts of the triode heptodes. The sawtooth voltages are built up on capacitors in the grid circuits. The capacitors are discharged by the grid currents and are charged from the 350-volt supply through resistors, so that a high degree of linearity is obtained.

3. The Horizontal Deflection Circuit

In order to clarify the working of the horizontal output stage, we draw attention to the ideal way in which a sawtooth current in a coil may be obtained. This problem has been recently treated by G. C. Sziklai.[2] The ideal circuit comprises the coil with its stray capacitances, a battery, and a switch that is opened and closed at certain predetermined times. As was pointed out in this reference, the switch may be more or less realized with the aid of an output tube and a diode. The cathode of the diode should be connected with the anode of the output tube. The anode of the diode should be connected to a suitable voltage supply, which in practice is replaced by a resistor by-passed with a capacitor. If this is done, the current through the diode is not fed back to the battery, and therefore the current does not contribute to the efficiency of the system. The only gain in efficiency arises from reducing to a certain extent the current of the output tube.

The present authors have found, however, that it is possible to connect the diode to the circuit in such a way that the full efficiency of the theoretical circuit is obtained. As indicated in Fig. 4, the anode side of the transformer winding is continued and the cathode of the diode is connected to the top of this additional winding, whereas the anode of the diode is connected to ground.

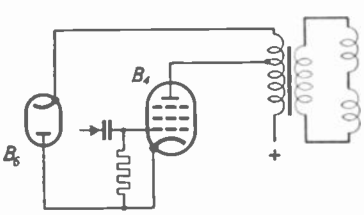

Fig. 4—The way in which the efficiency diode is used in the horizontal-deflection-circuit output stage. No separate filament voltage is needed for the efficiency diode.

To explain why this has to be done, it may be remarked that the output tube will always need a certain positive voltage on the anode in order to be able to provide the necessary current. During the stroke of the sawtooth this voltage must be practically constant, as the current has to increase linearly with time and the load of the transformer is almost entirely inductive.

The additional winding on the transformer is so dimensioned that during the stroke of the sawtooth the potential at the top of the additional winding is just that of ground. If the diode is connected to this point it will be able to stop the oscillations after the fly-back time, in which time the circuit has gone through a half-period of oscillation. From that time on, the voltage across the transformer will be kept at the necessary value, whereas the diode current has such a polarity

[2] G. C. Sziklai, "Current oscillators for television sweep," *Electronics*, vol. 19, pp. 120–123; September, 1946.

that the battery is charged, thereby giving the maximum theoretically possible efficiency.

Apart from suppressing unwanted oscillations and adding to the current efficiency of the system, the diode is also able to contribute to the linearity of the sawtooth current in the deflection coils during the entire stroke. This is achieved in the following way: The current furnished by the output tube is chosen somewhat higher than necessary. The diode takes over the excess current; it therefore passes current during the whole stroke, and by that means keeps the voltage across the transformer at a very nearly constant value. A nonlinearity of the characteristic of the output tube thus does not affect the linearity of the coil current.

During the first part of the stroke the diode will be able to provide all of the necessary current to the transformer. Therefore, the output tube may be cut off during this time. This is achieved by biasing the control grid at such a value that the first part of the sawtooth voltage on this grid falls in the cutoff region. In practice, the output tube is kept cut off during the first one-third of the stroke.

As already mentioned, the efficiency diode used in this circuit is of a new design, as this diode must satisfy some special requirements. It must stand a high inverse-peak voltage, which may reach a value of 4000 volts. Furthermore, it must have a low internal resistance, since otherwise the voltage across the output transformer will not remain sufficiently constant during the stroke of the sawtooth. The EA40 satisfies these requirements. It appears to be possible to feed the filament of this diode from the horizontal output transformer.

As the voltage across the transformer is practically constant during the stroke, the linearity of the sawtooth depends only on the resistance of the transformer and of the coils. The influence of the resistance of the transformer is in most cases negligible. The remaining factor which determines the linearity is the ratio of resistance to inductance, r/L, of the deflection coils. A simple calculation shows that if a maximum nonlinearity of p per cent is permitted at a frequency f of the sawtooth, r/L must approximately satisfy the following relation:

$$\frac{r}{L} < \frac{pf}{100}.$$

Hence, for $f=15{,}000$ and $p=10$ per cent, the requirement is $r/L<1500$. As is well known, the value of r/L depends largely on the dimensions of a coil. For direct-viewing tubes the necessary value is not difficult to obtain with air-core deflection coils. The projection tube MW6-2, however, has a narrow neck. If the horizontal deflection coils were to be mounted closely around the neck of this tube, the value of r/L would be much too high. For this reason the deflection yoke is so constructed that the vertical deflection coils lie adjacent to the neck. Around these are the coils for the horizontal deflection which now, due to their larger size, possess such a value of r/L that good linearity is ensured.

The total current consumption of the circuit is determined solely by the losses of the circuit. The losses are mainly caused by the anode and screen grid dissipation of the output tube and the losses of the output transformer. This is the reason that the new low-loss magnetic material "Ferroxcube," used in the pulse-type high-voltage supply described in the second paper[3] of this series, is also used as core material for the horizontal deflection output transformer.

The total current consumption of the circuit, including the screen-grid current, at a sweep frequency of about 15,000 and at a line length of 48 millimeters (1.89 inches) on the screen of the projection tube with 25 kilovolts on its final anode, is 23 milliamperes at a supply voltage of 350 volts, corresponding to a power consumption of about 8 watts.

4. The Vertical Deflection Circuit

The vertical sawtooth wave is fed through a special network to the control grid of a pentode. The vertical deflection coils are coupled to this tube by means of a transformer. If the impedance of the primary inductance of the transformer is very high with respect to the load impedance, the current through the deflection coils will again have the same shape. This leads, however, to very high values of the primary inductance and to a very uneconomical transformer.

When a smaller transformer with an underrated primary inductance is used, it is possible to compensate for the deformation of the sawtooth current through the deflection coils by a compensating network in the grid circuit.

The main effect of too small a value of the primary inductance is a phase shift of the current in the coil with respect to the tube current. This phase shift is dependent on frequency and is largest for the lowest frequencies. This fact is mainly responsible for the deformation of the sawtooth current.

In Fig. 5 is given the equivalent network of the transformer with load. The resistances and stray inductances of the transformer are omitted because they do not play an important role. L_1 represents the primary inductance of the transformer; r and L represent the resistance and inductance of the deflection coils transformed to the

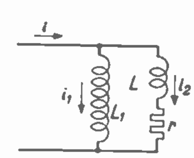

Fig. 5—Equivalent circuit for transformer and deflection coils in the anode circuit of the tube B_2.

[3] G. J. Siezen and F. Kerkhof, "Part II—Pulse-type high-voltage supply," Proc. I.R.E., this issue, pp. 401–407.

anode side of the transformer. The current i_2 through the coils possesses a phase shift Φ_1 with respect to the current i which is fed into the transformer. Φ_1 is determined by

$$\tan \Phi_1 = \frac{r}{\omega(L_1 + L)}.$$

The network of Fig. 6 is introduced between the capacitor C, where the voltage wave is built up, and the first grid of the output tube. A simple calculation will show

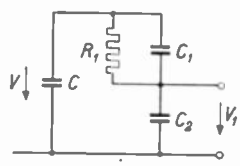

Fig. 6—The phase-correcting network that causes a phase shift opposite to that which occurs upon transmission through the transformer.

that the voltage V_1 on the grid shows a phase shift Φ_2 with respect to the voltage on the capacitor C, which is determined by

$$\tan \Phi_2 = \frac{-\omega R_1 C_2}{\omega^2 R_1^2 C_1 (C_1 + C_2) + 1}.$$

For the frequencies considered, the term containing ω^2 in the denominator is large with respect to 1. Therefore, to a close approximation one may write

$$\tan \Phi_2 = -\frac{C_2}{\omega R_1 C_1 (C_1 + C_2)}.$$

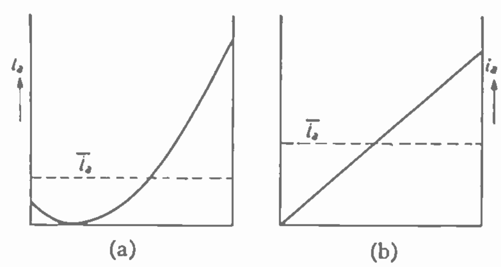

Fig. 7—(a) Shape of the voltage at the control grid, and thus also of the anode current of the tube B_2 when a transformer with too low a primary inductance is used and a correction is introduced for it. The average value of the anode current is indicated by the dotted line. (b) Shape of the voltage and current when a transformer with sufficiently high primary inductance is used, and the same amplitude of the sawtooth current in the deflection coils as for case (a) is obtained.

The phase correction will be correct if for all frequencies $\Phi_1 + \Phi_2 = 0$. The relation which satisfies this condition is

$$\frac{C_2}{R_1 C_1 (C_1 + C_2)} = \frac{r}{L_1 + L}.$$

In practice C_1 and C_2 are given a fixed value, and the value of R_1 is so adjusted that the best correction is obtained.

The phase-corrected voltage at the input of the amplifier tube is found to deviate considerably from the sawtooth form. The form of this voltage, and consequently also that of the anode current, is given in Fig. 7(a). When this is compared with the current form of Fig. 7(b) which refers to the case in which the primary inductance is very high, it is seen that the average current of the tube is lower in the first case. It follows from calculations that the greatest current economy is obtained in the event that $L_1/r = 0.29\tau$ where τ is the period of the sawtooth. In this event the current consumption for case (a) is less than 60 per cent of that for case (b).

In the vertical deflection circuit as described above, the average anode current of the tube B_2 is only 6 to 7 milliamperes.

5. The Deflection Coils

The maximum deflection angle of the projection tube[1] MW6-2 is relatively small, amounting to only 15 degrees from the center position. It is known that magnetic deflection generally introduces a raster distortion and an increase of the spot size. The raster distortion, known as barrel or pincushion distortion, increases with the third power of the deflection angle. The increase of the spot size is due to astigmatism, curvature of the image field, and coma. Of these causes, the first two are the most important.

The errors caused by astigmatism and curvature of the field increase with the second power of the deflection angle, and are proportional to the diameter of the electron beam within the deflection coils. All these errors are also dependent on the distribution of the deflection field. Most of the errors may, therefore, be eliminated by a special field distribution, but it is difficult to eliminate all errors at the same time. In this particular case, however, the coils can be designed in such a way that the raster distortion disappears, because, owing to the small deflection angle and the small beam diameter, the effects of astigmatism, curvature of the field, and coma are very small and hardly noticeable.

The deflection coils are wound as a flat, tapered winding and are bent afterwards into the desired shape. A cylindrical screen of soft-iron wire is wound around the outer coils in order to increase the efficiency of the coils and to lower the ratio of r/L, which is especially important for the horizontal deflection coils. As this iron shield has cylindrical symmetry with respect to the tube neck, it does not disturb the cylindrical symmetry of the focusing field and therefore does not introduce an astigmatism of the focusing lens.

A Developmental Pulse Triode for 200 Kw. Output at 600 Mc.*

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Summary—**The pulse triode A-2212 is a cylindrical triode which gives a peak power output of 200 kw. at 600 Mc. with a tunable external circuit. The tube and its pulse and c.w. performance are described. One of the single-tube circuits developed to test the tube is also described.**

Introduction

EARLY IN 1942, the development of the pulse triode now known as the A-2212 was undertaken under a Navy contract.[1] The tube was intended for use in search radar and was to meet the following electrical requirements: (1) it must have a peak pulse-power output in excess of 100 kilowatts at 600 Mc. with a duty of 0.1 per cent and a 5-microsecond pulse length; (2) it must be operable with external circuits capable of a wide tuning range; (3) it must be air-cooled; and (4) it must operate with no applied voltage in excess of 15 kilovolts. In addition, it was considered essential that the tube be compact, have few and simple parts, and be easy to manufacture on a mass-production basis.

All these requirements were met in the developmental tube H-2614 and the subsequent modification known as the A-2212. After improved circuits for the tube had been developed, it was found that the peak power output per tube at 600 Mc. with 0.1 per cent duty and a 5-microsecond pulse was about 200 kilowatts. This provides a comfortable safety factor over the original requirement.

The Tube

The first problem in the design of the tube was the choice of basic geometry. After both cylindrical and planar geometries had been considered, the cylindrical structure was chosen for the following reasons: (1) In the planar electrode structure with cylindrical symmetry, the voltage distribution across the cathode is a Bessel distribution. Because the area of the cathode can be increased only by increasing the radius of the cathode, the cathode area which can be usefully employed at a given frequency is definitely limited. With a cylindrical structure, the voltage distribution is sinusoidal axially and uniform angularly so that the cathode area can be increased indefinitely, in principle, by increasing the radius as long as the axial length is held constant. This argument was given considerable weight because it seemed likely that the H-2614, if successful, might be used as the basis for future tubes of higher power outputs. (2) The cylindrical structure is more likely to be mechanically stable under varying temperature conditions. (3) The cylindrical structure leads to a tube of smaller radius, which is of some consequence in the design of compact circuits for the tube. (4) The cylindrical structure is more economical of cathode-heating power than the planar structure.

When the basic geometry had been decided upon, the cathode was designed. Previous experience had led to two rough empirical relations for the design of oxide-coated cathodes for pulsed triodes: (1) The power output of pulsed triodes at about 600 Mc. is 1 kw. per ampere of emission. At first glance this relation seems a little absurd in that the operating voltage does not appear. However, the starting time and the peak power output of a pulsed triode both increase as the shunt load resistance is increased. When a reasonable compromise is made between the peak power output and the additional power dissipated at the anode because of the starting time, the load resistance turns out to be such that a peak power output of 1 kw. per ampere is obtained when the tube is operated up to its emission limit. (2) The peak-pulsed emission of an oxide-coated cathode is about 12 amperes per square centimeter with a 5-microsecond pulse.

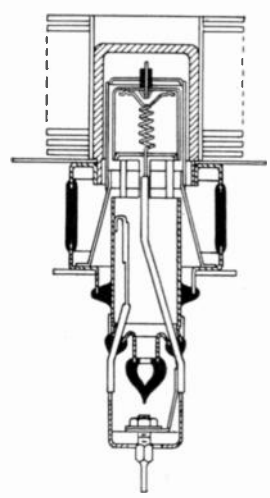

Fig. 1—Cross section of the H-2614 pulse triode.

On the basis of these two relations, a cathode area of 13.5 square centimeters was decided upon. This gives about 160 amperes of emission and allows a reasonable safety factor. From this point on, the electrical design proceeded along the usual lines for the design of high-frequency triodes.

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† Radio Corporation of America, RCA Laboratories, Princeton, N. J.

[1] Naval Research Laboratory Contract N-173S-4815.

Fig. 1 shows the tube H-2614 in section. The cathode is a thimble, oxide-coated on the cylindrical surface. The cathode is supported by tabs on a copper-plated Kovar cylinder about $\frac{3}{4}$ inch in diameter. This large-diameter cathode lead makes possible the use of a reasonably smooth transmission-line circuit between the cathode and grid. The lower end of the cathode lead is pierced by three eyelets. The axial eyelet carries the exhaust tubulation. The other eyelets carry the heater lead and the getter lead. The heater and getter currents are returned through the cathode lead. After the tube is exhausted, the cathode lead is extended by a copper thimble which serves to protect the tubulation and leads. The heater lead is brought out axially through this thimble.

The cathode is heated by a small tungsten helix on the axis of the cathode. An upper and a lower heat shield serve to reduce the end losses so that the cathode may be brought to operating temperature with about 40 watts of heater power.

The grid is a squirrel cage consisting of 90 platinum-clad molybdenum wires, 0.007 inch in diameter, supported by a cone welded to the grid flange. The upper end of the grid is held in alignment by a quartz bead on an axial pin on the cathode. The cathode-grid spacing is 0.019 inch.

The anode is a copper thimble with a U-shaped Kovar annulus silver-soldered to the lip. The internal diameter is such as to give a grid-to-anode spacing of 0.070 inch. The anode is cooled by a "horizontal-type" radiator having an external diameter of 2 inches.

The subsequent tube A-2212 differs from the H-2614 in that the getter is dispensed with, the exhaust tubulation is of copper and is brought out through the end of the anode, the heater lead is brought through the cathode lead axially, and a "vertical" radiator is used.

Fig. 2 is a photograph of the individual tube parts of the H-2614, some of the subassemblies, and of the assembled tube. An A-2212 is also shown.

The low-voltage plate characteristics of a typical tube are shown in Fig. 3. Other data of interest are given in Table I.

TABLE I

Heater voltage	5 volts
Heater current	8 amperes
Grid-to-plate capacitance	15 $\mu\mu$fd.
Grid-to-cathode capacitance	39 $\mu\mu$fd.
Plate-to-cathode capacitance	0.82 $\mu\mu$fd.
Amplification factor	30
Mutual conductance	0.024 mhos at $I_B=0.25$ amperes
Plate resistance	1250 ohms at $I_B=0.25$ amperes
Maximum anode dissipation	300 watts with 10 cubic feet of air per minute

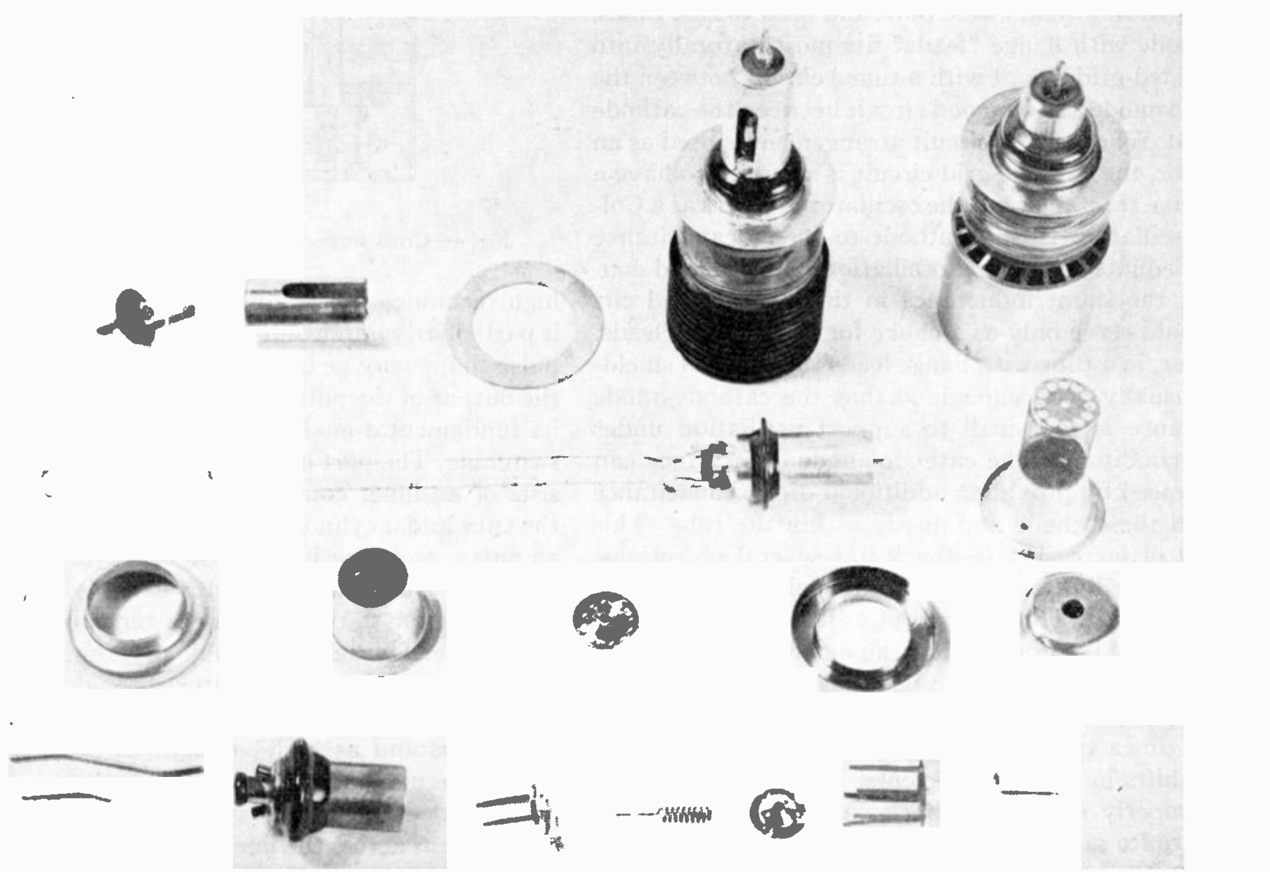

Fig. 2—Individual parts, subassemblies, and a completely assembled H-2614. An A-2212 is also shown.

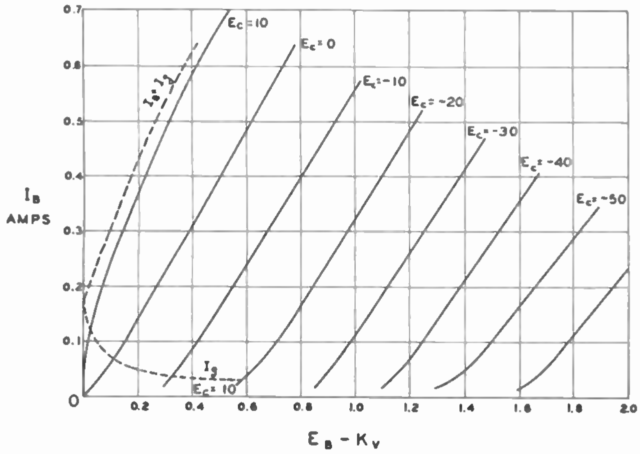

Fig. 3—Low-voltage plate characteristics of the H-2614.

Circuits

While the principal object of the work described in this paper was to design a tube to meet certain specifications, a considerable amount of work was devoted to circuits for the tube. This attention to circuits is natural in the case of high-frequency tubes because a considerable portion of the circuit reactances lie within the envelope of the tube. Hence, it is impossible to design the mechanical features of a tube without a circuit in mind. The H-2614 was designed with two circuits in mind, one with a cavity between grid and anode, the second with a half-wave transmission line between grid and anode. Both of these circuits were built and used to test tubes.

A triode with flange "leads" fits most naturally into a grounded-grid circuit with a tuned circuit between the grid and anode and a second circuit between the cathode and grid. When such a circuit arrangement is used as an oscillator, the cathode-grid circuit is adjusted to have a capacitive reactance and the oscillator operates as a Colpitts oscillator. If the cathode-to-anode capacitance were adequate to support oscillation under loaded conditions, the shunt inductance in the cathode-grid circuit would serve only as a choke for the filament leads. However, in a tube with flange leads the internal shielding is usually good enough so that the cathode-anode capacitance is too small to support oscillation under loaded conditions. The cathode-anode capacitance can be increased by providing additional direct capacitance between the cathode and anode within the tube. This method of increasing feedback has several objections. First, the feedback can be adjusted only by tuning the cathode circuit. This method of adjustment gives only one degree of freedom so that the magnitude and phase of the feedback cannot both be adjusted to obtain optimum operation, a severe limitation when electron-transit times are large enough to produce appreciable phase shifts in the tube currents. Hence, while the use of a properly chosen feedback capcitance within the tube is quite satisfactory for a narrow frequency range, it is not too satisfactory when a very wide frequency range must be covered. Second, a tube with enhanced feedback is not well suited to both oscillator and amplifier use. The alternative is to use external feedback. Then it is relatively easy to obtain two degrees of freedom for the adjustment of feedback, and in addition the oscillator tube is not totally unsuited to amplifier use. These considerations led to the adoption of external feedback systems for the oscillators built to test tubes.

The "half-wave oscillator" circuit is shown in Fig. 4. The grid-anode circuit is a coaxial transmission line, effectively a half-wavelength in length. One quarter-wave of this circuit may be considered as the plate tank and the other quarter-wave as a blocking capacitance which presents a very low reactance to r.f. currents and a very

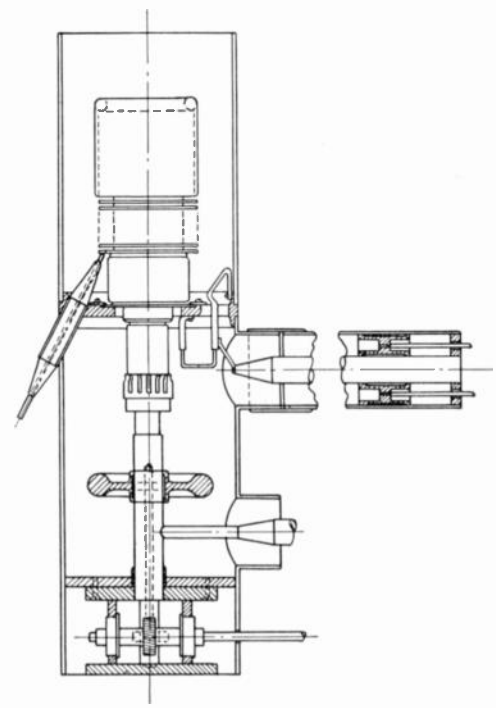

Fig. 4—Cross section of the "half-wave" oscillator circuit.

high reactance to low-frequency currents. This feature is particularly important in pulse applications where the pulse shape may be badly distorted by reactance across the output of the pulser. Because this circuit operates in its fundamental mode, the oscillator is free from mode switching. The part of the line external to the tube consists of an inner conductor comprising the radiator of the tube and a cylindrical extension of the radiator, and an outer conductor in the form of a cylinder 4 inches in diameter. The outer conductor extends beyond the inner conductor so that the line is terminated in a cutoff waveguide. This makes possible the admission of cooling air to a vertical-anode radiator through the end of the line without radiation losses. In order that the circuit may withstand as high voltages as possible with the given external diameter of 4 inches, the external line is proportioned to effect a compromise between the voltage gradient across the line, which varies approximately inversely as the spacing between conductors, and the step-up in voltage between the low-surge-impedance interelectrode line within the tube and the relatively high-surge-impedance line outside the tube, which varies approximately in proportion to the spacing between conductors. The plate voltage is fed through a lead lying in the nodal surface of the circuit.

The cathode-grid circuit consists of a relatively high-surge-impedance coaxial transmission line operating in the three-quarter-wave mode, with a coaxial blocking capacitor adjacent to the tube. The line is tuned by a torus-shaped capacitor which slides on the inner conductor of the line. The oscillator output is obtained from the cathode circuit by a tap directly on the line. The output is taken from the cathode circuit in order to keep the anode circuit clear of objects which increase the voltage gradients and hence induce spark-over, and to take advantage of the low transformation ratio necessary to match a 50-ohm load into the cathode circuit.

The external feedback system consists of two loops in series, one in the grid-anode circuit and the second in the cathode-grid circuit. These loops are tuned by a stub tapped on the loop in the cathode-grid circuit. The stub and tuning capacitor provide the two degrees of freedom necessary for the proper adjustment of the feedback.

The "half-wave" circuit described above and the "cavity" circuit, which is not described for lack of space, were used to test tubes while they were being made in the laboratory. Other circuits were subsequently built. One of these circuits was a modification of the half-wave oscillator which could be operated over the frequency range 400 to 1200 Mc. Push-pull circuits which will be described elsewhere were also built.

Performance

The performance data presented in this section were obtained in the half-wave circuit. In each case, the tube was biased by a cathode resistor. There were two reasons for the use of cathode bias. With a grounded-grid circuit, the oscillator shell and output system can be operated at ground potential when cathode bias is used. Secondly, with cathode bias, operation is stable even when grid emission is large enough to make the plate current exceed the cathode current slightly.

As is customary with pulse tubes having oxide-coated cathodes, the tube was anode-pulsed. What is called the anode voltage in the data is actually the grid-to-anode voltage, i.e., the sum of the anode voltage and the grid-bias voltage. Similarly, the quoted efficiency is the over-all efficiency, not the anode efficiency. In the pulse tests, the duty was 0.1 per cent and the pulse was substantially square, so the average power output and average anode current may be computed by dividing the pulse power and current, respectively, by 1000.

Fig. 5 shows typical pulse operating data on a laboratory-made tube at 600 Mc. The peak power output, the over-all efficiency, and the anode current are plotted against the pulse anode voltage for two values of cathode-bias resistor. It will be noted that in each case the peak power output varies as the 5/2 power of the anode voltage, and the anode current varies as the 3/2 power of the anode voltage until grid emission becomes appreciable, at which point the current begins to rise more rapidly. The efficiency increases slowly with voltage for the lower voltages, and then drops as grid emission sets in.

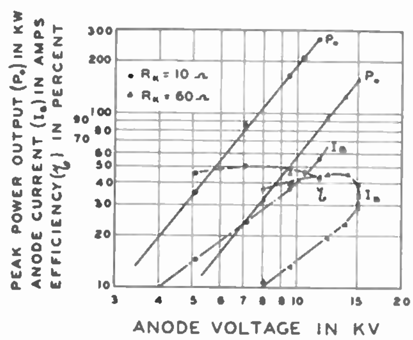

Fig. 5—Typical pulse operating data at 600 Mc.

With the 60-ohm bias resistor, an output of 160 kw. was obtained at 15 kv. In this case, the output was limited by the pulser, but the indications of grid emission suggest that the useful operating point has been passed. With the 10-ohm bias resistor, an output of 265 kw. was obtained at 12 kv. In this case, the output was limited by circuit flashover. In fact, carbon-tetrachloride vapor was used to coax the circuit up to this power level. However, again the evidence of grid emission suggests that the useful operating point has been passed.

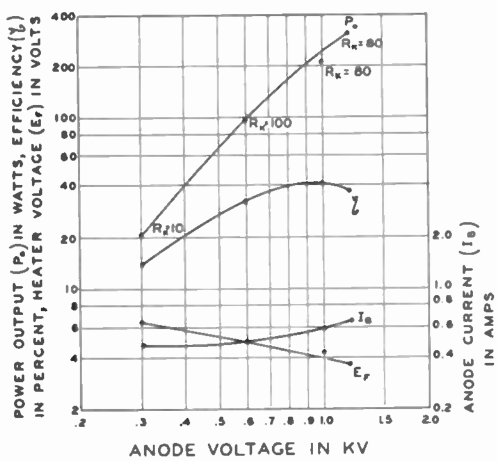

Fig. 6—Typical continuous-wave performance at 600 Mc.

Inspection of the efficiency curves shows that a given output may be obtained with a fixed input for a wide range of anode current. From the standpoint of voltage breakdown, it is advantageous to operate with a low voltage and a high current. At first, it was felt that high-current operation might seriously impair the cathode life. Life tests on tubes operating under conditions comparable to those with the 10-ohm cathode resistor and a power output of 150 kw. have shown that a life in excess of 1000 hours may be expected under these conditions.

While the H-2614 was designed for pulse operation and has a cathode much larger than is required for a continuous-wave tube of comparable average power rating, it was quite natural to make some tests of its performance under c.w. conditions. The results of such a test at 600 Mc. are shown in Fig. 6, in which the power output, plate current, efficiency, and filament voltage are plotted against the anode voltage. For each point on the curves, the cathode bias, feedback, and load were adjusted for maximum power output. The value of the cathode resistor for each point is shown on the power curve. The effects of the cathode-to-grid electron-transit time are evident in the reduction in heater power necessary to keep the tube stable and in the relatively large plate current required for best performance at low voltages. The effects of the grid-to-anode transit time are evident in the very rapid rise of power output with plate voltage. The maximum power output was limited by the tube stability. At the maximum power point, the heater power had been reduced to the point where any perturbation caused the emission to fail and the tube to drop out of oscillation, or caused the cathode temperature, and consequently the grid temperature, to rise so rapidly that the tube "ran away."

To get an idea of the "high-frequency limit" of the tube, it was operated as an oscillator with 250 volts on the anode. It oscillated at all frequencies up to 1100 Mc., at which frequency the efficiency dropped to zero. At 1000 Mc., a power output of 2 watts was obtained.

While the tube was designed specifically for pulse operation and is in some respects poorly designed for c.w. operation, its c.w. performance is such that it has had some application as a c.w. amplifier and oscillator.

Conclusion

In conclusion, it may be said that the pulse triode described above meets all the initial specifications with regard to power output, tunability, cooling, and maximum applied voltages. It meets the minimum pulse-power requirement of 100 kw. at 600 Mc., with a 5-microsecond pulse and 0.1 per cent duty, with a comfortable margin of about 100 per cent. It also gives a c.w. power output of 100 watts at 600 Mc., and has been operated as a c.w. oscillator at frequencies up to 1100 Mc.

Acknowledgment

The authors gratefully acknowledge the contribution to the work described above of John F. Dreyer, Jr., who was a co-worker with them in the early stages of the development.

Circle Diagrams for Cathode Followers*

JOSEPH M. DIAMOND†, ASSOCIATE, I.R.E.

Summary—**Universal circle diagrams are developed which represent the gain, input admittance, and output impedance of the cathode follower. The variables are transconductance and the components of cathode load. The Colpitts oscillator is considered as a cathode follower, and is analyzed with the aid of the circle diagrams, and also algebraically.**

Derivation of the Diagrams[1]

CIRCLE DIAGRAMS may be constructed to represent graphically the variation of the cathode-follower properties (input admittance, gain, and output impedance) with cathode load and transconductance. Fig. 1 shows a cathode follower with general grid-cathode and cathode-ground admittances, and Fig. 2 is its equivalent circuit. Arrows used in connection with voltages indicate the conventional direction of voltage drop. From the equivalent circuit of Fig. 2, these equations may be written:

$$E_kY_k = I_p + I_g \tag{1}$$

$$E_{gk}Y_{gk} = I_g \tag{2}$$

$$E_k + I_pr_p - \mu E_g = 0 \tag{3}$$

$$E_k - E + E_{gk} = 0. \tag{4}$$

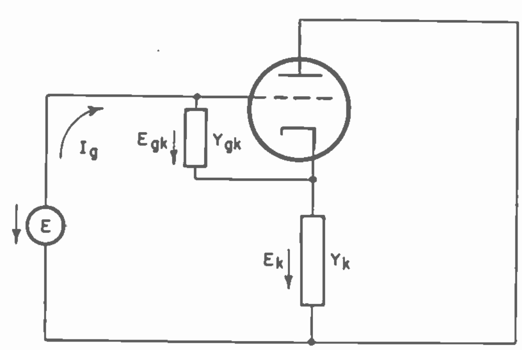

Fig. 1—General cathode follower.

A relation between input admittance and gain can be found immediately:

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[1] The material presented in this paper supplements to some extent the work of K. Schlesinger, "Cathode-follower circuits," Proc. I.R.E., vol. 33, pp. 843–855; December, 1945. See also, H. J. Reich, "Features of cathode follower amplifiers," *Electronic Ind.*, vol. 4, pp. 74–78, 170, 4, 8; July, 1945.

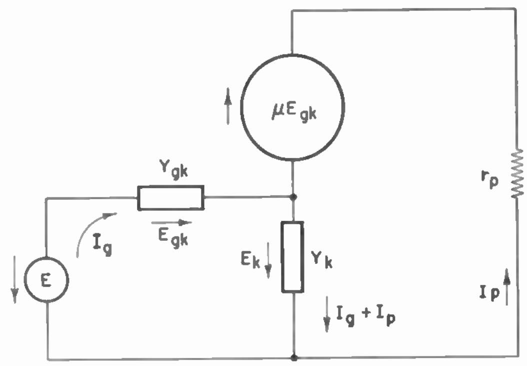

Fig. 2—Equivalent circuit of the general cathode follower.

$$\text{Input admittance} = Y_i = \frac{I_g}{E} = \frac{E_{gk}Y_{gk}}{E}$$

$$= Y_{gk}\left(\frac{E - E_k}{E}\right) = Y_{gk}(1 - A) \quad (5)$$

where the symbol $A = E_k/E$ represents vector gain. Solving (1), (2), (3), and (4) for E_k/E, we have

$$\frac{E_k}{E} = A = \frac{1}{1 + \dfrac{\dfrac{1}{r_p} + Y_k}{G_m + Y_{gk}}} \quad (6)$$

where G_m has its usual meaning of transconductance. Then, using (5)

$$\frac{Y_i}{Y_{gk}} = 1 - \frac{1}{1 + \dfrac{\dfrac{1}{r_p} + Y_k}{G_m + Y_{gk}}}. \quad (7)$$

Using (6), which gives the circuit gain for a given cathode load, it is easy to show that the total output impedance of the cathode follower, including Y_k as well as the output impedance of the tube itself, is:

$$\text{Output impedance} = Z_0 = \frac{1}{G_m + Y_{gk} + \dfrac{1}{r_p} + Y_k}$$

$$= \left[\frac{1}{G_m + Y_{gk}}\right]\left[\frac{1}{1 + \dfrac{\dfrac{1}{r_p} + Y_k}{G_m + Y_{gk}}}\right] \quad (8)$$

or

$$Z_0 = \frac{A}{G_m + Y_{gk}}. \quad (9)$$

Gain, input admittance, and output impedance may now be expressed in nondimensional forms, as functions of the nondimensional quantity

$$\frac{\dfrac{1}{r_p} + Y_k}{G_m + Y_{gk}}.$$

Let

$$\frac{\dfrac{1}{r_p} + Y_k}{G_m + Y_{gk}} = \xi = \alpha + j\beta. \quad (10)$$

Therefore,

$$A = \frac{1}{1 + \xi} \quad (11)$$

$$\frac{Y_i}{Y_{gk}} = 1 - \frac{1}{1 + \xi} \quad (12)$$

$$Z_0(G_m + Y_{gk}) = \frac{1}{1 + \xi}. \quad (13)$$

The quantities of (11), (12), and (13) are the ones which it is proposed to represent graphically. It is evident that A and $Z_0(G_m + Y_{gk})$ will be represented by the same family of curves, which will also apply to Y_i/Y_{gk} with a simple change of axes. Fig. 3 shows the two families of circles which represent the vector quantity Y_i/Y_{gk} as the

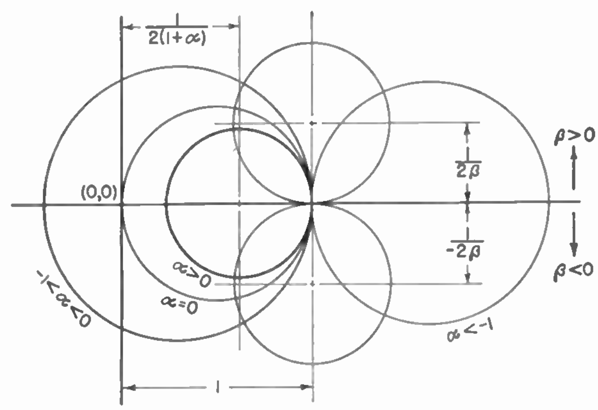

Fig. 3—Loci of $Y_i/Y_{gk} = 1-(1/1+\xi)$ as α and β vary.

components of ξ (α and β) vary. That they are circles may be shown by manipulation of (12).

Returning to the function

$$\xi = \alpha + j\beta = \frac{\dfrac{1}{r_p} + Y_k}{G_m + Y_{gk}},$$

it may be seen that α will not ordinarily be negative, as that would require both a fairly large value of Y_{gk}, and also that the reactive components of Y_k and Y_{gk} be of opposite sign—in other words, incipient series resonance between Y_k and Y_{gk}. Besides, the diagrams are most useful when α and β can be computed very simply from the circuit constants, as can be done for many cases. Fig. 4 shows the loci of Fig. 3 for non-negative α. If the components of Y_k are:

$$Y_k = \frac{1}{r_k} + jb_k, \quad (14)$$

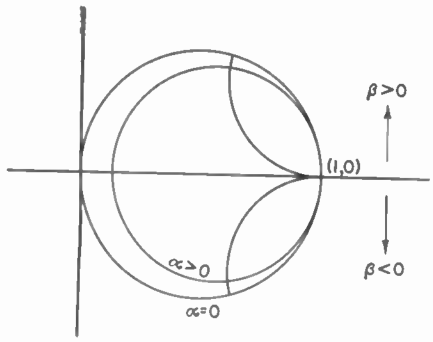

Fig. 4—Loci of $Y_i/Y_{gk}=1-(1/1+\xi)$ for non-negative α.

then

$$\xi = \alpha + j\beta = \frac{\frac{1}{r_p}+\frac{1}{r_k}+jb_k}{G_m+Y_{gk}} = \frac{\frac{1}{R_k}+jb_k}{G_m+Y_{gk}} \tag{15}$$

where

$$\frac{1}{R_k} = \frac{1}{r_p}+\frac{1}{r_k}. \tag{16}$$

If Y_{gk} may be neglected in the sum G_m+Y_{gk}, as is often the case,

$$\alpha = \frac{1}{R_kG_m} \tag{17}$$

$$\beta = \frac{b_k}{G_m}. \tag{18}$$

When (17) and (18) apply, the circle diagrams are very convenient, since the circles represent directly the components of cathode admittance. If Y_{gk} is small, but not small enough to be neglected entirely, simple corrections may be applied to the values of α and β given by (17) and (18). Let

$$Y_{gk} = g_{gk} + jb_{gk}. \tag{19}$$

Then the corrected value of G_m is

$$(G_m)_{\text{corrected}} = G_m + g_{gk} \tag{20}$$

and, approximately,

$$\left(\frac{1}{R_kG_m}\right)_{\text{corrected}} = \frac{1}{R_kG_m} + \frac{b_k}{G_m}\cdot\frac{b_{gk}}{G_m} \tag{21}$$

$$\left(\frac{b_k}{G_m}\right)_{\text{corrected}} = \frac{b_k}{G_m} - \frac{1}{R_kG_m}\cdot\frac{b_{gk}}{G_m}. \tag{22}$$

The right-hand sides of (21) and (22) should make use of the corrected value of G_m as in (20), but this correction is rarely of importance and was not used. If $1/R_kG_m$ and b_k/G_m are of the same order of magnitude, both corrections ((21) and (22)) may either be used or neglected, depending upon the magnitude of b_{gk}/G_m. If one of the components of cathode admittance is much smaller than the other, however, only the smaller one may have to be corrected.

Thus far circle families have been derived which express the variation of the cathode-follower properties with variation in the components of cathode load. Next, the derivation of a circle family which expresses the variation of these properties with transconductance will be indicated. To this end, (7) may be written in the following forms, neglecting Y_{gk} in the sum G_m+Y_{gk} as before:

$$\frac{Y_i}{Y_{gk}} = 1 - \frac{1}{1+\frac{1}{R_kG_m}(1+jb_kR_k)}$$
$$= 1 - \frac{1}{1+\frac{b_k}{G_m}\left(\frac{1}{b_kR_k}+j\right)}. \tag{23}$$

The nondimensional variables are now b_kR_k and either R_kG_m or b_k/G_m. With b_kR_k as parameter, either of the forms of (23) generates the family of circles shown by Fig. (5). If b_k and R_k are fixed, then moving along a constant b_kR_k circle can only represent a change of transconductance: the value of G_m at each point is given by either the R_kG_m or the b_k/G_m circle at the point.

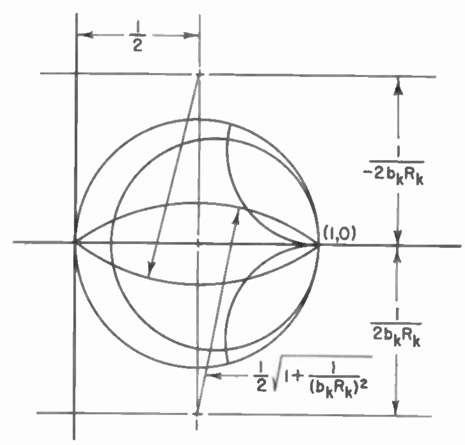

Fig. 5—Loci of Y_i/Y_{gk} with b_kR_k constant.

Before proceeding to specific circuits, a convenient form for R_kG_m may be developed:

$$R_kG_m = \left(\frac{r_pr_k}{r_p+r_k}\right)\left(\frac{\mu}{r_p}\right) = \frac{\mu}{1+\frac{r_p}{r_k}}. \tag{24}$$

If, as is usually the case, Y_{gk} is capacitive,

$$\frac{Y_i}{Y_{gk}} = \frac{Y_i}{j\omega C_{gk}},$$

and, therefore,

$$\frac{Y_i}{\omega C_{gk}} = j\left(\frac{Y_i}{Y_{gk}}\right), \tag{25}$$

from which it follows that a circle family for $Y_i/\omega C_{gk}$ can be obtained by rotating Fig. 4 90° counterclockwise, as shown by Fig. 6. For the general case, $Y_{gk}=|Y_{gk}|\angle\theta$; therefore,

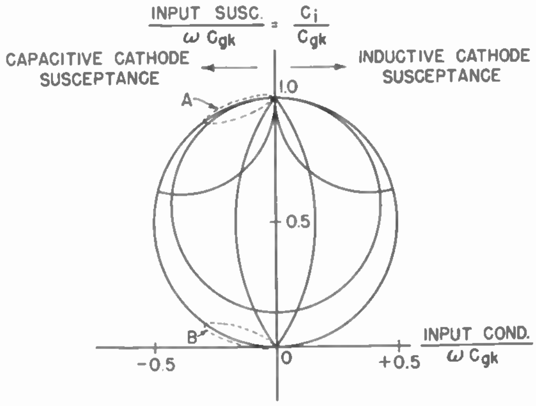

Fig. 6—Loci of $Y_i/\omega C_{gk}$ when $Y_{gk}=j\omega C_{gk}$.

$$\frac{Y_i}{Y_{gk}} = \frac{Y_i}{|Y_{gk}| \angle\theta}$$

and

$$\frac{Y_i}{|Y_{gk}|} = \left(\frac{Y_i}{Y_{gk}}\right)\angle\theta. \tag{26}$$

Therefore (see Fig. 7) the circle family for $Y_i/|Y_{gk}|$ is produced by rotating Fig. 4 counterclockwise through the angle θ.

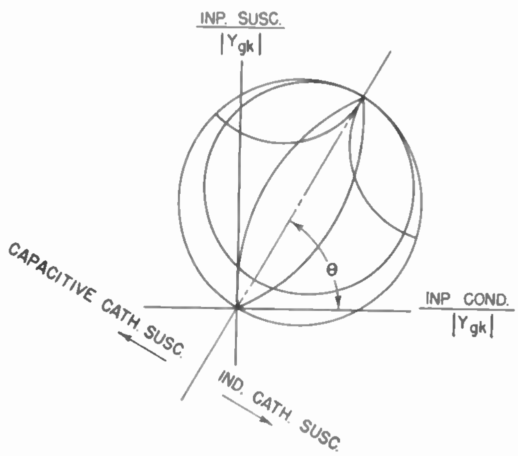

Fig. 7—Loci of $Y_i/\ |Y_{gk}|$.

The locus of gain and output impedance is the same for all these cases, being simply a function of ξ. In connection with output impedance, it may be seen from (9) that an approximate form for Z_0 is

$$Z_0 = \frac{A}{G_m} \tag{27}$$

if Y_{gk} is small compared to G_m. Fig. 8, the diagram provided for purposes of computation, has axes for purely capacitive grid-cathode admittance, since that is the usual case, and for gain; other cases are covered by changing axes. Since the diagram represents $Y_i/\omega C_{gk}$, the ordinate is equal to

$$\frac{\text{input susceptance}}{\omega C_{gk}} = \frac{\omega C_i}{\omega C_{gk}} = \frac{C_i}{C_{gk}}$$

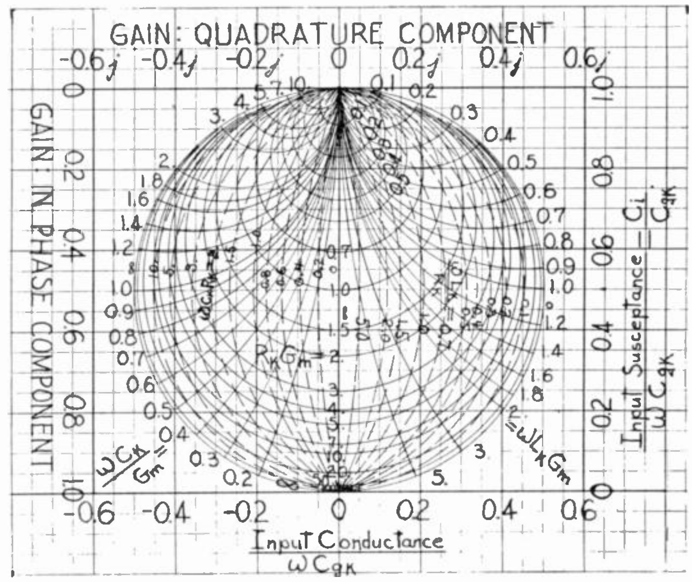

Fig. 8—General cathode-follower circle-diagram family.

where C_i = input capacitance. The value of input resistance, either positive or negative, is the reciprocal of input conductance.

The Cathode Follower as an Oscillator

For purely capacitive grid-cathode admittance, Fig. 6 indicates that the input conductance will be negative if the cathode circuit is capacitive. Thus the possibility of maintaining oscillation in the grid circuit is suggested. Such a circuit is, of course, nothing more than a Colpitts oscillator, as has been pointed out before.[1]

Adding a conductive component to the grid-cathode admittance (see Fig. 7 for $\theta=0$) adds to the input capacitance (assuming the cathode circuit capacitive) and adds a positive component of input conductance, both of which are undesirable in an oscillator design—though this method might be used in atempting to avoid oscillation. This effect also shows how grid current may limit the amplitude of oscillation in a circuit of this sort. In further discussion of cathode-follower oscillators, then, it will be assumed that any grid-cathode coupling is purely capacitive. The left half of Fig. 6 is, therefore the area which results in negative input conductance.

For fairly low frequencies, at which the possibility of maintaining oscillation at all is not a limitation, there is a wide choice of operating conditions, corresponding to various subareas of Fig. 6. This choice will be influenced by such factors as:

1. Whether or not power is to be taken from the circuit, and if so, the impedance level required.

2. The point from which voltage or power is taken. The cathode has the advantage here, since the output circuit is not directly coupled to the tuned circuit. Small voltages might also be taken from the plate, across a small impedance (compared to r_p) in the plate circuit.

3. If voltage is required, the magnitude of output voltage.

4. The factors against changes of which it is desirable to provide frequency stabilization. Subarea A (Fig. 6) will stabilize C_i against changes in C_k, R_k, and G_m, but not against C_{gk}. The output impedance and voltage will be low. Subarea B will stabilize C_i against changes in C_k, R_k, G_m, and C_{gk}. The output impedance and voltage will be high. It is worth while to note here that the input capacitance due to C_{gk} and C_k can be made very small (see Fig. 8 and also the algebraic analysis of the cathode-follower oscillator which will follow), and in particular much smaller than C_kC_{gk}/C_k+C_{gk}, which is the equivalent series capacitance of the two. This conclusion may be restated: In a Colpitts oscillator, the tuning capacitance contributed to the tank circuit by the voltage-dividing capacitors is not equal to their equivalent series capacitance, and may be much less.

At frequencies for which ωC_{gk} becomes comparable to G_m, the circle method requires the approximate corrections of (21) and (22), and possibly the exact solution of (15) for α and β; for such cases, therefore, an algebraic analysis of (7) is in order. This can be accomplished fairly easily if

$$\frac{1}{R_k} = \frac{1}{r_p} + \frac{1}{r_k}$$

can be neglected in the sum $1/r_p+Y_k$; that is, if the cathode circuit is predominately reactive. If so, (7) becomes, for capacitive cathode and grid-cathode admittances,

$$Y_i = j\omega C_{gk}\left(\frac{1}{1+\dfrac{G_m+j\omega C_{gk}}{j\omega C_k}}\right). \tag{28}$$

From which the following expressions are derived:

input conductance

$$= \frac{-G_mC_kC_{gk}}{(C_k+C_{gk})^2}\cdot\frac{1}{1+\left[\dfrac{G_m}{\omega(C_k+C_{gk})}\right]^2} \tag{29}$$

input capacitance

$$= \frac{C_kC_{gk}}{C_k+C_{gk}}\cdot\frac{1}{1+\left[\dfrac{G_m}{\omega(C_k+C_{gk})}\right]^2} \tag{30}$$

$$\frac{\text{negative input conductance}}{\text{input capacitance}} = \frac{G_m}{C_k+C_{gk}}. \tag{31}$$

Equation (31) can be considered a high-frequency figure of merit for a cathode-follower oscillator, since the larger the ratio of negative input conductance to input capacitance, the higher in frequency it will be possible to maintain oscillation. An expression for input resistance follows from (29):

$$R_i = \frac{-(C_k+C_{gk})^2}{G_mC_kC_{gk}}\left(1+\left[\frac{G_m}{\omega(C_k+C_{gk})}\right]^2\right). \tag{32}$$

For a given G_m, $-R_i$ will be a minimum if

$$\frac{G_m}{\omega(C_k+C_{gk})}\rightarrow 0, \quad \text{and} \quad C_k = C_{gk};$$

under these conditions, we have[2]

$$(-R_i)_{\min} = \frac{4}{G_m}. \tag{33}$$

The problem arises next of producing the required negative resistance with a minimum of input capacitance. Equation (31) shows that, for a given G_m, this condition corresponds to making C_k+C_{gk} a minimum. Equation (32) can be solved for C_k+C_{gk}:

$$C_k+C_{gk} = \frac{C_k(-R_i)G_m \pm \sqrt{C_k^2(-R_i)G_m[(-R_i)G_m-4]-4\left(\dfrac{G_m}{\omega}\right)^2}}{2}.$$

Using the minus sign before the radical for the smaller value of C_k+C_{gk}, and setting

$$\frac{d(C_k+C_{gk})}{dC_k} = 0,$$

we have

$$\frac{\omega C_k}{G_m} = \frac{\omega C_{gk}}{G_m} = \frac{1}{\sqrt{(-R_i)G_m-4}}. \tag{34}$$

The input susceptance corresponding to these values is

$$\omega C_i = \frac{2}{(-R_i)\sqrt{(-R_i)G_m-4}}. \tag{35}$$

Therefore, C_k should equal C_{gk}, both having the value given by (34). Equations (34) and (35) indicate that the minimum negative-input resistance obtainable with a given G_m is $4/G_m$, confirming (33).

At frequencies for which $\omega(C_k+C_{gk})\gg G_m$, the negative-input resistance approaches $(C_k+C_{gk})^2/G_mC_kC_{gk}$, and the input capacitance approaches C_kC_{gk}/C_k+C_{gk}, these being a minimum and maximum, respectively, with respect to frequency. If $C_k=C_{gk}$, these asymptotic values become $4/G_m$ and $C_k/2$, respectively.

ACKNOWLEDGMENT

The author is indebted to Jerry Shmoys, who suggested a number of changes in notation and presentation, and to members of the staff of Bendix Radio Corporation, who helped in the preparation of the manuscript.

[2] In equations (30) to (38a), Schlesinger (see footnote reference 1) derives the formula $(-R_i)_{\min}=8/G_m$. However, the derivation applies only to the case of variable G_m, keeping the circuit capacitances constant, while $4/G_m$ is the minimum negative input resistance that may be produced by a tube of given G_m, by varying C_k and C_{gk}.

A Note on Frequency Transformations for Use with the Electrolytic Tank*

W. H. HUGGINS†, ASSOCIATE, I.R.E.

Summary—**Two frequency transformations are described which transform the circular electrolytic tank described by Hansen and Lundstrom into rectangular and elliptical tanks, respectively.**

The rectangular tank is particularly suited for simulating highly damped circuits where response extending over many octaves in frequency is desired. Furthermore, the co-ordinates of the tank may be calibrated directly against "logarithmic frequency" and dissipation factor, respectively.

The elliptical tank is particularly adapted to representation of band-pass circuits having characteristics which are geometrically symmetric about a center frequency. Not only does this transformation expand the scale of the tank and increase its accuracy, but the geometrical relationships between the pole electrodes for simple resonant circuits become substantially independent of the relative bandwidth.

The Semicircular Tank

IT WILL BE assumed that the reader is familiar with the representation of rational functions in terms of complex poles and zeros,[1,2] and also with the potential analogue of these functions.[3] Although Hansen and Lundstrom discuss a *circular* electrolytic tank, it is important to realize that a *semicircular* tank will suffice in studying physical networks, since all poles and zeros will either be located upon the imaginary-frequency (real-p) axis or be distributed in "conjugate" pairs symmetrically about it. Since the semicircular tank is taken as the basis of the two transformations herein described, it seems advisable to illustrate this property with a simple example.

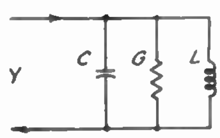

Fig. 1—Simple resonant circuit.

We consider the simple resonant circuit shown in Fig. 1. If this circuit were used as the plate load in an amplifier, the gain would be proportional to the impedance Z.

$$Z = -j\frac{1}{C}\frac{\omega}{(\omega-\omega^+)(\omega-\omega^-)} \tag{1}$$

where

$\omega^+ = \omega_0\underline{/\theta}$, $\omega_0 = \dfrac{1}{\sqrt{LC}}$, the resonant frequency

$\omega^- = \omega_0\underline{/\pi-\theta}$, $M = \sqrt{\dfrac{C}{L}}$, the characteristic admittance

$\theta = \sin^{-1} d/2$, $d = \dfrac{G}{M}$, the dissipation factor.

In the circular tank, the circuit impedance given by (1) would have "pole" electrodes at ω^+ and ω^-, and a "zero" electrode at the origin (i.e., at the conducting rim of the tank), as shown in Fig. 2. It is of interest to note that the radial distance of either pole from the origin is equal to the resonant frequency of the circuit, and the vertical displacement from the real-frequency axis is exactly equal to one-half of the 3-db bandwidth of the circuit. Thus, if the loading conductance were increased, both poles would move upward along a circular locus of constant radius ω_0. For critical damping, both poles would be coincident on the imaginary-frequency axis at $j\omega_0$, and the real frequency of oscillation would be zero.

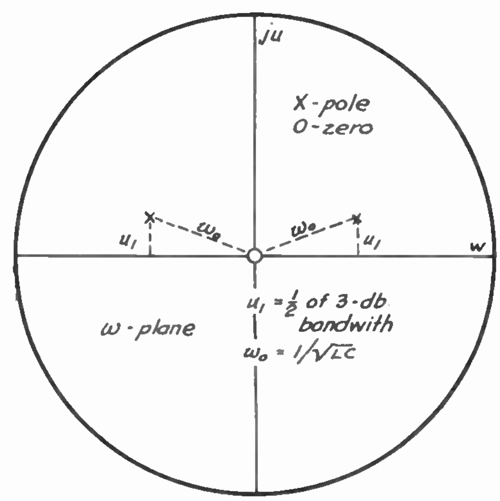

Fig. 2—Representation of the circuit of Fig. 1 on a circular tank.

But, regardless of the damping, symmetry will always be maintained between the right and left halves of the tank and *no current will cross the imaginary-frequency axis*. An insulating partition may, therefore, be inserted along this axis and, once this partition is in place, the left half of the tank may be removed entirely without upsetting in any way the potential field in the right half. In one sense, the insulator strip simulates *by reflection* the fields that would result from the poles and zeros located in the left half-plane.

* Decimal classification: 510×R140. Original manuscript received by the Institute, October 4, 1946; revised manuscript received, July 11, 1947.

† Cambridge Field Station, Air Materiel Command, U. S. Air Forces, Cambridge 39, Mass.

[1] W. E. Bradley, "Wide-band amplifier design," submitted for publication in Proc. I.R.E.

[2] H. W. Bode, "Network analysis and feedback amplifier design," D. Van Nostrand Co., New York, N. Y., 1945.

[3] W. W. Hansen and O. C. Lundstrom, "Experimental determination of impedance functions by use of an electrolytic tank," Proc. I.R.E., vol. 33, pp. 528–534. August, 1945.

A little consideration of the circular tank will show that only half of the current from any electrode lying along the imaginary-frequency axis or at infinity (i.e., at the outer boundary) will flow into the right half-plane. Therefore, in using the semicircular tank, *all boundary electrodes should be fed with only half the current that would be used for "internal" electrodes.* With this adjustment, the fields in the semicircular tank should be identical with those obtained in the circular tank with the advantage that only one-half as many "internal" electrodes are required and the tank may be made smaller physically.

The Logarithmic Transformation $z = \ln \omega$

If a semicircular tank in the ω plane is transformed conformally onto a z plane, where $z = \ln \omega$, it is found that a pole located at $\omega^+ = \omega_0/\sin^{-1} d/2$ corresponds to a pole on the z plane at

$$x = \ln \omega_c \tag{2}$$

$$y = \sin^{-1} d/2. \tag{3}$$

Hence, the entire right half of the ω plane is transformed onto a strip in the z plane lying between $y = \pi/2$ and $y = -\pi/2$. The semicircular tank therefore transforms into the rectangular tank shown in Fig. 3. The top and bottom walls are made of insulating material to simulate by reflection the poles lying in all other period strips (the value y given by (3) is multivalued), and the end pieces are conducting strips to simulate the circular equipotentials near zero and infinite frequencies.

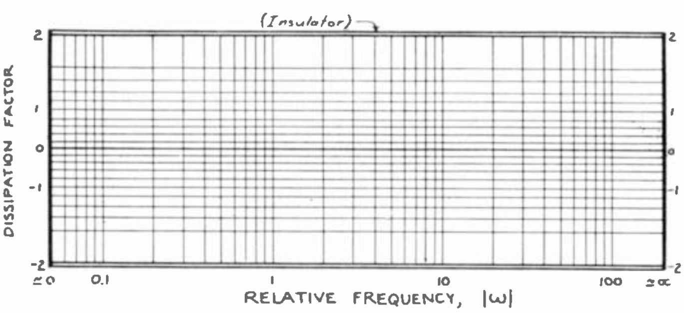

Fig. 3—Rectangular electrolytic tank.

The structural advantages of such a rectangular tank are obvious. It possesses the added advantage that the bottom of the tank may be calibrated directly in rectangular co-ordinates which show the resonant frequency and dissipation factor corresponding to each of the electrodes.

The outstanding electrical advantage of the rectangular tank is that the logarithmic frequency scale enables one to measure characteristics that extend over a very large ratio in frequency. It is, therefore, particularly well adapted for studying broad-band circuits in which the asymptotic behavior at high or low frequencies may be of interest. For this same reason there is little need for finite-boundary corrections, since the equipotentials very rapidly become parallel to those at infinity and zero. For example, the addition of a single decade on the end of the rectangular tank is equivalent to *increasing the size of the circular tank by 10 times!*

The rectangular tank also provides a "feel" and an understanding of the asymptotic behavior of network functions. To illustrate, the low-frequency response of a simple R-C coupling circuit is

$$A = \frac{\omega}{\omega - j\alpha}. \tag{4}$$

This gain function has a pole at $\omega = j\alpha = j(1/RC)$ and a zero at the origin. Since the imaginary-frequency axis corresponds to a dissipation factor of 2, a "pole" electrode is inserted along the upper rim of the tank at $\omega = 1/RC$ and a "zero" electrode is attached to the left end of the tank. Assuming that $\alpha = 10$, the potential distribution along the central-frequency axis would be found to vary as shown in Fig. 4.

Fig. 4 also illustrates the method of calibrating the tank. Since there is a zero at the origin, the gain must ultimately decrease linearly by 20 db per decade decrease in frequency. Hence, for making the initial calibration, a pole may be attached to the right edge of the tank, a zero to the left edge, and the current and probe-voltmeter sensitivity adjusted to give a voltage deflection of 20 units per decade motion in the middle region of the tank. Once this calibration has been made, it should in no way be affected by the addition of other pole and zero electrodes, and the *probe voltmeter will read directly in decibels.* It should be recalled, however, that, since the pole and zero were on the boundary, the current thus calibrated is one-half the value that would be used for electrodes not on the boundary of the tank.

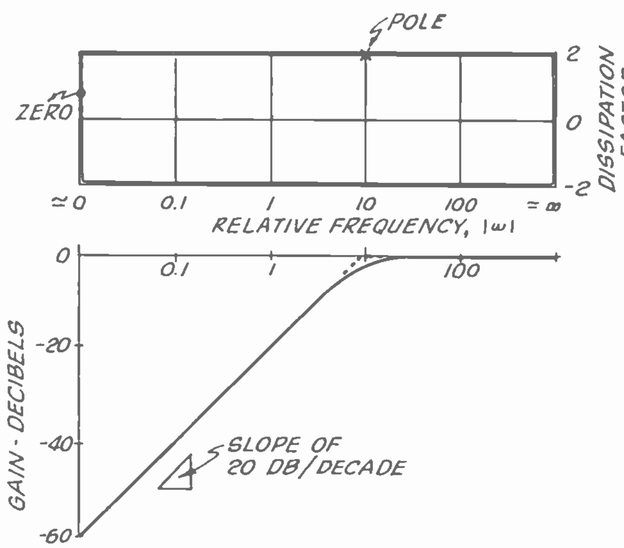

Fig. 4—Representation of the simple R-C coupling circuit in the rectangular tank. The gain characteristic is the plot of potential along the center axis of the tank.

It is of considerable practical value that the sum of the "straight-line" asymptotes will yield a very good approximation to the actual characteristic, provided the corners are rounded off by 3 db. As an illustration, suppose that low-frequency compensation were added to the R-C coupling network represented in Fig. 4. In terms

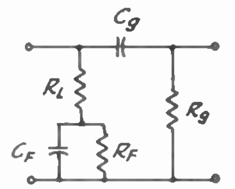

Fig. 5—Low-frequency coupling network.

of the circuit parameters shown in Fig. 5, this compensated network will have poles at

$$\hat{\omega}_1 = j\frac{1}{R_gC_g}, \qquad \hat{\omega}_2 = j\frac{1}{R_FC_F}$$

and zeros at

$$\bar{\omega}_1 = 0, \qquad \bar{\omega}_2 = j\left[\frac{1}{R_LC_F} + \frac{1}{R_FC_F}\right].$$

For a somewhat over-compensated case, the response would appear as shown in Fig. 6. Note that by placing $\bar{\omega}_2$ on top of $\hat{\omega}_1$ the network would behave exactly as a simple R-C circuit with a time constant 10 times that shown in Fig. 4.

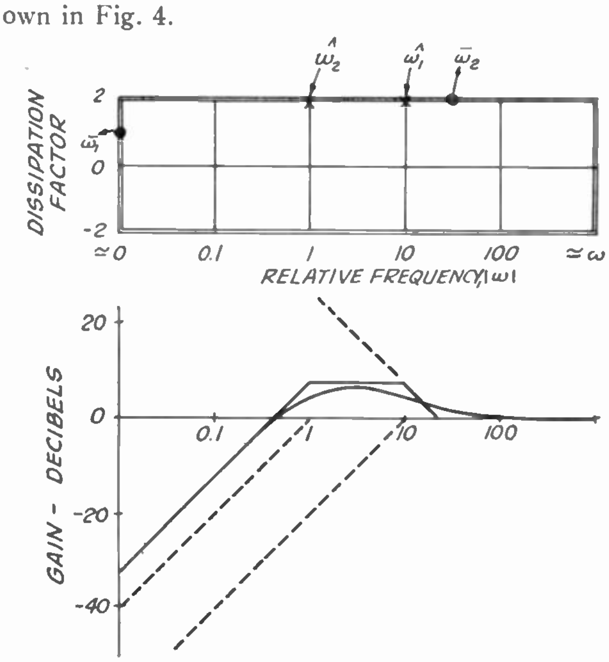

Fig. 6—Low-frequency-compensation circuit. The potential plot shows individual asymptotes as broken lines, and the sum of asymptotes and actual characteristic as heavy lines.

The Band-Pass Transformation $z = \omega - 1/\omega$

When the desired function F possesses symmetry about some "center" frequency, which in this discussion will be *normalized to unity* so that $|F(\omega)| = |F(1/\omega)|$, the transformation $z = \omega - 1/\omega$ may be used to achieve a further reduction in the number of electrodes required to represent the function. Thus, by means of this transformation, a stagger-tuned amplifier having n circuits in each group may be represented concisely by $n/2$ electrodes in the semielliptical tank, whereas the same representation in an unmodified circular tank would require $2n+1$ electrodes.

Consider a pole in the ω plane located at $\omega_1 = w_1 + ju_1 = \omega_0/\underline{\theta}$. On the z plane, this pole will be located at $z_1 = x_1 + jy_1$ where

$$z_1 = (w_1 + ju_1) - \frac{1}{(w_1 + ju_1)}$$

or

$$x_1 + jy_1 = w_1\left[1 - \frac{1}{w_1^2 + u_1^2}\right] + ju_1\left[1 + \frac{1}{w_1^2 + u_1^2}\right] \tag{5}$$

$$= \left[\omega_0 - \frac{1}{\omega_0}\right]\cos\theta + j\left[\omega_0 + \frac{1}{\omega_0}\right]\sin\theta. \tag{6}$$

Equation (6) shows that, for poles located at frequencies of magnitude considerably greater than the "center" frequency, the transformation has little effect, and that $x_1 \simeq w_1$ and $y_1 \simeq u_1$. The transformation maps all points of the ω plane lying inside of the unit circle (shown as a broken line in Fig. 7) onto the left half of the z plane,

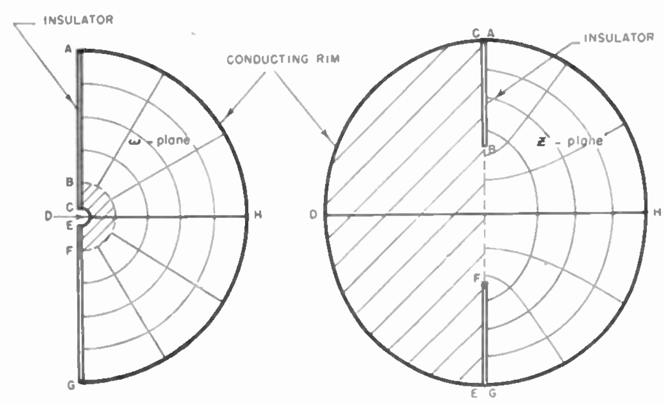

Fig. 7—The transformation of the "semicircular" tank into an "elliptical" tank by means of $Z = w - (1/\omega)$.

and it transforms the circumference of this circle onto the imaginary axis of the z plane. Further consideration will show that each point on the z plane corresponds to two points on the ω plane, one *inside* of the unit semicircle and one *outside* of the unit semicircle. To resolve this ambiguity, we chose the Riemann surface corresponding to the positive radical sign,

$$\omega = \frac{z}{2} + \sqrt{1 + \left(\frac{z}{2}\right)^2}, \tag{7}$$

and we simulate electrically the required branch cut by inserting into the elliptical tank two insulator partitions as shown in Fig. 7.

Now, when the function to be represented possesses geometric symmetry in the ω plane, it will have arithmetic symmetry in the z plane about the axis $ABFG$ of Fig. 7. It is then possible to extend the insulator completely across the elliptical tank and to remove the left half without disturbing in any way the potential field in the remaining right half of the tank. The semielliptical boundary is so nearly circular that a semicircular tank could probably be used without appreciable error. In fact, because of the two-times expansion of the scale near the center of the z plane, it is probable that a semicircular tank used in this way would yield more accurate results than if used to represent the ω plane. It must, however, be recalled that any pole or zero lying along the unit circle in the ω plane will be transformed onto the boundary of the semielliptical tank, and that these pole or zero electrodes should therefore be fed with one-half of the normal current, for reasons previously discussed.

It should be mentioned parenthetically that this particular transformation also possesses considerable utility in the analytic study of amplifiers having tuned circuits of the type shown in Fig. 1. The principal advantage is due to the fact that, since only the poles remain on the finite part of the z plane, the response of a symmetrically tuned amplifier having n staggered stages is characterized by an nth degree polynomial P_n in $(z/a)^2$ (where a is the normalized bandwidth) instead of by a rational fraction. It may be shown that, for "flat-flat" response of the Butterworth type,

$$P_n = 1 + (z/a)^{2n}, \tag{8}$$

while for maximum selectivity and bandwidth, consistent with an allowable tolerance ϵ in the gain within the pass band, the polynomial will be

$$P_n = 1 + \epsilon T_{2n}(z/a) \tag{9}$$

where T_{2n} is the Tchebyscheff polynomial of degree $2n$. These relations will be valid regardless of whether the desired bandwidth is small or large compared to the center frequency. Thus, it is found that the electrodes representing a "flat-flat" amplifier will all be located equally spaced upon the circumference of a circle having a radius equal to the normalized bandwidth a. For Tchebyscheff overstaggering, these poles will lie on an ellipse whose foci correspond to the real frequencies at the edges of the pass band and whose eccentricity is related to the gain tolerance ϵ.[4]

[4] Richard F. Baum, "Design of broad-band i.f. amplifier, Part II," *Jour. Appl. Phys.*, vol. 17, pp. 721–730; September, 1946.

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JOHAN HAANTJES

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For a biography and photograph of W. H. HUGGINS, see the September, 1947, issue of the PROCEEDINGS OF THE I.R.E.

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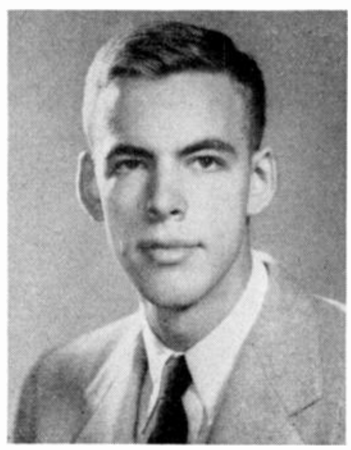

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Abstracts and References

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ACOUSTICS AND AUDIO FREQUENCIES

534.41 321
Industrial Stethescopes—(*Overseas Eng.*, vol. 20, pp. 388–389; July, 1947.) A commercial sound-detecting instrument for locating and comparing noises in machinery or detecting flow in or leakage from water mains, etc. Two forms are available, one for testing external sound when direct contact with the test part is impossible and the other for contact application.

534.75 322
Monaural Problems of Hearing—H. Mol. (*Tijdschr. ned. Radiogenoot.*, vol. 12, pp. 165–181; September, 1947. In Dutch with English summary.) Discussion of the action of the ear, and the way in which musical pitch is perceived.

534.78:621.395.61/.62 323
The Design of Speech Communication Systems—L. L. Beranek. (PROC. I.R.E., vol. 35, pp. 880–890; September, 1947.) "A method is presented for calculating the ability of a communication system to transmit speech intelligibly in the presence of noise."

621.395.61/.62 324
Proposed Method of Rating Microphones and Loudspeakers for Systems Use—F. F. Romanow and M. S. Hawley. (PROC. I.R.E., vol. 35, pp. 953–960; September, 1947.) The ratings are based on the ratios (in db) of the input, output and/or available powers. The overall sound rating of a system consisting of a microphone, input coupling network, amplifier, output coupling network, and loudspeaker is obtained by addition of the individual ratings.

621.395.623 325
Insertion Effects and Transfer Coefficients of Electroacoustic Systems—R. Stadlin. (*Tech. Mitt. Schweiz. Telegr.-Teleph.-Verw.*, vol. 25, pp. 133–143 and 187–194; August 1, and October 1, 1947. In German.)

621.395.623.75:518.4 326
The Design of Acoustic Exponential Horns—(*Electronic Eng.* (London), vol. 19, pp. 286–287; September, 1947.) Graphs from which the dimensions of horns of square or circular cross section may be determined for a given cutoff frequency.

621.395.623.8+621.395.97 327
Music Transmission System, with Individual Programme Selection, for Hospitals, Sanatoria, Hotels and Apartment Houses—E. Beutler. (*Tech. Mitt. Schweiz. Telegr.-Teleph.-Verw.*, vol. 25, pp. 194–199; October 1, 1947. In German.)

621.395.625.2+621.396.621+621.395.623.8 328
Portable Recorder-Player—J. G. Karnath. (*Radio Craft*, vol. 18, pp. 20–21, 65; August, 1947.) A combined recording unit, record player, radio receiver, and public-address sysystem. Records can be cut at either 33 and one third or 78 r.p.m. and all recordings may be monitored by a volume-level indicator on the control panel. Full circuit details are given.

621.395.625.3 329
Magnetic Tape Recorder for Movies and Radio—R. H. Ranger. (*Electronics*, vol. 20, pp. 99–103; October, 1947.) Response is flat within 4 db from 32 c.p.s. to 9.6 kc. Circuit diagrams are included and design considerations discussed. A device simplifying editing of the recording is also described. See also 20 of February.

AERIALS AND TRANSMISSION LINES

621.315.212 330
Broad-Band Noncontacting Short Circuits for Coaxial Lines: Part 1—TEM-Mode Characteristics—W. H. Huggins. (PROC. I.R.E., vol. 35, pp. 906–913; September, 1947.) Discussion of the design of an S-type noncontact plunger which will present an effective short-circuit to the TEM mode in a circular coaxial line over a frequency range of 3 to 1 or more. Equations showing the relation between the power loss and the physical shape of the plunger are derived and their application to a practical design is demonstrated.

621.315.212 331
The Capacity per Unit Length and Characteristic Impedance of Coaxial Cables with One Slightly Non-Circular Conductor—P. Parzen. (*Jour. Appl. Phys.*, vol. 18, pp. 774–776; August, 1947.)

621.392.029.64+621.317.763+621.396.616 332
The Transverse Electric Modes in Coaxial Cavities—R. A. Kirkman and M. Kline. (PROC. I.R.E., vol. 35, pp. 931–935; September, 1947.) Discussion on 888 of 1946.

621.392.029.64 333
The Propagation of Damped $H_{0,m}$ Waves near the Limiting Frequency—A. Käch. (*Helv. Phys. Acta*, vol. 20, pp. 341–356; August 4, 1947. In German.) Calculations of propagation in hollow metal conductors give incorrect results if losses in the tube walls are not taken into account. Formulas are derived, for tubes of rectangular cross section, which have general application and include a formula for the damping at the limiting frequency. Above and below this frequency the expressions agree with those given by other authors. See also 1328 of June (Kuhn).

621.392.029.64 334
Slow Transverse Magnetic Waves in Cylindrical Guides—G. G. Bruck and E. R. Wicher. (*Jour. Appl. Phys.*, vol. 18, pp. 766–769; August, 1947.) Because of the complex shapes given to waveguide walls by devices incorporated to reduce the phase velocity, calculations of the field pattern are inexact and complicated.

These calculations become easy and exact if, for theoretical purposes, the well-known equivalence of true and simulated dielectrics in producing a slow field is used.

Lining the guide walls with a natural dielectric might also prove to be a useful practical method of reducing the phase velocity.

621.392.029.64:621.315.68 335
Design of Simple Broad-Band Wave-Guide-to-Coaxial-Line Junctions—S. B. Cohn. (PROC. I.R.E., vol. 35, pp. 920–926; September, 1947.) Bandwidths greater than 2 to 1 with voltage s.w.r. of less than 2 are obtained. The theory of the designs and detailed mechanical drawings are given.

621.392.029.64:621.396.662.3 336
Lattice Filters for Decimetre Waves—Semenov. (*See* 387.)

621.392.1:621.3.012.2 337
Transmission-Line Calculations: Use of Impedance Circle Diagrams—W. C. Vaughan. (*Wireless Eng.*, vol. 24, pp. 314–322; November, 1947.) A simplified approach for the benefit of engineers with limited mathematical experience and mainly practical needs. Both the Car-

tesian and polar forms of circle diagram are considered. Proofs of constructions used are indicated; these assume little more than a knowledge of elementary geometry. A comprehensive account of circle diagrams was noted in 1031 of 1945 (Jackson and Huxley).

621.392.2:538.56 338
On the Electromagnetic Theory of the Lecher Line and Some Related Problems—J. Oswald. (*Onde Élec.*, vol. 27, pp. 330–340; August–September, 1947.) The problems of the Lecher line and the coaxial cable are identical and can be treated similarly. The only difference comes from the fact that, for the Lecher line, it is not possible to find a proper wave corresponding to a transverse electric or transverse magnetic solution. These waves of higher order do exist, however, as in the coaxial cable, but they are complex and do not constitute natural solutions of the type $F(x)G(y)$ exp $j(\omega t-\gamma z)$. For the coaxial cable, these higher order waves are of the type $J_n(K_r)$ cos $n\theta$ exp $j(\omega t-\gamma z)$.

On the other hand, the ordinary solution of type TEM is very simple and is conveniently treated by using the natural coordinates obtained by means of two systems of orthogonal circles. This system of coordinates is useful for all problems of electromagnetism where the limiting surfaces are circles, or arcs of circles, of the same system. Some examples of such problems are discussed, including the propagation along a cable with inner and outer conductors not concentric, and the radiation of a split cylinder.

If, as seems logical, the problem of the Lecher line is treated as a particular case of the general problem of guided waves, the line equations and constants, and the propagation constant, are obtained directly from the conditions at the limits, thus showing the close connection which exists between these quantities and the electromagnetic field.

621.392.43:621.317.3 339
Precision Measurement of Impedance Mismatches in Waveguide—Pomeroy. (*See* 468.)

621.396.67 340
An Electromagnetic Radiation Formula—G. Goudet. (*Onde Élec.*, vol. 27, pp. 313–317; August–September, 1947.) Asymptotic development of the formulas of Kottler gives a radiation formula which enables the effect produced at great distances by an aerial to be determined easily if the field distribution on a surface surrounding the aerial is known. The formula is particularly useful for centimeter λ aerials, horns, dielectric and slot aerials, etc. The formula also has applications in optical diffraction.

621.396.67 341
Recent Theories of the Aerial—É. Roubine. (*Rev. Tech. Comp. Franç. Thomson-Houston*, pp. 5–45; July, 1947.) Reprint of a series of articles in *Onde Élec.* See 2676 of 1947 and back references.

621.396.67 342
Circularly Polarized Antennas—W. Sichak and S. Milazzo. (*Elec. Commun.* (London), vol. 24, p. 273; June, 1947.) Summary of I.R.E. Convention paper. A formula is derived which gives the variation in received voltage as an elliptically polarized aerial is rotated in a plane transverse to the direction of propagation of an incident elliptically polarized wave. If elliptically polarized aerials are used, some signal will always be obtained, but two cases are noted in which there will be no signal with a circularly polarized aerial. Some methods of obtaining circular polarization are discussed. Experimental results confirm the theory.

621.396.67 343
Radiating Slit Systems—J. N. Feld. (*Compt. Rend. Acad. Sci.* (U.R.S.S.), vol. 53, pp. 615–618; September 10, 1946. In English.) Continuation of a paper on radiation through narrow slits in systems with axial symmetry, published in *Compt. Rend. Acad. Sci.* (U.R.S.S., vol. 51, no. 2; 1946. Here the potential difference between the edges of the slit is variable, and the slit is cut in the surface of a thin perfectly conducting ellipsoid of revolution. The field is excited from within by a linear conductor carrying a given current and shaped so that intense radiation through the slit occurs. Field equations are derived, and the relationship between slit aerials and the "dual" wire aerials is considered briefly. See also 2548 of 1946 and 1335 of 1947 (Booker).

621.396.67:621.396.97 344
The Broadcast Antenna—H. P. Williams. (*Jour. Brit. I.R.E.*, vol. 7, pp. 140–155; July–August, 1947. Discussion, pp. 155–156.) A discussion of the factors determining the radiation along the ground from medium-wave aerials and their fade-free radius of transmission.

The possibility of using slot aerials in the form of trenches or as a suspended system of wires is mentioned.

621.396.67.029.62 345
An Antenna that Multiplies by 50—J. A. Kmosko. (*QST*, vol. 31, pp. 50–53; September, 1947.) An array of 12 λ/2 elements fed in phase, six broadside and two high, with 12 reflectors spaced λ/4 behind the driven elements. Impedance and gain measurements are described and the beam pattern is given. The array is suitable for 144 Mc. or higher frequencies.

621.396.671 346
Method of Determining the Characteristic Reactance of Thin Aerials—M. L. Levin. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 11, no. 2, pp. 117–133; 1947. In Russian.)

621.396.671:621.396.611.33 347
The Matching Ranges of Transmitters—Mourmant. (*See* 595.)

621.396.674:621.314.2 348
A Note on Coupling Transformers for Loop Antennas—Kobilsky. (*See* 354.)

621.396.679.4† 349
On the Efficiency of H.F. Feeders—A. R. Vol'pert. (*Radiotekhnika*, (Moscow), vol. 2, pp. 22–24; September–October, 1947. In Russian.)

621.392.029.64 350
The Principles and Practice of Wave Guides [Book Review]—L. G. H. Huxley. Cambridge University Press, London, 328 pp., 21s. (*Wireless Eng.*, vol. 24, p. 344; November, 1947.) The first of a new Cambridge series on "Modern Radio Technique," edited by J. A. Ratcliffe. It is based on courses given by the author at the Radar School of the Telecommunications Research Establishment, and provides an introduction to the great practical war-time developments in the use of waveguides.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.012.2:621.392.1 351
Transmission-Line Calculations: Use of Impedance Circle Diagram—Vaughan. (*See* 337.)

621.3.015.3:517.36 352
Oscillations and Transient Phenomena. Their Study by Means of the Laplace and Cauchy Transformations—Bouthillon. (*See* 458.)

621.3.032.24:621.316.722 353
Diode Contact Potential for Negative Bias—H. T. Sterling. (*Electronics*, vol. 20, pp. 164, 172; October, 1947.) Discusses the use of contact potential as a source of bias in high-gain audio amplifiers, in the r.f. and i.f. stages of receivers and in automatic gain control circuits.

621.314.2:621.396.674 354
A Note on Coupling Transformers for Loop Antennas—M. J. Kobilsky. (Proc. I.R.E., vol. 35, pp. 969–973; September, 1947.) A theoretical analysis. For optimum signal-to-noise ratio, the loop inductance should equal the primary inductance; general expressions for sensitivity, gain, and selectivity are derived for this optimum condition assuming that circuit noise limits the sensitivity. See also 1683 of 1943 (Levy) and 2590 of 1944 (Bond).

621.316.727 355
Compensation of Phase Shift at Low Frequencies—F. McGee. (*Elec. Commun.* (London), vcl. 24, pp. 270–271; June, 1947.) Summary of I.R.E. Convention paper. Discussion of a method for simultaneous correction of phase errors in the cathode, screen, and coupling circuits.

621.316.86.001.8 356
The Thermistor in Biological Research—Andrew. (*See* 500.)

621.318.572 357
Electronic Switching—E.M.I. Laboratories—(*Electronic Eng.* (London), vol. 19, p. 282; September, 1947.) A sensitive cathode-coupled trigger circuit which is little affected by changes in tube characteristics, values of components, or supply voltages.

621.385.832:621.397.6 358
Magnetic-Deflection Circuits for Cathode-Ray Tubes—Schade. (*See* 574.)

621.392+621.385.1:621.396.694.012.8 359
Circuits and Valves in Electronics—R. Charbonnier and J. Royer. (*Télév. Franç.*, Supplement *Électronique*, pp. 29–32; September, 1947.) The first of a series of special articles. The following subjects are considered briefly:—(a) internal impedance of a complex system; (b) transfer impedance of a quadripole; (c) tube equivalent circuits; extension to pentode; (d) tube impedance.

621.396.611 360
On the Theory of Oscillators with a Bridge Oscillating Circuit—K. F. Teodorchik. (*Radiotekhnika*, (Moscow), vol. 2, pp. 3–7; September–October, 1947. In Russian.)

621.396.611.1 361
On the Interaction of Two Oscillators—B. N. Gorozhankin. (*Bull. Acad. Sci.* (U.R.S.S.) *sér. phys.*, vol. 11, no. 2, pp. 147–154; 1947. In Russian, with English summary.) It is shown, both theoretically and experimentally, that in the case of two identical oscillators separated by a distance great in comparison with λ, the synchronous frequency has a series of jumps when the distance between the oscillators is varied. The distance between the points where these jumps occur is approximately λ/2. With increase of separation, the frequency gradually diminishes, suddenly rises, then falls again, and so on, with the reverse for decrease of separation.

621.396.611.1 362
Free Oscillations of a Resonant Circuit with Nonlinear Self-Inductance—H. Miedema. (*Tijdschr. ned. Radiogenoot.*, vol. 12, pp. 155–163; September, 1947. In Dutch, with English summary.) Two particular solutions of the differential equation for such a circuit are given. The solution obtained for an undamped circuit also gives a good approximation to the true waveform with small amplitudes, and shows the relationship between frequency and amplitude.

621.396.611.21 363
The Series Trimming of Crystal Resonators—M. P. Johnson. (*Electronic Eng.*, vol. 19, pp. 281–282; September, 1947.) Formulas are derived for the series inductance or capacitance required for a given deviation from the crystal resonant frequency. The effect of temperature changes on the combination is considered. A

particular crystal was trimmed ±100 parts in 10^6 with little degradation of oscillator performance.

621.396.611.3 364

Coupled Circuits—B. D. H. Tellegen. (*Philips Res. Rep.*, vol. 2, pp. 1–19; February, 1947.) In the determination of the frequencies and dampings of the free oscillations, the theory of coupled circuits leads to a quartic equation. For resonance curves of small relative width, the quartic can be reduced to a quadratic equation which can easily be resolved into factors, each of which determines one of the free oscillations. Coupling factors are determined for oscillatory circuits coupled by inductance, capacitance, and resistance and the cases of circuits having dissimilar damping or tuning are considered.

"Systems are considered in which the circuits are coupled over an arbitrary four-terminal network which may also contain an amplifying valve."

621.396.611.3 365

General Calculation of a Regular Impedance Network. Application to Systems of Coupled Oscillatory Circuits—P. Fajon. (*Rev. Gén. Élec.*, vol. 56, pp. 377–391; September, 1947.) A general method is given for calculating the properties of a series of identical coupled circuits. This is applied to filters made up of (a) undamped and (b) damped circuits. Systems of coupled oscillatory circuits are also treated by a simple method, applicable to any number of circuits; the cases of e.m. and of capacitive coupling are particularly considered.

621.396.611.4 366

Some Results on Cylindrical Cavity Resonators—J. P. Kinzer and I. G. Wilson. (*Bell Sys. Tech. Jour.*, vol. 26, pp. 410–445; July, 1947.) "Certain hitherto unpublished theoretical results on cylindrical cavity resonators are derived. These are: an approximation formula for the total number of resonances in a circular cylinder; conditions to yield the minimum volume circular cylinder for an assigned Q; limitation of the frequency range of a tunable circular cylinder as set by ambiguity; resonant frequencies of the elliptic cylinder; resonant frequencies and Q of a coaxial resonator in its higher modes; and a brief discussion of fins in a circular cylinder.

The essential results are condensed in a number of new tables and graphs."

A bibliography of 89 items is included.

621.396.615:621.316.726.078.3 367

The Impulse Synchronized Oscillator and Its Applications—E. H. Hugenholtz. (*Tijdschr. ned. Radiogenoot.*, vol. 12, pp. 89–110; May, 1947. Discussion, pp. 111–112. In Dutch, with English summary.) Description of various systems in which frequencies can be obtained with crystal stability and accuracy by using this type of oscillator. Such oscillators permit selective high-ratio frequency multiplication or division; their principles and limitations are discussed, with particular reference to transmitters. A brief comparison is made with analogous systems.

621.396.615.029.3 368

A Resistance-Tuned Frequency-Modulated Oscillator for Audio-Frequency Applications—H. S. McGaughan and C. B. Leslie. (Proc. I.R.E., vol. 35, pp. 974–978; September, 1947.) The oscillator consists of a *RC* amplifier with two feedback circuits. Wide variation in frequency is obtained without excessive a.m. or harmonic distortion.

621.396.615.029.5 369

Practical Construction of a Beat-Frequency Oscillator—R. Aschen and R. Zahl. (*T.S.F. Pour Tous*, vol. 23, pp. 183–185; September, 1947.) Complete circuit details. Practical constructional details are given by R. Zahl in *T.S.F. Pour Tous*, vol. 23, pp. 213–215; October, 1947. See also 3827 of January (Aschen and Lafargue) and back references.

621.396.615.142.2 370

Reflex Oscillators—Pierce and Shepherd. (*See* 617.)

621.396.615.17:621.314.632 371

Variable Time Constant—T. Petrides. (*Electronics*, vol. 20, p. 138; October, 1947.) Crystal rectifiers as grid resistors in a multivibrator provide a variable time-constant, with a resultant square-wave output.

621.396.616+621.392.029.64+621.317.763 372

The Transverse Electric Modes in Coaxial Cavities—R. A. Kirkman and M. Kline. (Proc. I.R.E., vol. 35, pp. 931–935; Sepetmber, 1947.) Discussion on 888 of 1946.

621.396/.397].645 373

Bandwidth and Speed of Build-Up as Performance Criteria for Pulse and Television Amplifiers—D. G. Tucker. (*Jour. I.E.E.* (London), part I, vol. 94, pp. 382–383; August, 1947.) Summary of 3062 of 1947.

621.396.645:518.4 374

A Method of Graphically Analyzing Cathode-Degenerated Amplifier Stages—E. M. Lonsdale and W. F. Main. (Proc. I.R.E., vol. 35, pp. 981–984; September, 1947.) The method is based on the use of a curve relating grid-to-ground potential to the resulting anode current, taking into account the effects of the voltage drops across cathode and anode resistors, but neglecting reactive effects.

621.396.645:621.396.615 375

The Cathode-Follower V.F.O.—W. M. Scherer. (*CQ*, vol. 3, pp. 15–18, 84; September, 1947.) Combines crystal stability with easy frequency adjustment. Full circuit and constructional details and component ratings are given.

621.396.645:621.396.621 376

Intermediate-Frequency Amplifiers for Frequency-Modulation Receivers—Adams. (*See* 527.)

621.396.645.029.42/.52 377

General Purpose Portable Amplifier—C. R. Smitley. (*Electronics*, vol. 20, pp. 150, 160; October, 1947.) Voltage gain can be 20 db or 40 db, and is flat to within 0.5 db over the frequency range 1 to 100 kc. Equivalent noise input level is at least 90 db below 1 volt. The amplifier is operated from a 115-volt a.c. line. The cathode follower can be connected to input or output. Full circuit details are given.

621.396.645.029.62 378

Broad-Band Very-High-Frequency Amplifiers—A. M. Levine and M. G. Hollabaugh. (*Elec. Commun.*, vol. 24, pp. 269–270; June, 1947.) Summary of I.R.E. Convention paper. For another summary see 3479 of 1947.

621.396.645.029.64 379

On the Theory of U.H.F. Amplifiers—V. I. Siforov. (*Radiotekhnika* (Moscow), vol. 2, pp. 3–21; July–August, 1947. In Russian.) A theory of single and multistage amplifiers is developed in which the tubes and the intertube circuits are treated as active and passive 4-terminal networks respectively. Primary, secondary, and characteristic parameters of a tube are introduced and the relationship between them is investigated. From the knowledge of these parameters, the maximum amplification factor of a multistage amplifier can be determined and various elements of the system co-ordinated. General design formulas for obtaining maximum amplification with pentodes, grounded-grid triodes and other types of tubes are derived and the conditions necessary for the absence of parasitic oscillations in a multistage amplifier are established.

621.396.645.35 380

High Gain D.C. Amplifier—W. G. Shepard. (*Electronics*, vol. 20, p. 138, 182; October, 1947.) Circuit diagrams and component ratings are given. Stages are directly coupled; each has a separate power supply. The circuit has low drift, frequency response is flat to 50 kc., and phase shift negligible to 20 kc.

621.396.645.35 381

D.C. Amplifier for Low-Level Signals—C. B Aiken and W. C. Welz. (*Electronics*, vol. 20, pp. 124–128, 130; October, 1947.) To amplify signals below 1 microvolt, the circuit noise level is reduced by the use of input impedances not exceeding 20 ohms, bandwidths of only a few c.p.s. and averaging rectifiers.

621.396.645.371 382

Video-Frequency Negative-Feedback Amplifiers—M. G. Hollabaugh, J. A. Rado, and A. M. Levine. (*Elec. Commun.* (London), vol. 24, pp. 272–273; June, 1947.) Summary of I.R.E. Convention paper. For another summary see 3483 of 1947; see also 3517 of 1939 (Wheeler).

621.396.645.371:621-526 383

Impedances in the Amplifier with Counter-Reaction—P. M. Prache. (*Bull. Soc. Franç. Élec.*, vol. 7, pp. 515–528; September, 1947.) The amplifier with counter-reaction is treated as a servomechanism. The conditions for optimum operation of such amplifiers are determined theoretically.

621.396.662.3 384

Insertion Loss and Effective Phase Shift in Composite Filters at Cut-Off Frequencies—V. Belevitch. (*Elec. Commun.* (London), vol. 24, pp. 192–194; June, 1947.) Formulas are derived for the insertion loss in decibles and for the effective phase shift in radians at cutoff frequency for both low-pass and band-pass filters. The usual methods of estimating insertion losses fail at the cutoff frequencies; by these derived formulas, however, total losses are obtained by direct calculation.

621.396.662.3 385

Infinite-Rejection Filters—A. M. Stone and J. L. Lawson. (*Jour. Appl. Phys.*, vol. 18, pp. 691–703; August, 1947.) Analyses of several types of bridged-T filter structures show that they can be developed into the symmetrical lattice form, which is itself equivalent to the usual 4-arm bridge.

An expression is derived relating the circuit constants and frequency of the filter and the ratio of the output power passed by the filter to the power in the absence of the filter. In theory, it is possible to obtain infinite attenuation at a given frequency while retaining essentially the same bandwidth as that for the uncompensated filter. Certain bridged-T filter structures may be adapted to meet the requirements of distributed parameters circuits, such as u.h.f. lines and u.h.f. resonant cavities. Their usefulness is discussed.

A model has been constructed which has a bandwidth of one half Mc. at 3000 Mc. with a attenuation at resonant frequency of over 70 db, compared with 20 db for a similar uncompensated filter. The distortion produced by a u.h.f. filter of this type, in otherwise rectangular pulses of short duration, has been investigated both theoretically and experimentally, and curves are presented which show the resultant waveform as a function of the tuning of the filter.

621.396.662.3 386

Optimum Resistive Terminations for Single-Section Constant-K Ladder-Type Filters—L. J. Giacoletto. (*RCA Rev.*, vol. 8, pp. 460–479; September, 1947.) "The operation of a single non-dissipation section of a ladder-type constant-K filter terminated in a resistance is considered. It is found that depending upon the value of the terminating resistance in proportion to a filter-design parameter, different filter characteristics are obtained. Optimum values for the resistive termination are determined for different operating characteristics and different filter sections. It is found that the T-filter section has somewhat better operating characteristics than the π-filter section."

621.396.662.3:621.392.029.64 387
Lattice Filters for Decimetre Waves—V. F. Semenov. (*Radiotekhnika* (Moscow), vol. 2, pp. 44–47; July–August, 1947. In Russian.) A preliminary report on experiments with filters for H_1 waves in a cylindrical waveguide. The possibility of using such filters for tuning open-ended waveguides was also investigated.

621.396.662.3:621.396.611.21 388
Filter Crystals with Low Self-Inductance—J. J. Vormer. (*Tijdschr. ned. Radiogenoot.*, vol. 12, pp. 1–6; January, 1947. In Dutch, with English summary.) A tenfold reduction in inductance can be obtained with a $-18.5°$ rotated X-cut crystal by exciting a harmonic of the Y' wave, using a series of pairs of electrodes, alternate electrodes on the two faces of the crystal being connected in parallel. The ratio of the crystal dimensions in the z and y directions must exceed unity; a value near 1.8 minimizes spurious frequencies. Corrections *ibid.*, vol. 12, no. 2, insert March, 1947.

621.396.662.6 389
Tuning Without Condensers—F. E. Berhley. (*FM and Telev.*, vol. 7, pp. 35–37; August, 1947.) High sensitivity and effective static reduction are achieved by means of a tuned r.f. stage, one limiter, and a discriminator. The r.f. stage has three high-Q resonant lines but no variable capacitors.

GENERAL PHYSICS

53.08+621.317 390
Fundamentals of Measurement Technique—J. Hartmann. (*Rev. Sci.* (Paris), vol. 85, pp. 323–334; April 1, 1947.)

530.145.65 391
The Problem of the Polarization of the de Broglie Waves associated with Electrons—J. Brenet. (*Rev. Sci.* (Paris), vol. 85, pp. 357–359; April 1, 1947.) The phenomena of polarization of electronic waves should be examined with reference to the fundamental characteristics of the electron, essentially a simple particle in the Dirac sense, and of the photon, a particle obtained by the fusion of two corpuscles which are complementary in the Dirac sense. There appears to be no contradiction between the existence of diffraction phenomena for both photons and electrons, and the absence of polarization for electronic waves, since a state of polarization cannot be defined for a Fermi-Dirac elementary corpuscle. This will be discussed in a later paper.

534.13:539.21.001.572 392
Model to Demonstrate Oscillations in a Molecule Subjected to an Electric Field—A. J. Maddock. (*Jour. Sci. Instr.*, vol. 24, pp. 230–232; September, 1947.) The mechanical model demonstrates dipolar, atomic, and electronic relaxation oscillations occurring in the molecule, and the contribution to the total dielectric constant of each of the resulting polarizations.

535.37 393
The Light Emission from Fluorescent Screens Irradiated by X-Rays—Klasens. (*See* 437.)

536.33+536.24]:535.37 394
Radiation and Heat Conduction in Light-Scattering Material—H. C. Hamaker. (*Philips Res. Rep.*, vol. 2, pp. 55–67; February, 1947.) "On the basis of a set of simultaneous differential equations, originally due to Schuster, the transmission and reflection of light in light-scattering layers is discussed. Formulas previously developed by Kubelka and Munk are recapitulated briefly; they are extended so as to describe the luminescence of fluorescent screens excited by X-rays or electron bombardment. Likewise formulas are derived that include temperature radiation."

537.122 395
The Electron Jubilee—(*Elec. Times*, vol. 112, pp. 389–390; October 5, 1947.) A series of lectures arranged to celebrate J. J. Thomson's discovery of the electron. History and Early Development of the Electron, by J. A. Crowther. Electrons in Modern Theoretical Physics, by R. E. Peierls. The Electron Liberated, by C. C. Paterson. The Electron in Research, by G. Thomson. Electrons in Industry, by T. E. Allibone. See also *Jour. I.E.E.* (London), part I, vol. 94, p. 339; August, 1947; *Engineer* (London), vol. 184, p. 315; October 3, 1947; and *Electrician*, vol. 139, pp. 921–923; September 26, 1947.

537.291+538.691]:621.385.832 396
Electron Beam Deflection: Part 1—Small-Angle Deflection Theory—Hutter. (*See* 609.)

537.5 397
Discharge through Gases—L. B. Loeb. (*Science*, vol. 106, pp. 229–236; September 12, 1947.) The 38th Kelvin Lecture. A historical review, with bibliography of 90 items. See also 3854 of January.

537.525 398
On the Growth, Reaction Mechanism and Stability of Low-Current, Low-Pressure Discharges—H. Luz von Gugelberg. (*Helv. Phys. Acta*, vol. 20, pp. 307–340; August 4, 1947. In German.) Results and discussion of measurements of the development of glow discharges in He, Ar, Kr, XE, N_2, H_2, and in Ne-Ar mixtures.

537.527:536.5 399
On the Excitation Temperature, the Gas Temperature, and the Electron Temperature in the High-Pressure Mercury Discharge—W. Elenbaas. (*Philips Res. Rep.*, vol. 2, pp. 20–41; February, 1947.)

538.311 400
The Production of a Uniform Magnetic Field Over a Specific Volume by Means of Twin Conducting Circular Coils—H. Craig. (*Proc. Phys. Soc.*, vol. 59, pp. 804–814; September 1, 1947.) An investigation of the uniformity of axial magnetic field attainable on the central plane between twin parallel coaxial circular coils of given diameter arranged in conjunction. The optimum coil separation and the highest degree of uniformity of the field attainable over a given circular area are derived graphically. Improved uniformity can be obtained over an annular area. Volumes of cylindrical shape are also considered, and regions of remarkably constant axial field are shown. "The best ratio of radial depth to axial length of the coils is determined, and a method is given by which the equivalent separation of parallel coils of finite cross-section can be found very exactly by means of a search coil of particular shape."

538.541:538.221 401
On the Theory of Eddy Currents in Ferromagnetic Materials—H. B. G. Casimir. (*Philips Res. Rep.*, vol. 2, pp. 42–53; February, 1947.) The theory is developed for the limiting case where the depth of penetration d is small although the product μd is not necessarily small.

First the rigorous solution for a sphere of radius R is discussed for this limiting case. Next it is shown that the solution can be obtained from Laplace's equation with a new type of boundary condition. This boundary condition is then applied to a discussion of eddy currents in spheroids. Explicit formulas are found both for small and for very large values of $\mu d/R$.

Special attention is given to the limits for very long and very flat spheroids, respectively.

538.56:621.316.7.078 402
Radiophysics and the Theory of Automatic Control—G. S. Gorelik. (*Bull. Acad. Sci.* (U.R.S.S.), *sér phys.*, vol. 11, No. 2, pp. 103–115; 1947. In Russian, with English summary.) A review stressing the similarity between problems on the origin of self-oscillations in radiophysics and in automatic control.

538.56:621.392.2 403
On the Electromagnetic Theory of the Lecher Line and Some Related Problems—Oswald. (*See* 338.)

538.569.4+621.396.61.029.64 404
On the Emission of Microwaves and Their Absorption in the Air—Ginsburg. (*See* 594.)

538.569.4.029.64 405
A New Electronic System for Detecting Microwave Spectra—W. Gordy and M. Kessler. (*Phys. Rev.*, vol. 72, p. 644; October 1, 1947.) A single-crystal system using a l.f. cutoff filter and a modulation technique which allows the signal to be amplified at higher frequencies.

538.569.4.029.64:546.171.1 406
Saturation Effect in Microwave Spectrum of Ammonia—W. V. Smith and R. L. Carter. (*Phys. Rev.*, vol. 72, pp. 638–639; October 1, 1947.) Comment on 1399 of 1947 (Townes). The magnitude of intensity saturation of the 3,3 line here found is consistent with Townes' results, but there is no broadening of the line at low pressures. Reasons for this discrepancy are suggested. See also 3870 of January (Bleaney and Penrose).

538.569.4.029.64:546.171.1 407
The Ammonia Spectrum and Line Shapes Near 1.25 cm Wave-Length—C. H. Townes. (*Bell Sys. Tech. Jour.*, vol. 26, p. 689; July, 1947.) Summary of 1399 of 1947, U.D.C. of which should read as above.

538.652:546.74 408
Experimental Facts concerning the Magnetostriction of Nickel—Y. Rocard. (*Rev. Sci.* (Paris), vol. 85, pp. 195–204; February 15, 1947.) A review of data from many sources, with a short account of some applications.

538.691 409
Magnetic Focusing Between Inclined Plane Pole-Faces—H. O. W. Richardson. (*Proc. Phys. Soc.*, vol. 59, pp. 791–804; September 1, 1947.) The resultant magnetic field, if applied to β-ray spectroscopy, is said to give high dispersion and resolving power, together with a fair solid angle of collection Ω. With planes bevelled to become parallel at their closest parts, both lateral and longitudinal focusing should occur. Approximate calculation shows that the beveling makes high values of Ω possible but may result in lower resolving power.

539.3:518.3 410
Nomographic Representation of the Elastic Contact Conditions between Steel Pivot and Sapphire Jewel—G. F. Tagg. (*Jour. Sci. Instr.*, vol. 24, pp. 244–248; September, 1947.)

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.53:621.396.96 411
Radar Observations of Meteors—J. S. Hey and G. S. Stewart. (*Proc. Phys. Soc.*, vol. 59, pp. 858–883; September 1, 1947.) A comprehensive account of observations made on short-duration echoes at λ 4 to 5 meters near the E region, using British Army radar equip-

ment. The echoes came from a height of about 95 kilometers; the analysis indicates a close correlation between these echoes and meteors.

523.72.029.5:523.746 412

Solar Radiation at Radio Frequencies and Its Relation to Sunspots—L. L. McCready, J. L. Pawsey, and R. Payne-Scott. (*Proc. Roy. Soc. A*, vol. 190, pp. 357–375; August 12, 1947.) Experimental studies on a frequency of 200 Mc. are described. The radiation has characteristics similar to those of thermal radiation from the photosphere but always exceeds expected values by a factor of the order of 10^2 to 10^4. Day-to-day intensity variations show correlation with sunspot variations. Rapid intensity fluctuations or "bursts" received simultaneously at widely spaced points are presumably solar in origin. Directional observations indicate that the radiation originates in areas in the immediate vicinity of a sunspot group. Radiation from gross electrical discharges is suggested as the cause of values of received intensity higher than could be produced by thermal radiation.

523.745+523.854]:621.396.822:551.510.535 413

Solar and Terrestrial Radio Disturbances—J. S. Hey, S. J. Parsons, and J. W. Phillips. (*Nature* (London,) vol. 160, pp. 371–372; September 13, 1947.) Continuous recording of galactic radio emission for λ 12 meters approximately shows a significant increase of absorption by the enhanced D-layer ionization during periods of solar activity. This effect cannot easily be distinguished if a burst of intense solar radio emission takes place simultaneously. Two examples are discussed; for one the solar radio emissions were small; for the other the absorption preceded a solar radio burst.

523.75 414

Electromagnetic Forces in Solar Prominences—D. S. Evans. (*Mon. Not. R. Astr. Soc.*, vol. 106, no. 4, pp. 300–337; 1947.) The forms of solar prominences and the motion of associated knots may be explained in terms of the motion of ions in a magnetic field. In section 1, the dispersal of systems of ions under their mutual electrostatic repulsion is discussed. Agreement between computed and observed velocities (of knots) is obtained by assuming the existence of a background cloud of 10^{-11} or 10^{-12} elementary charges per centimeter^3. In section 2, the support of prominences and of highly ionized atoms in the corona is considered in relation to the possible existence of a general electrostatic field above the solar surface. Calculated space tracks of moving knots are found to be consistent with the observed forms of streamers. The forms of tornado and coronal prominences are also considered.

537.212+538.12]:521.15 415

Stellar Electromagnetic Fields—L. Davis, Jr. (*Phys. Rev.*, vol. 72, pp. 632–633; October 1, 1947. Formulas are obtained for the magnetic induction, potential, and electrostatic field outside a star, and rough values of these quantities for the earth, the sun, and two stars are tabulated. The limitations and implications of these formulas are discussed. See also 3891 of January (Babcock) and back references.

537.591.15 416

Cosmic-Ray Bursts and Shower Spread under Large Thicknesses of Lead—J. W. F. Juritz and C. B. O. Mohr. (*Proc. Roy. Soc. A*, vol. 190, pp. 426–434; August 12, 1947.

551.510.535 417

Evolution of Views on the Structure of the Ionosphere—V. N. Kessenikh. (*Bull. Acad. Sci.* (U. R. S. S.), *sér. phys.*, vol. 11, no. 2, pp. 155–163; 1947.) In Russian, with English summary.) A historical survey. The necessity of extending the network of ionosphere observations is stressed.

551.510.535:546.21 418

Distribution of Molecular and Atomic Oxygen in the Upper Atmosphere—H. Rakshit. (*Indian Jour. Phys.*, vol. 21, pp. 57–68; April, 1947.) The distribution is calculated by adapting Pannekoek's method of studying the effect of solar ultraviolet radiation on atmospheric ionization; it is essentially the method of Majumdar (3121 of 1938) but Majumdar's results were based on data not corroborated by recent observations. The density of O_2 molecules decreases rapidly with height above 100 kilometers; that of O atoms is almost zero at 80 kilometers, increases rapidly to a maximum at 105 kilometers and then gradually decreases.

551.510.535:621.396,11 419

The Forecasting of Ionosphere Critical Frequencies—K. Rawer. (*Rev. Sci*, (Paris), vol. 85, pp. 234–235; February 15, 1947.) A single quantity Q, representing the aggregate of all the critical frequencies measured during a month, is defined by the equation

$$Q = 1/24 \sum_{t=00}^{t=23} f^2(t)$$

where $f(t)$ is the critical frequency of the ordinary ray at hour t. The values of Q vary with place, season, and solar activity, but for a particular month and station, they furnish an excellent means of comparison between different years. For example, the mean monthly curve for January, 1944, is derived directly from that for January, 1942, by multiplying by the square root of the ratio of the corresponding Q values. When the Q values for preceding years are known for any station, the variations corresponding to the 11-year solar cycle can be derived. This is illustrated from the data published by Washington. The method has, hitherto, only been verified experimentally for the F_2 layer. It does not appear to be applicable to the sporadic-E layer.

551.510.535:621.396.11 420

One of the Reasons for a Change of Amplitude of a Single Pulse reflected from the Ionosphere—V. D. Gusev. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 11, no. 2, pp. 195–201; 1947. In Russian, with English summary.) An examination of problems related to the inhomogeneous structure of the E layer. Electron clouds whose dimensions exceed the wavelengths in question are postulated. Diffraction due to the inhomogeneities in the E layer is considered and shown to result in fading and in abnormally large values of the reflection coefficient in the case of reflection from the E or the F_2 layer. These conclusions were confirmed by continuous measurements of the reflecteon coefficient for vertical transmissions on frequencies of 3 and 4 Mc.

551.594.5:551.510.535:621.396.11 421

Ionospheric Perturbations in the Zone of Polar Auroras—K. Rawer. (*Rev. Sci.* (Paris), vol. 85, pp. 287–288; March 1–15, 1947.) Discussion of absorption effects shows that frequently in the auroral zone there is an ionosphere layer, at a height of about 100 kilometers, which is due to solar corpuscular radiation and which gives good radio reflections, often for very high frequencies. This accounts for the fact that when perturbations prevent communication by the normal frequencies and operation on lower frequencies is equally bad, it is often possible to maintain communication by using considerably higher frequencies. The development of perturbations is described and their seasonal effects are discussed.

551.594.5:551.510.535:621.396.822 422

Radio Echoes from the Aurora Borealis—A. C. B. Lovell, J. A. Clegg, and C. D. Ellyett. (*Nature* (London), vol. 160, p. 372; September 13, 1947.) Recently, radio echoes were obtained at 460-kilometer range, arising apparently from a luminescent cloud near the zenith which appeared and disappeared with the echo. Simultaneously with the appearance of the echoes, a noise level increase was observed on both 72 and 46 Mc. over the range 450 to 600 kilometers. The aurora echo was observed on 46 Mc. but not on 72 Mc., and assuming it was of the ionospheric type, its ionization density was calculated to be about 100 times the normal F-region ionization density during the night.

LOCATION AND AIDS TO NAVIGATION

621.396.663 423

A New British Radio Compass—(*Electronic Eng.*, vol. 19, p. 325; October, 1947.) A flattened loop aerial is housed either in a shallow blister or within the aircraft body. In automatic operation the loop, when tuned to signals from a radio beacon, sets itself in the position of minimum signal and indicators operated by a "Desynn" servo system give the correct bearing. The loop can also be continuously rotated at a variable speed and the minimum-signal position determined aurally. Accuracy is within 1°).

621.396.93+621.396.663 424

Investigation of Errors in Spaced-Collector Direction-Finder Systems—T. H. Clark. (*Elec. Commun.* (London), vol. 24, pp. 199–207; June, 1947.) Site errors are classified but not investigated. Errors resulting from the design of one directional pair are considered for a wave arriving in the horizontal plane. These are treated mathematically and general conclusions, applicable as specific design principles, are drawn. Errors due to the combination of two directional pairs are investigated generally as are errors resulting from the use of long cables connecting the aerials to the goniometer. Finally, possible errors due to faulty alignment of the mechanical parts of the goniometer are mentioned.

621.396.932:621.396.96 425

Peacetime Radar—B. B. Talley. (*Electronics*, vol. 20, pp. 113–115; October, 1947.) A description of radar equipment installed on river boats for navigation in fog and a survey of results obtained, illustrated by photographs of actual plan-position-indicator images.

621.396.933+621.396.96 426

Developments in Airline Radio and Radar Communications and Navigational Facilities: Parts 1 and 2—H. J. Brown. (Proc. I.R.E. (Australia), vol. 8, pp. 4–9 and 4–15; August and September, 1947.) In part 1, improvements in communication between air and ground are considered. The use of ionospheric predictions for the correct choice of frequency to suit season and time of day is discussed. The results of research on precipitation static interference are considered. In Part 2, radio and radar navigational facilities discussed include loran, consol and other long-range systems, omnidirectional ranges of the C.A.A. type, pulse-type multitrack ranges and radar distance-measuring equipment. Developments in airport control radar are also considered. The economic advantages of instrument-landing systems for airlines are stressed, and possible future developments are mentioned briefly.

621.396.933 427

Status of V.H.F. Facilities for Aviation—P. Caporale. (*Electronics*, vol. 20, pp. 90–95; October, 1947.) Details of the v.h.f. omnidirectional radio range now being installed throughout the United States by the Civil Aeronautics Administration for short-range air navigation, and operating principles of the instrument-landing system and phase-comparison localizer. See also 2388 of 1942, 2655 of 1945 (Luck) and back references.

621.396.933 428
First Tests on Navar System for Aerial Navigation and Air Traffic Control—P. R. Adams, S. H. M. Dodington, and J. A. Herbst. (*Elec. Commun.* (London) vol. 24, pp. 263–264; June, 1947.) Summary of I.R.E. Convention paper. A description is given of equipment and its performance in an experimental ground and aircraft installation, assembled for preliminary tests, and of the gathering of flight data on (a) a pulse-type airborne distance indicator with meter presentation, (b) an airborne azimuth indicator with meter presentation, (c) a ground search radar with plan-position indicator (d) assisted radar, using an airborne responder, with presentation on the same plan-position indicator.

621.396.933 429
Survey of Radio Navigational Aids—R. I. Colin. (*Elec. Commun.* (London), vol. 24, pp. 219–261; June, 1947.)Radio navigational aids, including the earliest, are classified into four basic types, and their fundamental principles, characteristics, and ambiguities are discussed. General air navigational requirements, including those of the radio altimeter, are mentioned, and the methods of fulfilling them described. A comprehensive list of basic radio navigational systems is given (p. 243).

621.396.933 430
Relations between Bandwidth, Speed of Indication, and Signal-to-Noise Ratio in Radio Navigation and Direction Finding.—H. Busignies and M. Dishal. (*Elec. Commun.* (London), vol. 24, pp. 264–265; June, 1947.) Summary of I.R.E. Convention paper. For signal-to-noise ratios of the order of 3 to 1, the signal required for a given signal-to-noise ratio is proportional to the square root of the bandwidth ratio for pre-detection narrowing and to its fourth root for post-detection narrowing. A possible new bucking detector method for reproducing signals at very low signal-to-impulse-noise ratios is described. The required speeds of indication for various aids to navigation are considered, and it is suggested that unnecessarily wide bandwidth is used in some of them. The navaglobe narrow-band automatic direction-finding system, which has 20 c.p.s. pass bandwidth, is described.

621.396.96 431
Technique and Evolution of Radar: Parts 1-4—Demanche. (*Onde Élec.*, vol. 27, pp. 173–183, 244–258, 292–304; 341–356; May–September, 1947.) Part 1 deals with the radar equation and the choice of fundamental parameters, part 2 with transmitting and receiving apparatus, part 3 with the many different types of aerials used and part 4 with indicating apparatus and methods. To be continued.

621.396.96 432
The Maximum Range of a Radar Set—K. A. Norton and A. C. Omberg. (PROC. I.R.E., vol. 35, pp. 927–931; September, 1947.) Discussion on 2130 of 1947.

621.396.96:551.515.43 433
Radar Storm Detection—R. Wexler and D. M. Swingle. (*Bull. Amer. Met. Soc.*, vol. 28, pp. 159–167; April, 1947. Reprint.) Basic radar theory is discussed, assuming it to be unfamiliar to meteorologists, and with particular reference to the work of Ryde (515 of 1947). The power received from rainstorms, assuming total interception of the beam by the rainstorm area, varies as $\Sigma a^6)/R^2\lambda^4$, where a is the radius of a typical drop and R is the range. Moderate rain reflects about 10^6 times as effectively as rainless clouds. Typical attenuation values for absorption by oxygen and by water vapor and for absorption and scattering by rain are given for λ 3.2 centimeters. For most storm detection purposes, λ should be 3 to 6 centimeters rather than 9 to 12 centimeters, but 10-centimeter radar gives stronger return signals than 3-centimeter equipment through heavy rain over distances in excess of about 40 kilometers.

621.396.96:621.396.11 434
Reflexion of Centimetric Electromagnetic Waves over Ground, and Diffraction Effects with Wire-Netting Screens—Hey, Parsons, and Jackson. (*See* 519.)

621.396.96 435
Radar System Engineering [Book Review]—L. N. Ridenour, (Ed.). McGraw-Hill, New York, 1947, 748 pp., $7.50. (*Tele-Tech*, vol. 6, p. 93; September, 1947.) The first of a series of 28 books intended to serve as a general treatise and reference work. It deals primarily with micro-wave pulse radar.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 436
An Investigation on Hot-Wire Vacuum Gauges—H. von Ubisch. (*Ark. Mat. Astr. Fys.*, vol. 34, part 2, section A, 33 pp; September 25, 1947. In English.) The theory of the thermal conductivity of gases is outlined in relation to hot-wire gauges for pressures low enough to preclude convection. The influence of wire dimensions and bridge circuits on the attainment of maximum sensitivity is discussed and the results are described of experimental work on W, Mo, Ni, and Pt wires of various sizes in air, H_2 and CO_2. Typical calibration curves are shown covering the pressure range 10^{-4} to 20 mm. Hg.; pressure changes of 10^{-8} mm. Hg. can be detected using a manually operated bridge at pressures below 10^{-4} mm. Hg. Numerous graphical and tabulated data and a list of 46 references are given.

535.37 437
The Light Emission from Fluorescent Screens Irradiated by X-Rays—H. A. Klasens. (*Philips Res. Rep.*, vol. 2, pp. 68–78; February, 1947.) "Applying Schuster's theory as extended by Hamaker for the scattering and absorption of light, general equations are deduced for the amount of light emitted by fluorescent screens, irradiated by X-rays. Some commercial screens are examined to measure the 'absorption' coefficient σ of the fluorescent light. Several means to increase the brightness of a screen are discussed."

537.228.1 438
The Rochelle-Electric Lattice of KH_2PO_4-Type and the Behaviour of the NH_4-Rotation Transformation for $(NH_4, Tl)H_2PO_4$ Mixed Crystals—B. Matthias, W. Merz, and P. Scherrer. (*Helv. Phys. Acta*, vol. 20, pp. 273–306; August 4, 1947, In German.) An investigation of the dielectric properties of $(NH_4)H_2PO_4$ and of $(NH_4, Tl)H_2PO_4$ mixed crystals. Introduction of Tl+ions causes large changes of the NH_4 transformation temperature.

539.232 439
The Theory of the Formation of Protective Oxide Films on Metals: Part 3—N. F. Mott. (*Trans. Faraday Soc.*, vol. 43, pp. 429–434; July, 1947.) "A new mechanism is proposed to account for the formation on metals of oxide films which grow to a limiting thickness. According to this, electrons can pass the film easily, but ions can only penetrate it in the presence of a very strong field. This mechanism is compared with the author's previous theory based on tunnel effect; further experimental work is required to determine which is correct." For earlier parts see *ibid.*, 1939, vol. 35, pp. 1175–1177, and 1940, vol. 36, pp. 472–483; see also *Nature* (London), vol. 145, pp. 996–1000; June 29, 1940.

546.883 440
Some Applications of Tantalum in Electronics—L. F. Yntema and R. W. Yancey. (*Beama Jour.*, vol. 54, pp. 324–325; September, 1947.) High melting point, strength at high temperatures, ease of working and welding, chemical inertness, ease of cleaning, and pronounced getter action make Ta particularly useful in the construction of transmitting tubes.

620.193.33+620.197 441
On Rust and the Protection of Iron—M. Ragg. (*Elektrotech. und Maschinenb.*, vol. 64, pp. 150–160; September–October, 1947.) A comprehensive discussion of the subject, dealing particularly with the chemistry of passivation processes, the selection of suitable pigments for protective coatings and various special surface treatments.

620.197:679.5 442
Resin-Plotting for Sub-Assemblies—(*Tele-Tech*, vol. 6, p. 53; September, 1947.) Another account of the resin noted in 3928 of January.

621.315.59 443
Excess-Defect Semiconductor Contacts—L. Sosnowski. (*Phys. Rev.*, vol. 72, pp. 641–642; October 1, 1947.) These contacts have been considered theoretically with the object of explaining photo voltaic and rectifying properties of a thin layer of PbS which are not associated with a contact with metal electrodes. Fuller information is available in Admiralty Research Laboratory reports ARL/R8/E320 and ARL/R9/E320.

621.315.612:621.315.62 444
Stabilized Insulators—G. H. Gillam. (*Elec. Times*, vol. 112, pp. 289–293; September 11, 1947.) A general account of the properties of semiconducting glazes. Examples are given of their use in suppressing arc breakdowns in the high voltage insulators used in power transmission lines.

621.315.612.2 445
Alkaline Earth Porcelains Possessing Low Dielectric Loss—M. D. Rigterink and R. O. Grisdale. (*Bell. Sys. Tech. Jour.*, vol. 26, p. 688; July, 1947.) Summary of 3550 of 1947.

621.315.612.4.011.5 446
Effect of Field Strength on Dielectric Properties of Barium Strontium Titanate—H. L. Donley. (*RCA Rev.* vol. 8, pp. 539–553; September, 1947.) For high dielectric constant ceramics, as for ferromagnetic materials, the variation of dielectric flux density with field strength shows saturation effects. There is also a critical or Curie temperature above which dielectric constant varies linearly with field strength. The dependence of dielectric properties on field strength is shown graphically for three different mixtures of $BaTiO_3$ and $SrTiO_3$. The results suggest the use of these materials as nonlinear circuit elements; their use in a frequency multiplier, frequency changer, and frequency modulator is discussed. See also 3551 of 1947 (Bunting, Shelton, and Creamer).

621.315.613.1:549.623.52 447
Capacitance Stability of Ruby Muscovite Mica—W. Schick. (*Jour. I.E.E.* (London), part I, vol. 94, pp. 371–376; August 1947.) Short-term stability of dielectrics in general and measurements of capacitance variations of mica with temperature and air pressure are discussed. For flawless mica in the temperature range 25°C to 95°C, the capacitance change is substantially linear and cyclic, and a single temperature coefficient, whose measured values are evenly spread between $+6\times10^{-6}$ and $+40\times10^{-6}$, defines the dielectric behavior adequately. Plates with gas inclusions produce irregular and mostly very large changes of capacitance with temperature and particularly with air pressure.

621.315.616:549.623.5 448
[German] Manufacture of Synthetic Mica—A. E. Link. (*Tele-Tech*, vol. 6, pp. 67–68; July, 1947.) Abstract of article in *Chimie and Industrie* (Paris), vol. 56, p. 21: The synthetic product, the composition of which is given, is stated to be comparable with natural mica. Large plates are obtained by controlling the cooling of the melted constituents and by applying a magnetic field. See also Field Information Agency Technical Final Report No. 746, entitled "Synthetic Mica Research," and British Intelligence Objectives Subcommittee Final Report No. 785, entitled "The German Mica Industry."

621.318.22 449
Control of Permanent-Magnet Alloy Quality—J. D. Seaver and R. E. Anderson. (*Gen. Elec. Rev.*, vol. 50, pp. 44–47; October, 1947.) Possible methods of control are discussed, and apparatus is described for testing the remanence, in a predetermined magnetizing field.

621.383.4 450
The Photo-Conductivity of "Incomplete Phosphors"—R. Frerichs. (*Phys. Rev.*, vol. 72, pp. 594–601; October 1, 1947.) Synthetic single crystals of CdS, CdSe, and CdTe have been produced by the reaction of cadmium vapor with H_2S, H_2Se, and H_2Te, respectively. These "incomplete phosphors" show no phosphorescence but strongly developed photo-conductivity. Photo cells made from these crystals are extremely sensitive in the whole region from the infrared down to the ultraviolet, x-rays, and gamma-rays and for corpuscular rays, alpha- and beta-rays. Two different mechanisms of photoconductivity occur, namely, the normal photoconductivity in the region of strong absorption from the blue to the ultraviolet, and the selective photoconductivity in the region of weak absorption in the visible, x-rays, and corpuscular-rays region. The phenomena observed are in general accordance with the zonal theory of phosphorescence.

621.775.7 451
British Advances in Powder Metallurgy—(*Machinery* (London), vol. 71, p. 295; September 11, 1947.) A review of Special Report No. 38 of the Iron and Steel Institute, London, containing the 28 papers presented at the symposium noted in 150 of February.

679.5:621.3 452
The Growing Importance of Plastics in the Electrical Industry—G. Haefely. (*Jour. I.E.E.* (London), part II, vol. 94, pp. 301–308; August, 1947. Discussion, pp. 308–312.) Full paper, of which summaries were noted in 1122 and 2161 of 1947.

535.37 453
Fluorescence et Phosphorescence [Book Review]—M. Curie. Hermann, Paris, 1946, 212 pp., 350 fr. (*Nature* (London), vol. 160, pp. 483–484; October 11, 1947.)

621.315.612 454
German Radio Ceramics [Book Notice]—B.I.O.S. Final Report No. 1459. H. M. Stationery Office, London, 261 pp., 30s. Report of an investigation into the war-time activities of the Hermsdorf Schomburg Isolatoren Gesellschaft, Steatite-Magnesia, and other companies. "The data obtained covered the composition, manufacture, and properties of the various ceramic bodies, also details of the numerous types of finished radio components produced. Measuring apparatus, research, and development were investigated in detail."

621.315.612.4 455
Experimental Low Temperature Coefficient Ceramics. Variation of Capacitance and Power Factor with Temperature [Book Review]—A. M. Thomas. Brit. Elec. and Allied Indus. Res. Assn., Tech. Rep. L/T 170, 1946, 15 pp., 7s. (*Beama Jour.* vol. 54, p. 321; September, 1947.) Ten experimental medium-permittivity ceramics were examined at frequencies in the range 800 c.p.s. to 2.5 Mc. and at temperatures between −31°C and 200°C. Permanent changes in dielectric properties were observed after the materials had been heated to 200°C.

MATHEMATICS

512.831 456
Application of the Small Parameter Method to [oscillatory] Systems Similar to Those of Sturm-Liouville—S. M. Rytov and M. E. Zhabotinski. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 11, no. 2, pp. 135–140; 1947. In Russian, with English summary.)

512.831:518.5 457
An Experimental Determination of the Eigenvalues and Functions of Certain Operators by Means of an RC Circuit—L. A. Lyusternik and A. M. Prokhorov. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 11, no. 2, pp. 141–145; 1947. In Russian, with English summary.) A method applicable to a certain class of symmetrical-A matrices, similar to the Sturm-Liouville operators and the positive symmetrical operators of Fredholm.

517.63:621.3.015.3 458
Oscillations and Transient Phenomena. Their Study by Means of the Laplace and Cauchy Transformations—L. Bouthillon. (*Ann. Radioélec.*, vol. 2, pp. 287–328; October, 1947.) Part 1 studies (a) mechanical systems nearly in equilibrium, (b) electrical networks, (c) gyroscopic systems, (d) electromechanical systems with magnetic coupling. All these systems are represented by linear differential equations with constant coefficients. The classical form of solution is recalled and the importance and diversity of its applications are shown by examples of second-order equations. Continuous media are then considered, with special reference to an equation of partial derivatives of the second order, particular cases of which are the telegraphy equation, the diffusion equation, and the wave equation. Part 2 discusses the application of the Laplace and Cauchy transformations. Certain definitions and results of the theory of functions of complex variables are recalled, the Laplace and Cauchy transformations defined, the conditions of the Laplace-Cauchy inversion enumerated and the principal rules given for the transformation calculations. Application is made to linear differential equations with contant coefficients, the method of integration is shown and solutions for various practical examples depending on second-order equations are discussed. Rules are given for the application of the method to the integration of linear equations with partial derivatives. Two important examples considered are the diffusion equation and the wave equation. In conclusion, tables are given showing the principal rules for calculation and also a number of pairs of associated functions.

518.5 459
Electrical Analogue Computing: Part 4—Pure Electronic Systems—D. J. Mynall. (*Electronic Eng.*, vol. 19, pp. 283–285; September 1947.) Blumlein time integrator *RC* circuits are connected in cascade to solve linear differential equations with constant coefficients, and the solution may be shown on a c.r. tube. Schemes for multiplying and dividing voltages by means of tube amplifiers are also outlined. For earlier parts see 3563 of 1947 and 157 of February.

518.5:621.385 460
Tube Failures in ENIAC—Michael. (*See* 601.)

518.61:621.396.619.13 461
Computation of the Solutions of $(1+2\epsilon \cos 2z)y''+\theta y=0$; Frequency Modulation Functions—N. W. McLachlan. (*Jour. Appl. Phys.*, vol. 18, pp. 723–731; August, 1947.) Carson has given an approximate solution of this equation, stable for the values of θ and ϵ encountered in radio broadcasting. Acoustic tests, using warble tones to reduce standing-wave effects, require an extended range of these parameters. Floquet's theory is applied to obtain stable solutions for such cases. The necessary formulas for accurate calculation are derived and a numerical example is given. An approximate normalization of the solutions is suggested in order to obtain standard f.m. functions.

519.271:539.16.08 462
Note on the Statistical Analysis of Counter Data—N. Hole. (*Ark. Nat. Astr. Fys.*, vol. 34, part 2, section B, 8 pp., September 25, 1947. In English.)

517.564.3 (083.5) 463
Bessel Functions: Vols. 3 and 4, Annals of the Computation Laboratory of Harvard University: Tables [Book Review]—Harvard University Press, 1947, vol. 3, 694 pp., $10; vol. 4, 662 pp., $10. (*Tele-Tech*, vol. 6, pp. 93–94; September, 1947.) Tables of $J_0(x)$ and $J_1(x)$ in vol. 3, and $J_2(x)$ and $J_3(x)$ in vol. 4, computed to 18 places by means of the automatic sequence controlled calculator (461 and 787 of 1947).

518.5 464
Calculating Machines—Recent and Prospective Developments [Book Review]—D. R. Hartree. Cambridge University Press, England Macmillan Co. New York, 1947, 40 pp., 2s. and 75¢. (*Nature* (London), vol. 160, p. 142; August 2, 1947. *Elec. Eng.* vol. 66, p. 1047; October, 1947.) Analogue and digital machines are described, with special reference to the ENIAC. See also 2480, 2481 of 1947, 3952 of January, and 158 of February (Burks).

MEASUREMENTS AND TEST GEAR

531.765:621.38 465
A Millisecond Chronoscope—R. S. J. Spilsbury and A. Felton. (*Jour. I.E.E.* (London), part II, vol. 94, pp. 316–322; August, 1947. Discussion, pp. 322–324.) Full paper, of which a summary was noted in 1137 of 1947.

621.317+53.08 466
Significance of Functional Analysis of Measurements—H. C. Dickinson. (*Gen. Elec. Rev.*, vol. 50, pp. 13–16; October, 1947.) For functional analysis, any measurement system is divided into three principal functional groups: the primary detector, the end device, and the intermediate means. Each group is further subdivided into basic elements. Measurement energy is regarded as flowing from the quantity measured through the apparatus to the end device. Specific examples are discussed.

621.317+53.08 467
Fundamentals of Measurement Technique—Hartmann. (*See* 390.)

621.317.3:621.392.43 468
Precision Measurement of Impedance Mismatches in Waveguide—A. F. Pomeroy. (*Bell Sys. Tech. Jour.*, vol. 26, pp. 446–459; July, 1947.) "A method is described for determining accurately the magnitude of the reflection coefficient caused by an impedance mismatch in waveguide by measuring the ratio between incident and reflected voltages. Reflection coefficients of any value less than 0.05 (0.86 db standing-wave ratio) can be measured to an accuracy of ±2.5%."

621.317.333.82 469
The Henley 1,200,000 Volt Impulse Testing Plant: Part 2—The Impulse Generator: Part 3—The High Speed Cathode-Ray Oscillograph—T. R. P. Harrison. (*Distrib. Elec.*, vol. 20, pp. 252–255, and 278–282; July and October, 1947.) Details of the charging unit, capacitor bank, capacitor potential divider, tripping mechanism, voltage measurement apparatus and control gear. Part 1: 2848 of 1947.

621.317.336 470
The Measurement of H.F. Impedance and Applications of the Standing-Wave Indicator—H. J. Lindenhovius. (*Tijdschr. ned. Radiogenoot.*, vol. 12, pp. 65–82; March, 1947. In Dutch, with English summary.) A survey of different methods. For frequencies below 300 Mc. the impedance is usually determined by its effect on a tuned circuit; for higher frequencies it is best to use an untuned transmission line and determine the standing-wave ratio and the position of the voltage minimum; a graphical method of deducing the impedance is discussed. A standing-wave indicator and its applications are described.

621.317.41+621.318.323.2.042.15 471
Determination of the Permeability of Powdered-Iron Ring Cores—R. Schiffermüller. (*Elektrotech. und Maschinenb.*, vol. 63, pp. 254–256; November–December, 1946.) A method for rapid testing. A small coil within the ring to be measured is fed from an 800-c.p.s. source. The voltage induced in two coils outside the ring which are connected in series depends on the permeability of the ring. This voltage is amplified and applied to an indicating instrument which is calibrated to give direct readings of permeability.

621.317.44 472
A Sensitive Recording Magnetometer—A. Butterworth. (*Jour. I.E.E.* (London), part II, vol. 94, pp. 325–330; August, 1947. Discussion, pp. 330–332.) Full paper, of which a summary was noted in 1137 of 1947.

621.317.44:621.395.623.741 473
Testing Loudspeaker Magnets—E. E. George. (*Gen. Elec. Rev.*, vol. 50, pp. 24–26; October, 1947.) Details of the requirements, construction, and operation of a new, portable, vibrating tester for the rapid production testing of magnets. The specimen is temporarily magnetized and a flux-density measurement is made, under normal operating conditions, by means of a continuously vibrating search coil and associated tube voltmeter. The search coil oscillates with constant amplitude.

621.317.7 474
Progress in Instrument Design—D. B. Fisk and J. M. Whittenton. (*Gen. Elec. Rev.*, vol. 50, pp. 8–11; October, 1947.) An illustrated general survey. Discussion with specific examples of (a) improvements of basic design, (b) extension of usefulness by means of accessories, (c) modifications of basic designs to meet special needs.

621.317.7.029.64 475
On Certain Instruments for Measurements at Centimeter Wavelengths—M. T. Grekhova, S. I. Averkov, D. I. Grigorash, and V. I. Anikin. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 11, no. 2, pp. 183–189; 1947. In Russian, with English summary.) Descriptions of (a) receiver-wavemeter with automodulation, (b) wavemeter with c.r. indicator, (c) voltmeter, (d) field-strength meter.

621.317.715.085.39 476
Note on "A Simple Galvanometer with Negative Feedback"—D. K. C. MacDonald. (*Jour. Sci. Instr.*, vol. 24, pp. 232–233; September, 1947.) A description of a sensitive galvanometer system employing photo cell amplification and series feedback. The circuit is suitable for measurement of small potentials such as those occurring in metallic conductivity experiments. It is shown how performance equations may be derived for both the series and parallel-feedback systems. For Preston's earlier design see 3670 of 1947 and 1513 of 1947.

621.317.725 477
A General-Purpose Valve Voltmeter—F. Gutmann. (Proc. I.R.E. (Australia), vol. 8, pp. 16–20; August, 1947.) Requirements and desirable characteristics of such an instrument are discussed, together with the construction, operation, and performance of an inexpensive instrument using a cathode-follower type probe. A balanced metering system permits the separate determination of a.c. and d.c. simultaneously present. The voltmeter can be used over a wide range of voltages and frequencies. A bibliography of 25 items is given.

621.317.725.027.7 478
The Design of an Ellipsoid Voltmeter for the Precision Measurement of High Alternating Voltages—In abstract 3584 of 1947, it was stated that "the voltage between the disks is deduced with an estimated error of less than 0.3 per cent." This should read 0.03 per cent.

621.317.755:[621.317.722+531.761 479
The Waveform Monitor: A Cathode-Ray Tube Equipment for the Measurement of Voltage and Time—H. L. Mansford. (*Electronic Eng.*, vol. 19, pp. 272–275 and 328–322; September and October, 1947.) Direct and alternating potentials are measured by a null method whereby a cathode-coupled amplifier gives push-pull deflections which indicate the balance of the input potential against another potential derived from and measured by the instrument. A similar amplifier is used to magnify and calibrate any portion of a synchronized timebase. The operation of the Miller integrator and of "long-tailed pair" circuits is also described.

621.317.755:[621.317.722+531.761 480
A Note on the Zero Setting Circuit of the Waveform Monitor—A. M. Spooner. (*Electronic Eng.*, vol. 19, p. 332; October, 1947.) Comment on 479 above.

621.317.761 481
An Instrument for Short-Period Frequency Comparisons of Great Accuracy—H. B. Law. (*Jour. I.E.E.* (London), part I, vol. 94, p. 377; August, 1947.) Summary of 2178 of 1947.

621.317.761:621.318.572 482
Frequency Standards and Electronic Counters—B. van Dijl. (*Tijdschr. ned. Radiogenoot.*, vol. 12, pp. 37–61; March, 1947. Discussion, pp. 62–64. In Dutch, with English summary.) Discussion of (a) the generation of standard frequencies, (b) the measurement of their constancy, (c) the use of counters for such measurements, and (d) results obtained.

621.317.761.029.64 483
A Microwave Frequency Standard—R. G. Talpey and H. Goldberg. (Proc. I.R.E., vol. 35, pp. 965–969; September, 1947.) A stabilized 10-Mc. quartz-crystal oscillator feeds a multiplier chain, and the voltage outputs at 20, 40, 120, and 360 Mc. are mixed in a silicon-crystal harmonic generator with a crystal current of 30 to 40 ma. The output spectrum extends to at least 10,000 Mc.; frequencies are identified by means of a coaxial-line wavemeter. Images are easily identified using a superheterodyne detector with a midband frequency of 0.5 Mc. and a bandwidth of 0.6 Mc.

621.317.763+621.396.616+621.392.029.64 484
The Transverse Electric Modes in Coaxial Cavities—R. A. Kirkman and M. Kline. (Proc. I.R.E., vol. 35, pp. 931–935; September, 1947.) Discussion on 888 of 1946.

621.317.763.089.6 485
A Method for Calibrating Microwave Wavemeter—L. E. Hunt. (Proc. I.R.E., vol. 35, pp. 979–981; September, 1947.) A quartz-controled oscillator and harmonic generator system is used as a frequency reference. The frequency of a calibrating oscillator, sufficiently powerful to operate the wavemeter, is continuously compared with the standard by separately heterodyning both outputs with that of a third oscillator to which a sawtooth f.m. voltage is applied. The resulting products are displayed on an oscilloscope so that frequencies can be continuously compared with an inaccuracy of less than 12 kc.

621.317.79 486
The Characteristic Recorder—T. Kammerloher. (*Funk-Technik* (Berlin), vol. 2, nos. 17 and 18, pp. 6–8 and 6–7; 1947.) The recorder has two moving-coil instruments with their pivotal axes at right angles. Each coil system carries a small mirror and a beam of light is reflected in turn from each mirror, falling finally on a screen. A detailed description is given of the use of the instrument for obtaining (a) resonance curves of tuned circuits and, (b) tube characteristics.

621.317.79:621.396.615 487
Special Applications of Ultra-High-Frequency Wide-Band Sweep Generators—J. A. Bauer. (*RCA Rev.*, vol. 8, pp. 564–575; September, 1947.) Discussion of the use of wide-band f.m. signal generators for (a) r.f. impedance measurements of wide-band terminal and other devices on a transmission line; (b) overall frequency response measurements of television receivers; (c) frequency measurements in the range 250 to 10,000 kc. within about 0.001 per cent. Large savings of laboratory and factory test time have already been achieved.

621.317.79:621.396.712 488
Monitoring Equipment for Frequency-Modulation Broadcasting—M. Silver. (*Elec. Commun.* (London), vol. 24, pp. 273; June, 1947.) Summary of I.R.E. Convention paper. The design and method of operation of such equipment are discussed, together with the measurements required to prove its performance. It is claimed that noise can be measured to -80 db, distortion to 0.2 per cent, and station carrier frequency to within ± 100 c.p.s. under full-modulation conditions with long-time stability of 1 part in 5×10^6.

621.317.79:621.396.822:621.385.1 489
Measurement of Valve Background Noise—M. Chamagne and G. Guyot. (*Télév. Franç.*, Supplement *Électronique*, pp. 36–39; September, 1947.) A short discussion of the origin of tube noise and a description of practical apparatus for its measurement. The method is that of Ziegler and uses a saturated diode as the comparison noise source. Measurements on EF5 and AC2 tubes are shown graphically. The results for AC2 tubes are in excellent agreement with those of M. J. O. Strutt.

621.317.79:621.396.822:621.385.2 490
How Sensitive Is Your Receiver?—B. Goodman. (*QST*, vol. 31, pp. 13–21; September, 1947.) Sources of noise are discussed, noise factor is defined as the ratio of the equivalent noise power of a receiver to that of an ideal receiver, and a detailed description is given of a simple diode noise generator which has proved very useful in measurements of noise factor. The method of use is described and illustrated by results obtained during the development of a cathode-coupled preamplifier.

621.317.79.029.64:621.396.81 491
On the Thermometric Method of Measuring

the Field Strength of Centimeter Waves—S. M. Rytov. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 11, no. 2, pp. 191–194; 1947. In Russian, with English summary.) Means are described for increasing considerably the sensitivity of the thermometer.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

519.271:539.16.08 492
Note on the Statistical Analysis of Counter Data—Hole. (*See* 462.)

535.61-15:621.317.755:535.33 493
An Infra-Red Spectroscope with Cathode-Ray Presentation—E. F. Daly and G. B. B. M. Sutherland. (*Proc. Phys. Soc.*, vol. 59, p. 901; September 1, 1947.) Discussion on 2860 of 1947.

539.16.08 494
Some Properties of Counters with Beaded Wires—S. C. Curran and E. R. Rae. (*Jour. Sci. Instr.*, vol. 24, pp. 233–238; September, 1947.) A description of the effect on the performance of a Geiger counter of attaching closely spaced beads on the central wire. The main result is a reduction in overall efficiency due to the production of local insensitive regions around the beads. The localization of the discharge and its variation with voltage, nature of the gas mixture, and pressure are discussed.

539.16.08 495
The Discharge Mechanism of Self-Quenching Geiger-Müeller Counters—S. H. Liebson. (*Phys. Rev.*, vol. 72, pp. 602–608; October 1, 1947.) Full paper: summary in 3982 of January.

539.16.08 496
Accurate Method of Measuring the Efficiency of Geiger-Müller Counters—A. Rogozinski and A. Voisin. (*Compt. Rend. Acad. Sci.* (Paris), vol. 225, pp. 409–411; September 1, 1947.)

539.16.08:531.717.1 497
Beta-Ray Thickness Gage for Sheet Steel—O. J. M. Smith. (*Electronics*, vol. 20, pp. 106–112; October, 1947.) "G-M counters and integrating circuits, responding to absorption of beta rays by steel strip moving over a radiostrontium source, measure thickness over range of 7 to 24 mils. Sheets can be sorted automatically by a mechanical gate after cutting. Accuracy is analyzed."

621.316.7.078 498
Theory of Automatic Control Systems—M. A. Melvin. (*Jour. Appl. Phys.*, vol. 18, pp. 704–722; August, 1947.)

621.316.71:629.12.014.6 499
The Use of High Permeability Materials in Magnetometers. The Application of a Saturated Core Type Magnetometer to an Automatic Steering Control—L. D. Armstrong. (*Canad. Jour. Res.*, vol. 25, sec. A, pp. 124–133; May, 1947.) The accuracy of steering was $\pm\frac{1}{3}°$, or better, with hunting so small as to have no noticeable effect on the ship's course.

621.316.86.001.8 500
The Thermistor in Biological Research—B. L. Andrew. (*Electronic Eng.* (London), vol. 19, pp. 288–289; September, 1947.) "Used as a resistance thermometer it has applications where there is a requirement for remote indication of temperature small size of sensitive element and a rapid response to changes of temperature." Suitable circuits are described and their operation discussed. See also 3044 of 1947 (Rosenberg).

621.317.083.7 501
Problems of Telemetry and Their Solutions—W. Boesch. (*Rev. Gén. Élec.*, vol. 56, pp. 355–368; September, 1947.) Part 1 discusses the requirements of and errors in telemetry systems. Part 2 describes briefly a wide range of devices, from simple voltmeter circuits to selsyns, frequency variation or compensation systems, and many pulse systems. A synoptic table of all the methods noted is given and its use is illustrated by practical examples.

621.317.083.7:551.46.018.1 502
Telemetering Fathometer—E. F. Kiernan. (*Electronics*, vol. 20, pp. 96–98; October, 1947.) For continuous depth indications beneath a remote-controlled pilotless ship. Frequencies near 8 and 70 Mc. are used.

621.317.39:620.178.3 503
A Twelve-Channel Recorder for Use with Resistance Strain Gauges—A. Watson. (*Jour. Sci. Instr.*, vol. 24, pp. 239–242; September, 1947.) For investigating stresses on a ship's hull. See also 3217 of 1947 (Cogman).

621.365.92:641.3 504
U.H.F. Heating of Frozen Foods—P. W. Morse and H. E. Revercomb. (*Electronics*, vol. 20, pp. 85–89; October, 1947.) Discussion of problems encountered in developing 1050-Mc. c.w. magnetron oscillator equipment for thawing and heating pre-cooked frozen food.

621.365.92.001.8 505
An Electronic Development of Growing Importance: Dielectric Heating and Its Many Practical Applications—G. R. Cooper. (*Overseas Eng.*, vol. 20, pp. 370–372; July, 1947.) A brief survey of the practical applications, with special reference to curing thermosetting glue bonds in the wood-working industry.

621.384 506
Travelling-Wave Linear Accelerator for Electrons—D. W. Fry, R. B. R.-S.-Harvie, L. B. Mullett, and W. Walkinshaw. (*Nature* (London), vol. 160, pp. 351–353; September 13, 1947.) A new application of the principle used in the Sloan and Lawrence positive ion accelerator (*Phys. Rev.*, vol. 38, pp. 2021–2032; December 1, 1931 is possible now that radar magnetrons can give high peak power in pulse operation at short wavelengths.

The 0.5-Mev. accelerator described uses a wavelength of 10 centimeters. A travelling wave is set up in a 40-centimeter length of circular waveguide with deep corrugations whose depth determines the phase velocity. Experimental and theoretical results agree closely. The accelerator would be working at optimum efficiency if the waveguide were extended to 20 meters.

With an applied r.f. power of 1 milliwatt, energy spectrum measurements show a maximum peak pulse beam current of 36 ma. at an energy very close to 540 kv. (the designed value) with a width of 65 kev. between half-amplitude points. A very large proportion of the injected electrons were trapped by the wave and formed into stable bunches.

621.384.6 507
Theory of the Proton Synchrotron—J. S. Gooden, H. H. Jensen and J. L. Symonds. (*Proc. Phys. Soc.*, vol. 59, p. 901; September 1, 1947.) Corrections to 209 of February.

621.384.6:621.313.322 508
Power Supply for the 100,000,000-Volt Betatron—F. L. Kaestle. (*Elec. World*, vol. 128, pp. 42–44; August 30, 1947.) Details of the synchronous-synchronous motor-generator set, 3-phase to single-phase, with amplidyne voltage-regulating equipment and reduced-voltage starting control, delivering the required 24,000-kva. excitation from a bank of 24-kv., 1000-ampere capacitors.

621.384.6(43) 509
European Induction Accelerators—R. Wideröe. (*Jour. Appl. Phys.*, vol. 18, p. 783; August, 1947.) Comment on 2527 of 1947 (Kaiser).

621.385.833 510
Coaxial Electron Lenses—J. W. Dungey and C. R. Hull. (*Proc. Phys. Soc.* (London), vol. 59, pp. 828–843; September 1, 1947.) A detailed mathematical account of these lenses, which contain a central conductor surrounded by a number of annular electrodes. The electrostatic fields in these lenses can be calculated by superimposing fields of a certain simple type. This type of field, which is tabulated, corresponds to a "one-element" coaxial lens with an annular electrode in the form of a perforated disk with rounded edges, preceded and followed by cylindrical guard rings. At least two such lenses are required to correct the spherical aberration inherent in the ordinary electron microscope, and a three-element correcting lens is better. These systems are analyzed in detail. The workmanship involved in their construction must be of the highest order. Careful focusing is required because of the small focal depth, which, however, suggests the possibility of examining objects in depth with an accuracy of the order of 100Å.

621.385.833 511
The Use of the Fluxball Method in the Measurement of the Axial Distribution of the Magnetic Field in Electron Microscope Lenses—K. I. Williamson. (*Jour. Sci. Instr.*, vol. 24, pp. 242–243; September, 1947.) A ballistic method in which a specially wound coil is withdrawn suddenly, by an elastic suspension, from a known position in the lens to a field-free region. The method is suitable for measurement of magnetic fields with rapid space variations.

PROPAGATION OF WAVES

621.396.11+621.396.65.029.64 512
Microwave Communication Link—Lamont, Robertshaw and Hammerton. (*See* 549.)

621.396.11 513
The Propagation of Radio Waves and the Inhomogeneity of the Atmosphere—H. Bremmer. (*Tijdschr. ned. Radiogenoot.*, vol. 12, pp. 7–29; January, 1947. In Dutch with English summary.) The theory of propagation of radio waves which suffer reflection at the ionosphere is considered from the point of view of geometrical optics, allowing for vertical atmospheric inhomogeneities. The condition for superrefraction is derived, and the behavior of short waves is compared with that of long waves.

621.396.11:551.510.535 514
Calculation of the Field of a Space Wave—K. Rawer. (*Rev. Sci.* (Paris), vol. 85, pp. 361–362; April 1, 1947.) The basic principle of the method described for calculating the field at a great distance from a short-wave transmitter consists in determining separately the fields corresponding to the different possible paths and then combining them. The results indicate that the inverse square law is only followed for a short distance from the transmitter and that beyond this distance the value of the field is greater than would be given by the inverse square law. The effects of the various ionosphere layers are discussed briefly.

621.396.11:551.510.535 515
The Prediction of Optimum Working Frequencies for Short Wave Radio Circuits—H. Stanesby and G. H. M. Gleadle. (*P. O. Elec. Eng. Jour.*, vol. 40, part 2, pp. 76–79; July, 1947.) A short general discussion of optimum frequencies and an account of methods for their determination.

621.396.11:551.510.535 516
The Forecasting of Ionosphere Critical Frequencies—Rawer. (*See* 419.)

621.396.11:551.510.535 517
On the Reasons for a Change of Amplitude of a Single Pulse Reflected from the Ionosphere—Gusev. (*See* 420.)

621.396.11:551.594.5:551.510.535 518
Ionospheric Perturbations in the Zone of Polar Auroras—Rawer. (*See* 421.)

621.396.11:621.396.96 519
Reflexion of Centimetric Electromagnetic Waves over Ground, and Diffraction Effects with Wire-Netting Screens—J. S. Hey, S. J. Parsons, and F. Jackson. (*Proc. Phys. Soc.*, vol. 59, pp. 847–857; September 1, 1947.) An account of a simple technique for determining the echo signal strength pattern in the vertical plane for a British Army G.L. III radar equipment, operating at λ 10.7 centimetres on natural sites both with and without artificial screening. The transmitter radiated 1-microsecond pulses at a peak power of 200 kw. Separate transmitter and receiver paraboloids with apertures of 1.22-meter diameter, mounted adjacently with foci 3.6 meters above the ground, could be traversed together in bearing or elevation. Vertical polarization was used and the reflector was a papier mâché sphere 0.6 meter in diameter, metallized with sprayed zinc and supported by a balloon. The range of the reflector was 2500 to 3000 meters and height varied from ground level to 500 meters.

The measurements obtained over different natural sites were in good agreement with theory. The effect of wire-netting screens of 22 standard wire gauge galvanized iron wire and 1.4-centimeter mesh, erected about 50 meters from the equipment, was measured and shown to be in agreement with simple diffraction theory.

621.396.11.029.6 520
Ten-Meter Propagation by Rebound Scattering—D. W. Heightman. (*CQ*, vol. 3, pp. 19–21, 87; September, 1947.) Fairly weak but reasonably consistent signals can be received on λ 10 meters from stations 50 to 500 miles away, well within the skip zone and too far away for the ground wave to be audible. Both receiving and transmitting aerials must point in approximately the same direction. It is suggested that these signals are not due to "long" and "short" scatter, but to direct scattering from the F_2 layer.

621.396.11.029.64 521
3- and 9-Centimeter Propagation in Low Ocean Ducts—M. Katzin, R. W. Bauchman, and W. Binnian. (Proc. I.R.E., vol. 35, pp. 891–905; September, 1947.) Description of one-way radio propagation measurements between ship and shore, coupled with meteorological measurements, made in the British West Indies area. Persistent low-level ducts (20 to 50 feet in height), whose strength and height appeared to depend on wind speed, were found. Various aerial-height combinations were used and very effective trapping was found on λ 3 centimeters for height between 6 and 15 feet. On λ 9 centimeters, the trapping was only partial. Attenuation rates on λ 9 centimeters averaged 0.85 db/nautical mile up to about 80 miles and 0.2 db/mile beyond, while the rate on λ 3 centimeters was 0.45 db/mile for all ranges up to 150 miles. Radar measurements on λ 3 centimeters gave similar results. Sed also 2900 of 1947 (Pekeris).

621.396.81+621.396.72 522
2½-Watt F.M. Transmitter Permits City-Wide Coverage—(*See* 554.)

RECEPTION

621.396.621 523
The Reception of Short Waves with a Standard Commercial Receiver—G. Trestchenkoff. (*Radio Franç.*, pp. 5–11; September, 1947.) The lack of sensitivity of some receivers on short waves is discussed and found to be frequently due to one or more of the following causes:—(a) poor gain for medium frequencies; (b) poor gain of high-frequency tuned circuits; (c) inefficient frequency changer; (d) grid current. Means are suggested for eliminating these sources of trouble.

621.396.621+621.396.69:06.064 London 524
Radio Exhibition [Olympia, London]—W. E. Miller. (*Elec. Rev.* (London), vol. 141, pp. 543–546; October 10, 1947.) A short account of some of the principal features. See also 617 below.

621.396.621:621.396.619.11 525
The Design of a Synchrodyne Receiver: Part 1—Design Principles—D. G. Tucker. (*Electronic Eng.* (London), vol. 19, pp. 241–245; August, 1947.) Design principles, together with all the information required for the construction of a receiver of this type. A locked oscillator is used for demodulation. See also 2364 of 1947 and 526 below.

621.396.621:621.396.619.11 526
The Design of a Synchrodyne Receiver: Part 2—Some Suitable Designs—D. G. Tucker and J. F. Ridgway. (*Electronic Eng.* (London), vol. 19, pp. 276–277; September, 1947.) Three circuits are given: (a) a cathode-coupled Cowan ring demodulator for low sensitivities (discussed more fully in 525 above); (b) two r.f. stages and a ring demodulator for high sensitivities; (c) a very simple receiver using a triode-hexode for demodulation.

621.396.621:621.396.645 527
Intermediate-Frequency Amplifiers for Frequency-Modulation Receivers—J. J. Adams. (Proc. I.R.E., vol. 35, pp. 960–964; September, 1947.) Voltage feedbacks must be reduced to a minimum to obtain good results in mass production without stagger tuning. Selectivity and stability formulas, stabilizing methods, and methods of aligning double-tuned transformers are discussed.

621.396.621.54:518.4 528
Superheterodyne Tracking Charts—Y. P. Yu. (*Tele-Tech*, vol. 6, pp. 46–47, 108; September, 1947.) Charts for determining the values of inductances and capacitances in permeability-tuned and capacitance-tuned systems.

621.396.822:621.317.79:621.385.2 529
How Sensitive Is Your Receiver?—Goodman. (*See* 490.)

621.396.828:621.319.74:629.135 530
Electrostatic Dischargers for Aircraft—W. C. Hall. (*Jour. Appl. Phys.*, vol. 18, pp. 759–765; August, 1947.) Methods using high-mobility gaseous ions are preferred to those using low-mobility charge carriers such as water spray. The discharger invented and developed at the Naval Research Laboratory is described. See also 2916 of 1947 (Beach).

621.396.828:621.396.619.16 531
Noise-Suppression Characteristics of Pulse Modulation—S. Moskowitz and D. D. Grieg. (*Elec. Commun.* (London), vol. 24, pp. 271–272; June, 1947.) Summary of I.R.E. Convention paper. Discussion of the manner in which noise may enter a pulse-time modulation system and of the effectiveness of various methods used in its elimination. Another summary noted in 3269 of 1947.

621.396.828:621.397.62 532
Interference with Television Broadcasting—Grammer. (*See* 579.)

621.396.828:621.397.62 533
Electrical Interference Suppression in Television Receivers—Flach. (*See* 580.)

621.396.822 534
Radio Noise. Radio Research Special Report No. 15 [Book Notice]—H. A. Thomas and R. E. Burgess. H. M. Stationery Office, London, 1947, 3s. 4d. (*Elec. Rev.* (London), vol. 141, p. 546; October 10, 1947. *Govt. Publ.* (London), p. 9; September, 1947.) A survey of existing information about, and data on, radio noise within the 1- to 30-Mc. frequency band. Issued by the Department of Scientific and Industrial Research as a foundation for investigations now being started in various parts of the world.

STATIONS AND COMMUNICATION SYSTEMS

621.396/.397](47) 535
Broadcasting and Television Methods in the Soviet Republics—A. Huth. (*Tele-Tech*, vol. 6, pp. 30–33, 114; September, 1947.) The development, scope, and methods are discussed and the 1946 to 1950 Five Year Plan is outlined. Rediffusion methods, serving individual listeners and public places, are used extensively. High power medium and long-wave transmitters predominate; lists of these are included. A television center was opened in Moscow in 1938.

621.396.1 536
Narrow-Band F.M. Authorized—(*Electronics*, vol. 20, pp. 146, 240; October, 1947.) For class A amateurs, the frequency bands 3.85 to 3.90 Mc. and 14.20 to 14.25 Mc. have been authorized for f.m. R/T on an experimental basis. Frequencies in the ranges 28.5 to 29 Mc. and 51.5 to 52 Mc. may also be used at any licensed amateur radio station.

621.396.1 537
Space Diversity Reception at Super-High Frequencies—G. H. Huber. (*Bell Lab. Rec.*, vol. 25, pp. 337–341; September, 1947.) Describes the use, in California, of vertically spaced aerials for diversity reception of pulse position modulated 4350 to 4800-Mc. signals. 510 miles were covered in optical stages of from 24 to 170 miles, some over land and some over water. Results indicate that the complementary vertical space diversity method is of great value in improving performance of super-high-frequency relay paths operating over sea or smooth land.

621.396.1:621.397.828 538
Engineering Problems Involved in TV Interference—A. Francis. (*Tele-Tech*, vol. 6, pp. 42–45, 109; September, 1947.) Mutual interference makes it impracticable for other services in the United States to share television channels. It is recommended that the television frequency spectrum should be extended and the frequencies allotted to different services reallocated.

621.396.332 539
Tape Relay System for Radiotelegraph Operation—S. Sparks and R. G. Kreer. (*RCA Rev.*, vol. 8, pp. 393–426; September, 1947.) Telegrams are received in a form suitable for immediate retransmission, so that service is quickened and operating costs and the possibilities of human error are reduced.

621.396.41:621.396.97 540
Ultra-High-Frequency Multiplex Broadcasting System—A. G. Kandoian and A. M. Levine. (*Elec. Commun.* (London), vol. 24, p. 268; June, 1947.) Summary of I.R.E. Convention paper. Pulse time modulation is used. A highly directive receiving aerial is permanently focused on the transmitter. The required program is chosen by push-button selection of simple timing circuits.

621.396.41:621.396.97:621.396.619.16 541
6-Channel Multiplex Equipment for Broadcasting—Chamagne and Guyot. (*Onde Élec.*,

vol. 27, pp. 318–329; August–September, 1947.) A short discussion of the principles of pulse modulation, with some details of apparatus using pulse duration modulation. This apparatus combines high-fidelity reproduction with low distortion and background noise.

621.396.41.029.64:621.395.43 542
Multiplex Microwave Radio Applied to Telephone Systems—T. H. Clark. (*Elec. Commun.* (London), vol. 24, pp. 265–266; June, 1947.) Summary of I.R.E. Convention paper. Two systems are described. One has a single wide-band channel capable of transmitting the frequency spectrum presented by a conventional frequency-division-multiplex carrier system. In the other, a number of telephone conversations are applied as voice bands to time-division-multiplexing equipment.

621.396.619.11/.13 543
Amplitude and Frequency Modulation—A. A. McKenzie. (*Wireless Eng.*, vol. 24, p. 332; November, 1947.) Criticism of 3660 of 1947 (Nicholson). For a direct comparison between an a.m. and a f.m. system, reference is made to 3528 of 1939 (Weir). A short paper by C. M. Jansky, Jr., entitled "The Demonstrated Potentialities of Frequency Modulation Broadcasting on Very High Frequencies," prepared for the International Telecommunications Conference at Atlantic City on August 6, 1947, also summarizes the advantages inherent in f.m. broadcasting.

621.396.619.16 544
Pulse Count Modulation System—D. D. Grieg. (*Tele-Tech*, vol. 6, pp. 48–52, 98; September, 1947.) Transmission is effected by dividing the amplitude of the modulating signal into finite levels, determining the instantaneous amplitude at short time intervals and then representing each amplitude level by different combinations of constant amplitude pulses. The demodulated signal is discontinuous, but intelligible speech can be obtained with seven modulation levels.

Time-division multiplex may be used. Relaying does not increase noise since the pulses are regenerated. Oscillograms of signals, and block schematic diagrams are given, together with tables of the distortion obtained and the number of pulses required per signal for different numbers of levels and typical pulse repetition rates.

621.396.619.16:621.395.43 545
Telephony by Pulse Code Modulation—W. M. Goodall. (*Bell. Sys. Tech. Jour.*, vol. 26, pp. 395–409; July, 1947.) Another pulse code modulation system was noted in 258 of February (Batcher).

621.396.65+621.396.7] (494) 546
High-Altitude [radio] Stations and Links—W. Gerber and F. Tank. (*Tech. Mitt. Schweiz. Telegr.-Teleph. Verw.*, vol. 25, pp. 177–186; October 1, 1947. In German.) The mountainous character of Switzerland offers distinct possibilities for the use of cm-, dm-, and m-wave technique for communication purposes, including telephony to alpine huts and other relatively inaccessible stations, broadcasting, facsimile transmission, and television. A station at Chasseral, 1608 meters above sea level, provides coverage over a wide area by line-of-sight links using a multichannel R/T system.

621.396.65:523.3 547
Considerations of Moon Relay Communications—D. D. Grieg, S. Metzger, and R. Waer. (*Elec. Commun.* (London), vol. 24, pp. 266–267; June, 1947.) Summary of I.R.E. Convention paper. Discussion of the possibilities of the moon as a passive repeater for radio links. The Doppler frequency shift due to the relative motion of earth and moon, the effect of cosmic noise, and the moon's radio reflecting properties are considered. See also 106 of 1947 (Clarke).

621.396.65:621.396.41:621.396.619.16 548
Pulse-Time-Modulated Multiplex Radio Relay System—Radio-Frequency Equipment—D. D. Grieg and H. Gallay. (*Elec. Commun.* (London), vol. 24, pp. 141–158; June, 1947.) Useful microwave transmission is limited to the band 1000 to 7000 Mc. At these frequencies, high frequency-stability and i.f. amplification are necessary, and transmission lines are inefficient. Many problems are solved by the use of pulse-time modulation.

The four-stage receiver in the experimental New York-Trenton link has a gain of 80 db, bandwidth of 8 Mc. and an image rejection ratio of 72 db. The transmitter operates at frequencies between 1200 and 1300 Mc., incorporates a 2C43 tube and can modulate at frequencies up to 3.5 Mc.

Tower heights should be 100 to 200 feet and repeater stations 20 to 30 miles apart. Power supplies may be derived from the wind. Details of the performance of the link are given. For a description of the terminal equipment see 1213 of 1947 (Grieg and Levine).

621.396.65.029.64+621.396.11 549
Microwave Communication Link—H. R. L. Lamont, R. G. Robertshaw, and T. G. Hamerton. (*Wireless Eng.*, vol. 24, pp. 323–332; November, 1947.) A single-channel duplex 3.2-centimeter R/T system is described. A single parabolic mirror is used as aerial for both transmitted and received signals, which are separated in a waveguide system. The transmitter is a klystron oscillator of about 75 milliwatts output. The superheterodyne receiver is fitted with automatic frequency control and automatic gain control. The equipment has been in operation for over a year; it was installed to provide R/T communication over one of the 57-mile oversea optical paths used in the propagation experiments described in 518 of 1947 (Megaw). Variations in signal strength occur mainly in fine weather, within a range of ± 10 db, but the automatic gain control makes them unnoticeable. Occasional very deep rapid fading was also observed at intervals over a period of perhaps an hour. Operation over other optical paths up to 70 miles long is also described.

621.396.65.029.64 550
Microwave Radio Relay Systems—E. Labin. (*Elec. Commun.* (London), vol. 24, pp. 131–140; June, 1947.) The effects of topography, of atmospheric refraction and of absorption on line-of-sight microwave propagation are discussed; a 30-db overall margin in power should be sufficient to make failure improbable. Both frequency selection and time division methods are available for multiplex transmission, but each requires a large bandwidth. The determination of the transmitter power required for such systems is discussed; powers of the order of several watts are necessary to allow for fading. Details of a Paris-Montmorency experimental link and of a New York-Trenton experimental link are given; see also 548 above and 551 below, 3283 of 1947 (Gerlach) and back references.

621.396.65.029.64 551
Paris-Montmorency 3000-Megacycle Frequency-Modulation Radio Link—A. G. Clavier and G. Phélizon. (*Elec. Commun.* (London), vol. 24, pp. 159–169; June, 1947.) English version of 540 of 1947.

621.396.7 552
Tangier Radio Relay Station—(*Elec. Commun.* (London), vol. 24, p. 208; June, 1947.) The station provides a through teletype service between New York and Moscow, and 24-hour service between New York and certain European and Asiatic cities. The reliability of the relayed circuits is high because they avoid direct routes which are very long, or which pass near to the north magnetic pole and are subject to severe ionospheric disturbances.

621.396.712 553
The Allouis (France) Short-Wave Broadcasting Center—M. Matricon. (*Jour. Brit. I.R.E.*, vol. 7, pp. 184–193; September, 1947.) A description of the transmitting equipment. There are two separate transmitters with independent power supply. Each transmitter consists of three h.f. chains, any two of which can be connected at will to two independent l.f. chains so that four programs may be transmitted simultaneously.

Each l.f. chain consists of line amplifier, equalizers and limiters, input amplifier, submodulator and modulator. The h.f. chains comprise crystal oscillators (stability 1 in 10^6), frequency doublers, 1-kw. amplifier, 10-kw. amplifier, and power stage using two continuously evacuated demountable tubes. The h.f. power output is not less than 100 kw. Twelve rhombic aerials are provided; connections to the transmitters are made from a control desk by servomechanism.

621.396.72+621.396.81 554
2½-Watt F.M. Transmitter Permits City-Wide Coverage—(*Tele-Tech*, vol. 6, pp. 34–35; September, 1947.) Block diagram and performance of an 88-Mc. General Electric phasitron-based transmitter installed at Syracuse University, United States. A coverage radius of 7 miles for a field strength of 50 microvolts per meter at a height of 30 feet was obtained with a horizontal circular loop transmitting aerial at a height of 100 feet. Calculated and measured field strengths are compared; discrepancies are attributed to the hilly terrain.

621.396.931 555
The Problems of Radio Communication with Moving Trains—G. H. Leversedge. (*Jour. Brit. I.R.E.*, vol. 7, pp. 157–163; July–August, 1947. Discussion, pp. 163–164). Brief survey of operational requirements for speech transmission and difficulties met in practice, with particular reference to British railways. Results of recent tests on the London and North Eastern Railway are summarized.

621.396.931.029.62 556
Mobile Frequency-Modulation 30–44-Megacycle Equipment—R. B. Hoffman and E. W. Markow. (*Elec. Commun.* (London), vol. 24, pp. 170–178; June, 1947.) The system "is characterized by simplicity of operation, small size of mobile equipment improved receiver squelch circuit, low power drain for mobile equipment during stand-by, plug-in arrangement of units for ease of replacement and servicing, and a self-contained selective calling system." See also 3665 of 1947.

621.396.97 557
U.N. Telecommunications Facilities—J. Peterson. (*Tele-Tech*, vol. 6, pp. 24–28, 102; September, 1947.) A $6,000,000 plan to provide world-wide coverage for United Nations programs by linking them to national and local broadcast systems.

621.396.619.13 558
Frequency Modulation Engineering [Book Review]—C. E. Tibbs. Chapman and Hall, London, 310 pp. 28s. (*Elec. Rev.* (London), vol. 141, p. 771; November 21, 1947.) "The book . . . can be recommended to all requiring information about frequency modulation."

SUBSIDIARY APPARATUS

621.3.016.3.029.64 559
Power Loads at Very- and Ultra-High Frequencies—A. G. Kandoian and R. A. Felsenheld. (*Elec. Commun.* (London), vol. 24, pp. 267–268; June, 1947.) Summary of I.R.E. Con-

vention paper. The following new designs of power load are described (a) transmission-line: a coaxial line with circulating water used as a dielectric and cooling agent; (b) radiator-type: an aerial or radiator enclosed in a tank of water; (c) resonant-cavity type: a $\lambda/4$ coaxial resonant circuit of low Q.

621.314.65:621.396.71 560
The Application of High-Voltage Steel-Tank Mercury-Arc Rectifiers to Broadcast Transmitters—P. A. T. Bevan. (*Jour. I.E.E.* (London), part II, vol. 94, pp. 299–300; August, 1947.) Discussion on 206 of 1946.

621.314.65:621.396.71 561
High-Voltage Steel-Tank Mercury-Arc Rectifier Equipments for Radio Transmitters—J. C. Read. (*Jour. I.E.E.*, (London), part II, vol. 94, pp. 299–300; August, 1947.) Discussion on 205 of 1946.

621.316.54+621.318.5 562
Glass-Sealed Switches and Relays—C. G. McCormick. (*Bell Lab. Rec.*, vol. 25, pp. 342–345; September, 1947.) The dry-reed and mercury-contact types of switch are discussed and performance figures given. They are desgined to withstand extreme climates.

621.316.722 563
Optimum Parameter for Gas Tube Voltage Regulators—W. R. Berg. (*Electronics*, vol. 20, pp. 136, 138; October, 1947.) Determination of values of the circuit constants for maximum regulation for gas-tube voltage regulators with linear current versus voltage characteristic.

621.396.68:621.316.722.1 564
The Stabilization of Power Supplies in Radio Technique—J. Moline. (*Radio Franç.*, pp. 16–19; September, 1947.) Discussion of various regulation systems of the variable series or shunt impedance type, with practical examples.

621.396.69 565
The Reconstruction of Leafield Long Wave Radio Masts—J. P. Harding and J. F. Harmon. (*P.O. Elec. Eng. Jour.*, vol. 40, parts 1 and 2, pp. 1–7 and 63–68; April and July, 1947.) The considerations which led to the decision to reconstruct in reinforced concrete the ten 305-foot tubular steel masts at Leafield radio station are discussed and the design problems which this entailed are considered. An account is given of the organization and constructional methods employed on site.

TELEVISION AND PHOTOTELEGRAPHY

621.396/.397](47) 566
Broadcasting and Television Methods in the Soviet Republics—Huth. (*See* 535.)

621.397.3 567
Finch Facsimile-in-Color Process—(*Tele-Tech*, vol. 6, p. 29; September, 1947.) The picture to be transmitted in color is scanned with red, blue, yellow, and white light. At the receiver, the picture is dotted on a linear raster by red, blue, yellow, and black pencils mounted in a turret rotating in synchronism with the scanner. A solenoid-operated device presses the correctly colored pencil against the paper at the correct moment. Picture definition is 100 lines per inch and the speed of reproduction 4 square inches per minute (one-half-inch travel of 8-inch roll). See also *Electronics*, vol. 20, pp. 104–105; October, 1947.)

621.397.3 568
Colorimetry in Television—W. H. Cherry. (*RCA Rev.*, vol. 8, pp. 427–459; September, 1947.) Colorimetrically exact reproduction of color in simultaneous television is now possible. The basic concepts and relations of trichromatic colorimetry are developed. See also 3297 of 1947 and 572 below.

621.397.5 569
American Television—M. Lorach. (*Télév. Franç.*, pp. 2–6, 2–5, 8, 6–10, 6–8, and 8–11, 18; June–October, 1947.) A general description of the principal features of the various systems, with some details of aerials, cable and radio links, relay stations, special transmitting tubes, service areas, etc. To be continued.

621.397.5:535.37:621.385.832 570
Application of I.C.I. Color System to Development of All-Sulfide White Television Screen—A. E. Hardy. (*RCA Rev.*, vol. 8, pp. 554–563; September, 1947.) Full paper; summary noted in 4051 of January.

621.397.5:621.396.65 571
Television Radio Links—(*Electrician*, vol. 138, p. 1769; June 27, 1947.) Some details of the equipment installed on Danbury Hill, near Chelmsford, for relaying the television transmissions from Alexandra Palace, 31 and one-half miles away, to Great Bromley, near Colchester, 24 miles beyond Danbury Hill. For another account see *Electronic Eng.* (London), vol. 19, p. 240; August, 1947.)

621.397.6 572
An Experimental Simultaneous Color-Television System—R. D. Kell, G. C. Szlikai, R. C. Ballard, and A. C. Schroeder, K. R. Wendt, and G. L. Fredendall. (Proc. I.R.E., vol. 35, pp. 861–875; September, 1947.) The paper describes a system in which the three primary color pictures are transmitted simultaneously. The standard scanning speeds are used so that a monochrome picture can be received on present receivers.

Pickup equipments for both films and live subjects are described. The film is scanned by means of a flying-spot kinescope. The transmitted light is divided into the three primary colors by dichroic mirrors and the three light beams are converted into video signals by multiplier photo cells. The live-subject equipment is similar in principle, the light reflected from the subject being picked up by a bank of red-, green- and blue-filtered photo cells. The development of the kinescope, the video amplifiers, correction circuits, and the construction of the equipment are described.

For the transmission of the three video signals a subcarrier system is used, the red subcarrier frequency being 8.25 Mc. and the blue 6.25 Mc. The sound is inserted between the green and the blue channels on a 4.5-Mc. subcarrier. The reproduction system consists of a three-gun kinescope which produces three separate images on different areas of the tube face. The images are filtered to produce the three colors and combined by a system of mirrors and a lens to form a registered color image. Details of the kinescope and the associated circuits are given.

621.397.6 573
Magnetic Deflection of Kinescopes—K. Schlesinger. (Proc. I.R.E., vol. 35, pp. 813–821; August, 1947.) The energy of the deflecting field is calculated for various deflecting angles and beam voltages. The efficiency of sweep generators, having the preferred positive rate of change of anode current, is discussed. Transients during the retrace and their elimination are considered; a special damping method by secondary emission within the output tube is described. Some basic forms of sweep distortion are discussed and means for their correction are indicated. Finally, a sweep circuit of improved efficiency is described. Flyback energy is rectified and the resulting d.c. power added to the anode power supply. See also 3220 of 1946 (Sziklai), 272 of 1947 (Cocking) and back references, and 3305 of 1947 (Friend).

621.397.6:621.385.832 574
Magnetic-Deflection Circuits for Cathode-Ray Tubes—O. H. Schade. (*RCA Rev.*, vol. 8, pp. 506–538; September, 1947.) In principle, an ideal cyclic system for deflecting an electron beam requires only wattless power. A practical system may be based on the fact that the inherent capacitance associated with a deflecting circuit will form a tuned circuit with the deflecting coil. This will rapidly reverse the field in the deflecting coil when the energizing potential is removed. An electronic switch can be used to control this potential. The graphical repreentation of the circuit resistance as a load line in the anode characteristics of electron tubes functioning as an electronic switch, furnishes an accurate means of obtaining operating conditions and specifications for the design of practical tubes and circuits.

"A substantial fraction of the circulating power in certain deflecting systems can be recovered as d.c. power output from the circuit and, by the use of specific transformation ratios, may be recirculated through the system."

621.397.61 575
Power Stage of a Television Transmitter—M. R. Labadie. (*Télév. Franç.*, pp. 10–12 and 12–15; June and July, 1947.) A general treatment of the inverse amplifier with cathode excitation, and discussion of neutrodyning, modulation, anode circuit, choice of tubes and results obtained with the Eiffel Tower transmitter.

621.397.62 576
RCA Television Receiver, "Popular" Type—(*Radio Franç.*, pp. 23–25; September, 1947.) A short general description, with detailed circuit diagram.

621.397.62 577
A Modern High-Quality Television Receiver for the Home Constructor—W. I. Flach and N. H. Bentley. (*Electronic Eng.* (London), vol. 19, pp. 320–321; October, 1947.) Photographs and a few details of a receiver which is fully described in a booklet entitled "A Modern Home-Built Televisor," published by *Electronic Eng.*

621.397.62:535.88 578
Optical Design of Philco Television Projection Receiver—W. F. Bradley and E. Traub. (*Tele-Tech*, vol. 6, pp. 36–40, 105; September, 1947.) Supplementing preliminary details noted in 2963 of 1947. For analysis of wide aperture optics see also 3587 of 1945 (Epstein and Maloff).

621.397.62:621.396.828 579
Interference with Television Broadcasting—G. Grammer. (*QST*, vol. 31, pp. 24–30; September, 1947.) Amateur interference with television is principally, though not entirely, a question of transmitter harmonics. The various circuits of television receivers are discussed with a view to finding out what frequencies are likely to cause interference. Details are given of experiments carried out by the Central Jersey Radio Club to trace the causes of individual cases of interference and, if possible, to find remedies. See also 254 of February (Seybold).

621.397.62:621.396.828 580
Electrical Interference Suppression in Television Receivers—W. I. Flach. (*Electronic Eng.* (London), vol. 19, pp. 326–327; October, 1947.) Tests on a receiver near a main road showed that frequently the sound was completely swamped, while about halt the picture was obliterated. Considerable improvement was effected, for sound, by the use of diode limiters similar to those used in Pye post-war and in Murphy receivers. Video interference was reduced by limiting to the peak-white level. Suitable circuits are given.

621.397.645:621.397.62 581
Wide-Band Amplification, by Wobbulation

of the Carrier Wave, in the "Ontra" Receiver—(*Radio Franç.*, pp. 21–22; September, 1947.) The values of R and C for the oscillator of the frequency changer are so chosen that the tube functions as a blocked oscillator. In addition, the capacitance of the tuned circuit is reduced to a minimum, so that the circuit is practically tuned by the tube input capacitance. Under these conditions a sawtooth wobbulation of the frequency is automatically obtained; it is thus possible to use i.f. tuned circuits of high impedance, of the order of 20,000 ohms for circuits tuned to 15 Mc. The "Ontra" video receiver uses only three tubes, an ECH3 as oscillator and frequency changer, a high-slope i.f. tube and an EBL1 as detector and video amplifier.

621.397.7 582
Plan for a Television Station—N. Q. Lawrence. (*Electronic Eng.* (London), vol. 19, pp. 322–324; October, 1947.) A survey of the general requirements for flexible and smooth working, with a description of a model of a station in which five studios, of different sizes, are conveniently arranged round a central control tower. Use of the basement for artists' accommodation, with direct access to each studio, permits free movement of performers and operational staff. Audiences are restricted to the first floor.

621.397.828:621.396.1 583
Engineering Problems Involved in TV Interference—Francis. (*See* 538.)

621.397.5 584
Television Techniques [Book Review]—H. Bettinger. Harper Bros., New York, 1947, 237 pp., $5.00. (*Electronics*, vol. 20, pp. 260–261; October, 1947.) "Based on the production experience at WRGB, Schenectady. . . . Pictorial composition and continuity, visual and audio techniques, television script writing, producing, and directing are covered in the text."

TRANSMISSION

621.391.63:621.325.53 585
The Concentrated-Arc Lamp in a Light-Beam Communication System—W. D. Buckingham, C. R. Deibert, and R. V. Morgenstern. (*Elec. Eng.*, vol. 66, pp. 975–979; October, 1947.) For another account see 3691 of 1947 (Buckingham and Deibert).

621.396.61 586
The Practice of Frequency Modulation—R. Gosmand. (*Télév. Franç.*, pp. 14–16; September 1947.) Description with detailed circuit diagrams, of a transmitter with stabilized frequency and either a.m. or f.m.

621.396.61.029.56/.58 587
2, 6 and 10 [meters] with Crystal Control—J. Millen. (*QST*, vol. 31, pp. 66–70; September, 1947.) Circuit and operational details of a 10-meter transmitter with an input of about 100 watts. Operation on 6 meters makes use of a second crystal. For 2-meter operation, a frequency tripler circuit is used with the 6-meter oscillator.

621.396.61.029.58 588
Medium Power—Living-Room Style—L. T. Waggoner. (*QST*, vol. 31, pp. 37–46; September, 1947.) Circuit and constructional details of a 350-watt transmitter for c.w. or telephony, housed in a metal cabinet 36 and three quarter inches×21 and one half inches ×15 inches. Circuits are conventional and 1250 volt and 600-volt power supplies are included.

621.396.61.029.58 589
Revamping the 150-B for 14-Mc Operation—J. M. Murray. (*QST*, vol. 31, pp. 22–23; 120; September, 1947.) Details of alterations required to increase the maximum frequency from 12 to 14 Mc.

621.396.61.029.62 590
25 Watts H.F. on 60 Mc.—L. Liot. (*Télév. Franç.*, Supplement *Électronique*, pp. 1–4 and 10–12; June and July 1947.) A transmitter using $\lambda/4$ resonant bifilar lines for grid and anode tuning in the oscillator, the tube for which is a QQE 04/20. The modulation and feed circuits and aerial coupling are described in detail.

621.396.61.029.62 591
F.M. Broadcast Transmitters using Phasitron Modulation—L. O. Krause. (*FM and Telev.*, vol. 7, pp. 31–36, 54; June, 1947.) Design and circuit details for a series of commercial power amplifiers which may be combined to give outputs of 1, 3, 10, 25, and 50 kw. in the frequency band 88 to 100 Mc. The basic 250-watt f.m. exciter was described in 277 of 1947 and the theory of operation of the phasitron tube was given in 1405 and 2767 of 1946.

621.396.61.029.62 592
KSBR's 50-kw [100.5-Mc.] High-Band F-M Transmitter—R. L. Norton, B. O. Ballou, and R. H. Chamberlin. (*Electronics*, vol. 20, pp. 80–84; October, 1947.) Detailed description with drawings, photographs, and a block diagram. Special tank circuits and a new multiunit thoriated-filament triode, Type 3X12500A3, are used.

621.396.61.029.62:621.396.931 593
A One Kilowatt V.H.F. Frequency Modulated Transmitter—J. B. L. Foot. (*Jour. Brit. I.R.E.*, vol. 7, pp. 195–203; September, 1947. Discussion, pp. 203–204.) For high fidelity broadcasting.

621.396.61.029.64+538.569.4 594
On the Emission of Microwaves and Their Absorption in the Air—V. L. Ginsburg. (*Bull. Acad. Sci.* (U.R.S.S.), *sér. phys.*, vol. 11, no. 2, pp. 165–182; 1947. In Russian, with English summary.) An analysis of certain new methods of generating waves with wavelength <1 centimeter with particular reference to emission by a relativistic electron and by an electron moving near a dielectric.

621.396.611.33:621.396.671 595
The Matching Ranges of Transmitters—P. Mourmant. (*Radio Franç.*, pp. 12–15; September, 1947.) A discussion of the problem of matching a transmitter to an aerial by means of an interposed device. It is shown that in general matching devices must include at least two variable elements. When only two variable elements A and B are included, there are normally two pairs of values of A and B which satisfy the matching condition. To increase the matching range, it is often desirable to use more than two variable elements. See also 1958 of 1947 (Glazer and Familier).

621.396.615.17 596
A 5-kW Pulse Generator—L. Liot. (*Télév. Franç.*, no. 29, Supplement *Électronique*, pp. 33–35; September, 1947.) Uses an EE50 for the pulse circuit, an EL6 for the amplifier and an 807 for the modulator. An oscillator using two 955 tubes and fed from this modulator gives a peak h.f. power of about 300 watts at a frequency of 375 Mc.

621.396.619.13:518.61 597
Computation of the Solutions of $(1+2\epsilon\cos 2z)y''+\theta y=0$; Frequency Modulation Functions—McLachlan. (*See* 461.)

621.396.645:621.396.61 598
Characteristics of the Quadriline Amplifier—J. R. Day and M. H. Jennings. (*FM Telev.*, vol. 7, pp. 43–46; August, 1947.) A push-pull-parallel arrangement of four 4 to 1000 ampere internal anode tetrodes, using 4-wire balanced transmission lines as input and output circuits. These circuits are open at one end and closed at the other by short-circuits between adjacent line elements. The symmetry of the arrangement makes the external field very small, so that operation is substantially independent of surrounding screens and little useful power is lost by radiation or external dissipation.

VACUUM TUBES AND THERMIONICS

621.314.67 599
The Rating of Small- and Medium-Power Thermionic Rectifiers—H. T. Ramsay. (*Jour. I.R.E.* (London) part III, vol. 94, pp. 260–274; July, 1947.) It is suggested that the rating of such rectifiers could be improved by specifying recommended values of the ratio of the source to the load resistance (R_S/R_0), and the product of the load resistance and the admittance of the input capacitor ($_{\omega}CR_0$). This specification ensures that full-load operation is always accompanied by a particular set of waveforms. Tables are given from which the full-load performance of a tube may be evaluated from its basic properties. Fractional-load conditions are investigated, and performance curves indicating the best use of the tube at a particular fraction of its maximum output are deduced.

621.383.5 600
Electromotive Force and Internal Resistance of Blocking Layer Photo Elements—A. E. Sandström. (*Ark. Mat. Astr. Fys.*, vol. 34, part 2, section B, 7 pp.; September 25, 1947. In English.) The equivalent network of such elements is assumed to contain an e.m.f. E of internal resistance R_b, shunted by the barrier-layer resistance R_b', while an external resistance R_e represents that of the electrodes and semiconductor. Typical experimental values for R_b, R_b' and R_e are given and it is shown that E is always higher than the potential difference between the electrodes for zero current.

621.385:518.5 601
Tube Failures in ENIAC—F. R. Michael. (*Electronics*, vol. 20, pp. 116–119; October, 1947.) Analysis of the causes of 644 tube failures occurring in the 18,800-tube ENIAC (158 of February) during one year of operation. The five major causes of failure were found to be (a) open heater wire, (b) damaged oxide-cathode coating, (c) internal leads and supports dangerously close, (d) open electrode spot welds, and (e) burned-open shorts. Each is discussed with photographic examples. Experiments showed no difference in the rate of failure between tubes to which full heater voltage was applied at once and those receiving a gradual application of voltage. These results could be used to improve the odds against tube failure in industrial service.

621.385.029.63/.64 602
Kinetic Theory of the Exchange of Energy between an Electron Beam and an Electromagnetic Wave—A. Doehler and W. Kleen. (*Ann. Radioélec.*, vol. 2, pp. 232–242; July, 1947.) The alternating electron current resulting from the interaction of the electron beam and the electric vector of the traveling wave is first determined. From consideration of the transfer of energy, three waves are found to exist, traveling in the same direction as the electron beam; the amplitude of one wave is strongly amplified, and this wave predominates after traversing a sufficiently great path. The power gain is calculated in a homogeneous waveguide with and without attenuation. Discussion of the particular cases of a plane or cylindrical waveguide partially filled with a dielectric shows that the kinetic theory and the deductions from it are consistent with Maxwell's equations.

621.385.1 603
The Valves to be used in the [French] Receivers of Tomorrow—G. Giniaux. (*TT.S.F. Pour Tous*, vol. 23, pp. 177–179; September, 1947.) Future French tubes should embody recent improvements in construction introduced both in France and in other countiies. The advantages of the all-glass technique and "Rimlock" base are discussed briefly.

621.385.1:621.317.79:621.396.822 604
Measurement of Valve Background Noise—Chamagne and Guyot. (*See* 489.)

621.385.1:621.396.694.012.8+621.392 605
Circuits and Valves in Electronics—Charbonnier and Royer. (*See* 359.)

621.385.1.032.216 606
Oxide Cathodes. The Effect of the Coating-Core Interface on Conductivity and Emission—D. A. Wright. (*Proc. Roy. Soc. A*, vol. 190, pp. 394–417; August 12, 1947.) A potential barrier occurring at the interface between oxide and metal leads to a rectifier action restricting the flow of electrons from metal to coating. In a well-activated coating this restriction determines the thermionic emission that can be drawn from it and accounts for the rapid decay of emission immediately after the application of anode voltage.

621.385.1.032.216:535.215.9 607
Effect of Light on the Behaviour of Oxide Cathodes—J. Debiesse and R. Champeix. (*Compt. Rend. Acad. Sci.* (Paris), vol. 225, pp. 404–405; September 1, 1947.) The electron current between the cathode and a surrounding anode is found to increase appreciably when the cathode is illuminated, through a hole in the anode, by light from a Hg arc or a 150-watt lamp. The sensitivity as a photoelectric device, for the tubes used, was at a maximum for anode voltages of +30 to +50 volts and cathode temperatures from 850°K to 900°K.

621.385.4 608
Experimental Audio Output Tetrode—W. S. Brian. (*Electronics*, vol. 20, pp. 121–123; August 1947.) A tetrode in which the first grid is connected by a resistor to the 250-volt supply and acts as a space-charge grid, while the second grid is used as the control electrode. Harmonic distortion is less than with a beam tetrode.

621.385.832:[537.291+538.691 609
Electron Beam Deflection: Part 1—Small-Angle Deflection Theory—R. G. E. Hutter. (*Jour. Appl. Phys.*, vol. 18, pp. 740–758; August, 1947.) The "path" and "iconal" mathematical methods for studying the effects of electric and/or magnetic fields on electron beams are discussed.

"These methods are then applied to describe the action of balanced two-dimensional electric deflection fields on electron beams. It is shown that both methods yield essentially the same results. Expressions are derived describing the magnitude of deflection and the distortions of an electron beam.

"Only the 'path' method is used in a similar investigation of magnetic-type deflection fields." A summary of part of this paper and of part 2 (to be published later) was noted in 3346 of 1947.

621.396.615.141.2 610
On the Effect of an External Electromagnetic Field on a Split Anode Magnetron—S. Ya. Braude. (*Zh. tekh. Fiz.*, vol. 13, nos. 7/8, pp. 431–449, 1943. In Russian.) A magnetron with a static characteristic represented approximately by a polynomial of the 5th degree is considered. Formulas are derived determining the oscillations for various conditions and also the amplification factors for weak and strong signals.

621.396.615.142.2 611
Reflex Oscillators—J. R. Pierce and W. G. Shepherd. (*Bell. Sys. Tech. Jour.*, vol. 26, pp. 460–681; July, 1947.) A comprehensive account. A broad theoretical discussion is given first; reflex oscillators vary so widely in construction that theoretical results form a better basis for generalization about their properties than experimental results. Mathematical calculations are relegated to a series of appendices. Many factors, such as multiple transits of electrons, different drift times for different electron paths and space charge in the repeller region, are not ordinarily taken into account although they can be quite important. It would not be difficult to fit a large body of data to a theory, correct or incorrect, which takes into account all observed effects. The theory given must be regarded only as a guide to the capabilities of these oscillators and to their design rather than as an accurate, quantitative tool. The following oscillators developed at the Bell Telephone Laboratory are then discussed: (a) beating oscillators, (b) the 707, which has an external resonator, (c) the 723, which has an integral cavity, (d) the 2K29, designed to eliminate hysteresis, (e) the 2K25, a broadband oscillator, (f) the 2K45, thermally tuned, (g) the 2K50, with waveguide output, (h) the 1464, a millimeter-range oscillator, and (i) the 2K23 and 2K54, for pulsed applications.

621.396.615.142.2 612
On the Effects of Space Charge in Velocity-Modulation Valves with Drift Bunching—R. Warnecke, P. Guénard, and C. Fauve. (*Ann. Radioélec.*, vol. 2, pp. 224–231; July, 1947.) A discussion of the two-cavity klystron, used as an amplifier with high output and medium gain. Certain assumptions are made concerning the dimensions of the drift tube, the constitution of the electron beam, the magnetic focusing and the modulation of the initial velocity, and an approximate formula is derived for the fundamental component of the electron current in the two cases where the beam is (a) infinitely wide and (b) of limited cross section.

621.396.615.142.2 613
Facts about Klystrons—O. P. Ferrell. (*CQ*, vol. 2, pp. 16–17. 63; July, 1946.) Discussion of the theory of their operation, and their use for super-high-frequency amateur radio.

621.396.645.029.64 614
On the Theory of U.H.F. Amplifiers—Siforov. (*See* 379.)

621.396.694 615
Simplified Extrapolation Method of Obtaining the Saturation Curves of Transmitting Valves—R. Suart. (*Radio Franç.*,: pp. 21–24; July, 1947.)

621.396.615.142.2 616
Klystron Tubes [Book Review]—A. E. Harrison. McGraw-Hill, New York, 1947. 271 pp., $3.50. (*Electronics*, vol. 20, p. 258; September, 1947.) The general behavior, including the functions of cavities and electron beams, is explained for various types of klystron. Methods of modulation and measurement are also described.

MISCELLANEOUS

06.064 London:621.396 617
The Olympia Show—(*Wireless World*, vol. 53, pp. 362–386; October, 1947.) An illustrated stand-to-stand guide to the National Radio Exhibition, Olympia, London, October, 1947. See also *Electronic Eng.* (London,) vol. 19, pp. 306–319; October, 1947; *Elec. Times*, vol. 112, pp. 383–388; October 2, 1947; *Wireless Eng.*, vol. 24, pp. 333–342; November, 1947.

06.064 London:[621.396.621+621.396.69 618
Radio Exhibition [Olympia, London]—Miller. (*See* 524.)

06.064 Manchester:621.38/.39 619
Manchester Electronics Exhibition—(*Electronic Eng.* (London) vol. 19, pp. 296–297; October, 1947.) Brief descriptions of a few of the exhibits.

5+6](054) 620
Research: A Journal of Science and Its Applications—A new monthly journal whose first number appeared in October, 1947, published by Butterworths Scientific Publications Ltd, 4–6 Bell Yard, Temple Bar. London W.C. 2 The annual subscription is 45s. ($10 in U.S.A.). One of the aims will be to fill the gap uncovered by journals of learned societies, technical journals, and popular journals, so that the technical specialist may get a general idea of what is going on in other fields. It also has the object of helping the pure scientist to bridge the enormous gap between invention and production.

53 Langevin 621
The Scientific Work of Professor Paul Langevin—M. R. Lucas. (*Onde Élec.*, vol. 27, pp. 357–358; August–September, 1947.) A short account of Langevin's contributions to the kinetic theory of gases, the theory of diamagnetism and of paramagnetism, the theory of relativity, and the practical applications of supersonics.

539.17 622
Nucleonics—A new monthly journal with this title has been published since September, 1947; *Atomic Engineering* and *Atomic Power* are incorporated in it. The editorial and circulation offices are at 330 West 42nd St., New York 18; the subscription rate in the United States is $15 per annum.

621.316.98/.99 623
H. F. Testing of Lightning-Conductor Earths—V. Fritsch. (*Elektrotech. und Maschinenb.*, vol. 64, pp. 142–148; September–October, 1947.)

621.38(42)+621.38(73) 624
Current Overseas Technical Developments—J. N. Briton. (Proc. I.R.E. (Australia,) vol. 8, pp. 10–14; August, 1947. Discussion, pp. 14–15.) A survey of recent technical and commercial progress made in Britain and America, with special reference to the development of extended frequency range disk recording, monochromatic and color television and v.h.f. f.m. broadcasting systems.

621.396.69:384 625
Future Trend in Radio Component Design—G. A. T. Burdett. (*Tech. Bull. Radio Component Mfrs' Fed.*, vol. 1, pp. 2–5; September, 1947.) A high standard of technical reliability is now possible at a competitive price provided that sufficient quantities are run off at one time. The technical implications of this fact are discussed.

621.396.712 626
Guide to Broadcasting Stations [Book Review]—Iliffe and Sons, London, 3rd edn., 64 pp., 1s. (*Wireless Eng.*, vol. 24, p. 344; November, 1947.)

SECTION MEETINGS

ATLANTA SECTION

"Pulse DME in the Air Navigation Program," by H. I. Metz, Civil Aeronautics Administration; December 19, 1947.

BUFFALO-NIAGARA

"A Universal Auto Radio Receiver Featuring a New Type of Automatic Station Selection," by V. H. Wiley, Colonial Radio Corporation; December 17, 1947.

CHICAGO SECTION

"New Developments in Electronic Air Navigation and Traffic Control," by E. A. Post, United Air Lines, Inc.; November 21, 1947.

"Omnidirectional Range Multifrequency Navigation Receiver," by E. W. Sheridan, Bendix Radio Division; November 21, 1947.

"Signal Communication for Supreme Headquarters Allied Expeditionary Forces," by P. J. Moore, Signal Corps Reserve; December 19, 1947.

CLEVELAND

"Intermodulation Effects," by H. C. Williams, Ohio Bell Telephone Company; December 11, 1947.

CONNECTICUT VALLEY

"The Application of Capacitors for Power Factor Correction," by J. S. Williams, Westinghouse Electric Corporation; January 13, 1948.

DALLAS-FORT WORTH

"Navy Airborne Radar Equipment," by H. E. McDaniel, United States Navy; December 18, 1947.

"The Megacycle Meter," by J. B. Mintner, Measurements Corporation; December 29, 1947.

DAYTON

"Dielectric and Induction Heating," by H. Toner, Westingthouse Electric Company; January 6, 1948.

LOUISVILLE

"Operation of The Institute of Radio Engineers," by W. R. G. Baker, 1947 President, The Institute of Radio Engineers; January 9, 1948.

MONTREAL

"Design of Broadcast Antenna Tuning and Phasing Equipment," by E. Farmer, Canadian Marconi Company; January 14, 1948.

NEW YORK

"Industrial Application of High-Power Ultra-Sonics," by S. Y. White, Consulting Engineer; January 7, 1948.

PHILADELPHIA

"Reproduction of Sound," by H. F. Olson, RCA Laboratories; January 8, 1948.

PITTSBURGH

"Traveling Wave Tubes," by J. R. Pierce, Bell Telephone Laboratories; October 13, 1947.

"Recent Developments in Television Transmitter Equipment," by L. Mauntner, Allen B. DuMont Company; November 10, 1947.

"Electronic Heat," by H. Toner and M. Montague, Westinghouse Electric Corporation; December 8, 1947.

"Electronic Computers," by J. W. Mauchly, Eckert-Mauchly Computer Corporation; January 12, 1948.

PRINCETON

"Pulse Code Modulation," by L. A. Meacham, Bell Telephone Laboratories; December 11, 1947.

"Communications Theory," by W. G. Tuller, Melpar, Inc.; January 8, 1948.

(Continued on page 36A)

SECTION MEETINGS

(Continued from page 34A)

ROCHESTER

"The Future of Radio and Television in Rochester," by R. H. Manson, Stromberg-Carlson Company; December 9, 1947.

"Christmas, Today and Yesterday," by E. M. Poteat, Colgate-Rochester Divinity School; December 17, 1947.

"Behind the Instrument Dial," by L. F. Parachini, Weston Electrical Instrument Corporation; January 8, 1948.

SAN DIEGO

"Some Propagation Factors at Television and F.M. Frequencies," by L. G. Trolese, United States Navy Electronics Laboratory; December 2, 1947.

"Multivibrator Uses and Designs," by D. J. Green, Consolidated-Vultee Aircraft Corporation; January 6, 1948.

Election of Officers; January 6, 1948.

SEATTLE

"Problems in Radio Telemetering," by H. A. Price, Boeing Aircraft Company; December 29, 1947.

Election of Officers; December 29, 1947.

WASHINGTON

"Microwave Repeater Research," by K. G. Jansky, Bell Telephone Laboratories, Inc.; January 12, 1948.

WILLIAMSPORT

"Applications and Techniques of Printed Circuits and Miniature Electronics," by C. Brunetti, National Bureau of Standards; January 15, 1948.

SUBSECTIONS

FORT WAYNE

"Short-Time Motor Coast Measurements," by E. G. Downie, General Electric Company; November 24, 1947.

"The Technique of Radio Noise Suppression," by S. A. Zimmerman, General Electric Company; November 24, 1947.

NORTHERN NEW JERSEY

"New Developments in the Traveling-Wave Tube," by J. R. Pierce, Bell Telephone Laboratories; October 8, 1947.

"Television Field Strength Measurements in the New York Area, and Problems of Television Shared-Channel Interferences between Cities," by T. T. Goldsmith, Allen B. DuMont Laboratories; December 10, 1947.

STUDENT BRANCH MEETINGS

UNIVERSITY OF ALBERTA, I.R.E. BRANCH

"Magnetic Recording," by J. H. Scrimgeour, Student; January 14, 1948.

UNIVERSITY OF ARKANSAS, I.R.E. BRANCH

"Possibilities for Use of Electronic Control Devices in the Rice Industry," by C. G. Leonard, University of Arkansas; January 7, 1948.

(Continued on page 38A)

STUDENT BRANCH MEETINGS

(*Continued from page 36A*)

"Applications of Electronics to Agriculture," by P. H. Matisheck, University of Arkansas, January 21, 1948.

CARNEGIE INSTITUTE OF TECHNOLOGY, I.R.E.-A.I.E.E.

"Telephone Systems," by L. R; Pakkensen, D. C. Brewer and F. J. Rau, Students Dceember 3, 1947.

"Phototubes," by E. H. Fritz and W.W. Calloway, Students; December 3, 1947.

"Electrical Engineering as a Public Servant," by R. E. Rodger, G. F. Nielsen, and J. F. Rentz, Students; December 3, 1947.

"Individual Topics," by J. Lucas, R. Ketterer, and C. Jacoby, Students; December 10, 1947.

"E.E.—A Public Servant," by W. W. Ege, R. Crago, and W. Douglass, Students; December 10, 1947.

"C.I.T. Automatic Telephone," by J. Neenan, R. Langkamp, and R. Musgrave, Students; December 10, 1947.

"Geiger Counters," by R. J. Heh, C. W. Dutfett, and E. Brazon, Students; December 17, 1947.

"Individual Topics," by E. R. Hornbake, H. L. Brenton, and R. N. Frantz, Students; December 17, 1947.

"Transportation—EE Point of View," by H. I. Sherman, F. R. Lee, J. B. Ramsey, and F. L. Pirkheim, Students; December 17, 1947.

"Individual Topics," by C. F. Freund, C. D. Pollis, E. K. Weiss, and E. P. Worgs, Students; January 7, 1948.

"Insulation," by W. G. Gribble, J. H. Cronin, and S. W. Hunt, Students; January 7, 1948.

"Insulation," by H. K. George, V. A. Gibel, E. L. Johnson, and W. C. Tice, Students; January 7, 1948.

"Cathode-Ray Tubes," by J. McCauslin, D. F. Paull, and W. J. Keefer, Students; January 14, 1948.

"Railway Traffic Control," by H. Birnkrant, J. Augustine, and J. Barclay, Students; January 14, 1948.

"Medical Applications of Electrical Engineer ing," by J. Falgen, F. Dickson, and J. Kostyo, Students; January 14, 1948.

UNIVERSITY OF MICHIGAN, I.R.E.-A.I.E.E.

"Stroboscopes and High Speed Photography," by K. Adams, General Radio Company; January 13, 1948.

NORTH CAROLINA STATE COLLEGE, I.R.E.

Election of representatives to the North Caroline State College Engineer's Council; January 15, 1948.

NORTHWESTERN UNIVERSITY OF TECHNOLOGICAL INSTITUTE, I.R.E.-A.I.E.E.

Business meeting; January 15, 1948.

WAYNE UNIVERSITY, I.R.E.-A.I.E.E.

"Television," by L. Spragg, Radio Station W.W.J.; January 13, 1948.

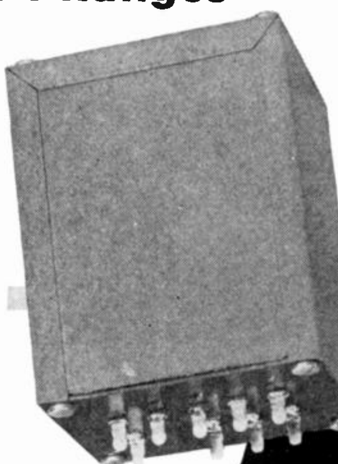
WORCESTER POLYTECHNIC INSTITUTE, I.R.E.-A.I.E.E.

"Electrical Problems in the Early Days of Power Distribution," by G. Hardy, Retired Engineer of the Worcester Electric Light and Power Company; January 6, 1948.

ADC TRANSFORMERS

Fill *Super* Specifications From Subsonic to Ultrasonic

For Ultrasonic Ranges

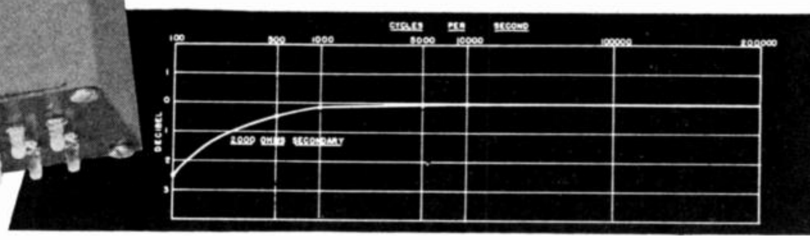


This **ADC** Transformer is custom-built to couple the output of an amplifier (30 watts) to a transducer. Impedance Ratio is 2500 ohms (4-2A3) to 500/700/1000/1500/2000 ohms. Transformers have been designed at **ADC** to operate up to 3 mc with useful band width in excess of 1:1000.

Whenever your equipment requires the unusual in transformers, **ADC** has the ingenuity, skill and capacity to produce them. If your problem requires . . .

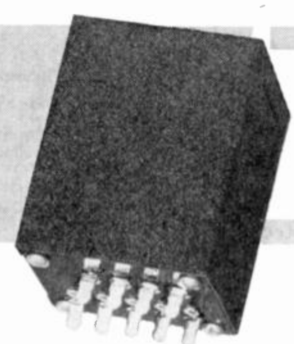
- **reliable performance**
- **extreme compactness**
- **unusual frequency ranges**
- **hermetic sealing (for severe service conditions—commercial or Army-Navy)**

. . . it will pay you to submit your specifications to **ADC** for reasonable prices, quality products and prompt delivery.



For Audible Ranges

This transformer has no unusual electrical properties, but it was designed for extreme dependability. It is an output from pp 6V6 to line, for voice range only (1 db—150 to 4000 cps). It was ordered from ADC simply because the equipment manufacturer required *unfailing* performance.



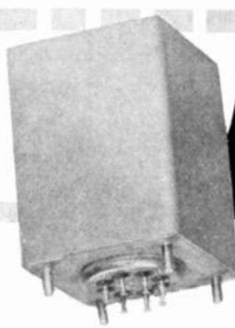
All **ADC** transformers have built-in reliability . . . a feature especially necessary for radio broadcasting, communications, wire recording, telemetering equipment, etc. A slightly higher original cost is more than offset by the dependability and quality of **ADC** design and manufacture.

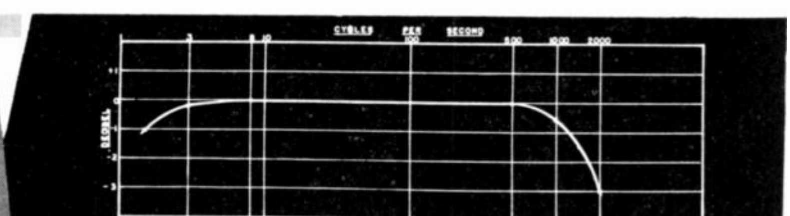
For Subsonic Ranges

(such as geo-physical work)

This transformer operates from pp plates (20,000 ohms) to pp grids (320,000 ohms) down to 2 cps. Secondary inductance is over 60,000 henries. It also has tertiary low impedance winding. Hermetically sealed—10 cubic inches.

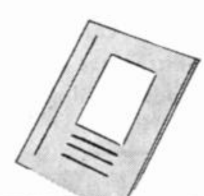
ADC has designed and made many low frequency transformers—some to operate from frequencies as low as 0.1 cps.





HERE are several charts showing characteristics of unusual transformers developed by ADC engineers. Many transformers have a wide range of requirements shown in the ADC Transformer Catalog. If you do not have your copy—*WRITE TODAY FOR CATALOG NUMBER 46-S*

Send us your special specifications for prompt and able transformer service.



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MEMBERSHIP

The following transfers and admissions were approved on February 3, 1948, to be effective as of March 1, 1948:

Transfer to Senior Member

Albersheim, W. J., Bell Telephone Laboratories, Inc., Deal, N. J.
Caldwell, P. G., General Electric Co., Electronic Park, Syracuse, N. Y.
Clark, D. E., 2501 N. Keeler Ave., Chicago, Ill.
Dempster, B., 2008 West Seventh St., Los Angeles, Calif.
Fano, R. M., Massachusetts Institute of Technology, Cambridge, Mass.
Fisher, F. J., 1 Richard Rd., Port Washington, L. I., N. Y.
Grundmann, G. L., 16 Oriental Ave., Westmont, N. J.
Guillemin, E. A., Massachusetts Institute of Technology, Cambridge, Mass.
Hachemeister, C. A., 85 Livingston St., Brooklyn, N. Y.
Hodson, W. G., 524 Hampton Rd., Burbank, Calif.
Johnson, E. O., 285 Merion Ave., Haddonfield, N. J.
Kilheffer, L. D., 804 Buckingham Rd., Dayton, Ohio
Koch, W. R., Victor Division 5-3, RCA Laboratories, Camden, N. J.
Mautner, R. S., 201 West 16 St., New York, N. Y.
Morgan, M. G., Thayer School of Engineering, Dartmouth College, Hanover, N. H.
Palmateer, R. E., Sylvan Heights, Emporium, Pa.
Sanborn, J. W., 15 Morris St., Merchantville, N. J.
Schultz, M. A., 635 Cascade Rd., Pittsburgh, Pa.
Skrivseth, A. G., 203 Rogers Dr., Fenwick Park, Falls Church, Va.
Tanner, R. H., 292 Charles St., Belleville, Ont., Canada
Taylor, H. A., RCA Communications, Inc., Riverhead, L. I., N. Y.
Wicks, D. S., 3895 Rodman S., Washington, D. C.

Admission to Senior Member

Isely, F. C., 3131 Westover Dr., S.E., Washington, D. C.
Kennedy, T. R., Jr., 601 West 112 St., New York, N. Y.
Meyer, R. B. 4702 Brandywine St., N.W., Washington, D. C.
Shaw, V. G., Electricl Engineering Department, Carnegie Institute of Technology, Pittburgh Pa.
Word, J. A., 238 Washington Ave., Chatham, N. J.

Transfer to Member Grade

Beins, J. K., 435 Kenilworth Ave., Toledo, Ohio
Broersma, C. B., c/o Radio-Holland, 562 Keizersgracht, Amsterdam C, Netherlands
Hicken, J., 40 Andrews Ave., Binghamton, N. Y.
Karlson, A. T., Severen Heights, Severna Park, Md.
Kidd, W. E., 585 Turner Ave., Glen Ellyn, Ill.
Kopecky, G. F., 304 E. 77 St., New York, N. Y.
Maron, M., Room 303, YMCA, Passaic, N. J.
Peach, S. E., Baldwyn, Miss.
Robinson, A. J., Jr., 226 W 5 St., Emporium, Pa.
Senn, G. F., 81 Garden Rd., Red Bank, N. J.
Shaw, R. G., 1406 Benton Way, Los Angeles, Calif.
Taylor, G. O., 5244 Arrow Rd., Cincinnati, Ohio
Ward, J. F., London House, Guilford St., W. C. 1, London, England
Williams, G. T., Box 440, Sect. 69, Anchorage, Alaska

(Continued on page 42A)

MEMBERSHIP

(Continued from page 40A)

Admission to Member Grade

Arnett, H. D., 18 Danbury St., S.W., Washington, D. C.
Bose, L., 34 Allenby Rd., Calcutta 20, India
Bryant, V. D., 22 High St., Franklin, Ohio
Burke, J. F., 33-17-201 St., Bayside, L. I., N. Y.
Capelli, M. P., 12, Cambridge Dr., Potters Bar, Middx., England
Carlander, A., Ymsenvagen 10, Enskede, Sweden
Carson, R. E., R.F.D., Maple Lane, N. Syracuse, N. Y.
Castilla, A., Calle Evaristo S. Miguel 20, Madrid, Spain
Clute, D. G., 520 Forrer Blvd., Dayton, Ohio
Cramer, R. E., Jr., 500 Jessamine Ave., West Collingswood, N. J.
Eland, R. C. A., 847 E. Chelten Ave., Philadelphia, Pa.
Frey, H. B., Jr., 163 Wade Lane, Oak Ridge, Tenn.
Green, M., North-West Telephone Company, 1955 Wylie St., Vancouver, B. C.
Hope, R. S., Thom & Smith Pty. Ltd., 919–929 Botany Rd., Mascot, N. S. W.
Houghton, E. G., Headquarters Fifteenth Air Force, A3CM, Colorado Springs, Colo.
Kelleher, K. S., 825 Church St., Alexandria, Va.
King, E. F., 3171 Federal Ave., Los Angeles, Calif.
Simler, L. L., 1711 Wait St., Seattle, Wash.
Stovall, J. R., Jr., 500 N. 12 St., Philadelphia, Pa.
Sullivan, E. F., 105 Harrison St., Oak Park, Ill.
Warchol, E. J., 1442 E. 29 St., Tacoma, Wash.
Webb, E. L. R., Electrical Engineering & Radio Division, National Research Council, Sussex St., Ottawa, Ont., Canada

The following admissions to Associate were approved on February 3, 1948, to be effective as of March 1, 1948:

Abernathy, H. D., 4927 Byers, Ft. Worth 7, Tex.
Albert, S. L., 1655 N. Cherokee St., Hollywood 28, Calif.
Aldridge, K. S., Casilla No. 2562, Santiago, Chile, S. A.
Anderson, M. E., 1906 Grismer Ave., Burbank, Calif.
Anderson, R. H., 1400 N. State Pkwy., Chicago 10, Ill.
Augenblick, H. A., Jr., 67 S. Munn Ave., East Orange, N. J.
Austin, F. H., 62-34—60 Ave., Maspeth, L. I., N. Y.
Aveni, G., 1966 S. Normandie Ave., Los Angeles 7, Calif.
Bailey, H. R., 52 Fernwood Ave., Dayton 5, Ohio.
Bates, A. G., c/o Radio Station KFAB, Omaha, Neb.

Bauman, J. P., 1207 Boynton Ave., New York 59 N. Y.
Bird, D. W., 620 E. Bridge St., Blackwell, Okla.
Blankenship, W. B., 2708 Ave. V, Lubbock, Tex.
Boellhoff, L. E., Box 32, Montandon, Pa.
Brennan, R. F., 200 E. 38 St., New York 16, N. Y.
Bricker, L., 2926 W. 24 St., Brooklyn 24, N. Y.
Brougher, R. M., 2912 Robinhood, Houston 5, Tex.
Casen, J. G., 109-01—72 Rd., Forest Hills, L. I., N. Y.
Castle, J. L., 218, Ash-Grove, Heston, Middx., England
Christiansen, T. A., 1929 Morse Ave., Chicago 26, Ill.
Coffin, B. E., Jr., 734 N. 20 St., Philadelphia 30, Pa.
Cooper, P. L., Jr., 980 Thomas Rd., Beaumont, Tex.
Cordella, D. P., 3605 Arthur Ave., Brookfield, Ill.
Coyne, J. M., Jr., 4429½ N. Kedzie Ave., Chicago, Ill.
Crawford, A. L., 1931 N. Tibbs Ave., Indianapolis 22, Ind.

(Continued on page 44A)

MEMBERSHIP

(Continued from page 42A)

Dathe, L. C., 502 Burlington Ave., Bradley Beach, N. J.
Dantine, W. A., 1211 San Pasqual, Pasadena 5, Calif.
David, E. E., Jr., 607C, Graduate House, Massachusetts Institute of Technology, Cambridge 39, Mass.
Dooley, L., 1940 E. Tremont Ave., New York 62, N. Y.
Drechsel, R. E., 205 W. Willow, Prospect Hts., Ill.
Everson, C. T., 321—22 St. N.E., Cedar Rapids, Iowa
Fanning, G. B., 527 Aberdeen Ave., Dayton 9, Ohio
Fink, E. A., 5603 Arlington St., Philadelphia 31, Pa.
Fischer, R. E., 4424 N. Clifton Ave., Chicago, Ill.
Foster, W. E., Hazeltine Electronics Corp., 58-25 Little Neck Pkwy., Little Neck, L. I., N. Y.
Friedman, I. B., 5114 S. Kimbark Ave., Chicago 15 Ill.
Fritsch, P. C., 5814 Third Pl., N.W., Washington 11, D. C.
Goldstein, B. M., 1820 Bryant Ave., New York 60, N. Y.
Grant, R. L., 1048 N. Front St., Sunbury, Pa.
Hamilton, D. H., Jr., 1320 Fairmont St., N.W., Washington 9, D. C.
Hatzakortzian, H. T., 424 E. 174 St., New York 57, N. Y.
Heald, E. T., 1313 Third Ave., S.W., Cedar Rapids, Iowa
Henley, E. J., 4208 Ivy St., East Chicago, Ind.
Hiller, F. L., 1288 French Ave., Lakewood, Ohio
Horniman, N. L., 528 Seventh St., Honolulu 57, T. H.
Hostetler, W. E., 5 Winthrop Ter., East Orange, N. J.
Hoyt, W. A., R.F.D. 3, Angola, Ind.
Huckaby, J. H., 67 Waverly Ave., Dayton 5, Ohio
Hymen, C., 686 McDonough St., Brooklyn, N. Y.
Ingram, C., 4746 S. Indiana, Chicago 15, Ill.
Johnson, M. P., 502 Newland Ave., Jamestown, N. Y.
Jorgensen, B., 365 Midland Ave., Syracuse 4, N. Y.
Kaufmann, W. S., 432 Haddon Ave., Camden, N. J.
Kellerman, R. B., Castle Creek Rd., Castle Creek, N. Y.
Keys, D. D., 4 Monroe, Denver 6, Colo.
Langan, J. E., 985 Amsterdam Ave., New York 25, N. Y.
Laymance, T. D., 1117 Elgin St., Houston 4, Tex.
Leferson, J., 41 Weedhill Ave., Springdale, Conn.
Lord, D. L., 107 N. Inglewood Ave., Inglewood, Calif.
Luecke, M. W., 2113 A West Galena St., Milwaukee 5, Wis.
Martin, L. H., N. Z. Broadcasting Unit, 2 NZEF (JAPAN) B.C.O.F., c/o APO 301, Japan
McBeath, H. M., Jr., 302 Palm St., Abilene, Tex.
McCracken, L. G., Jr., Ordnance Research Laboratory, State College, Pa.
McCusker, R. W., 15 Walker Ave., Pikesville 8, Md.
McKay, G. C., Jr., 602 Reid St., Houston 9, Tex.
Mead, F. S., 4420 Lily Ponds Dr., N.E., Washington 19, D. C.
Mercado, C. R., Box 156, Barceloneta, Puerto Rico
Merchant, V. V., 25-27 Clinton St., Brooklyn 2, N. Y.
Midcalf, W. A., 6217½—11 Ave., Los Angeles 43, Calif.
Miller, C. E., Box 516, Route 7, Mt. Clemens, Mich.
Miller, L. C., 232 E. 81 St., New York 28, N. Y.
Mitchell, J. L., 2802 LaBranch, Apt. 1, Houston 4, Tex.
Moiseve, C. J., 3707 Hope St., Huntington Park, Calif.
Mass, J. W., 1836 W. Maypole Ave., Chicago 12, Ill.
Noil, P. L., Route 10, Green Haven, North Kansas City, Mo.

(Continued on page 46A)

Membership

(Continued from page 44A)

Nelinson, S., 206 Bowers St., Jersey City 7, N. J.
Nelson, E. A., 2104 Jefferson Ave., St. Paul 5, Minn.
Nelson, H. G., 219 N. Yale, Vermillion, S. Dak.
Nelson, J. L., 7402 Bay Pkwy., Brooklyn, N. Y.
Osterberg, E. K., 2151 E. 96 St., Chicago 17, Ill.
Panian, R. M., 31 Gebhart St., Dayton 10, Ohio
Parker, W. L., 3906—50, Des Moines 10, Iowa
Pattison, J. W., Josephine, St. Wingham, Ont., Canada
Pettingill, S. A., 1809 G St. N.W., Apt. 108, Washington, D. C.
Ramamoorti, K. V., 10, Edward Elliots Rd., Mylapore, Madras, India
Reed, J., 88 Hillcrest Rd., Belmont 78, Mass.
Reinheimer, H. J., Jr., 45 Haight Ave., Poughkeepsie, N. Y.
Rew, T. G., 222 Shotwell Park, Syracuse 6, N. Y.
Ritter, G., Aeroporto Federal, Porto Alegre, Rio Grande Do Sul, Brasil
Rotenberg, A. B., 10750 Barman Ave., Culver City, Calif.
Schwartz, M. J., 302 Fifth St., Angola, Ind.
Sharp, H., Box 960, Denver 1, Colo.
Shirley, R. C., 604 Harper Ave., Jenkintown, Pa.
Shook, W. A., 6708—39 S.W., Seattle 6, Wash.
Silverman, H. M., 33 Crawford St., Roxbury 21, Mass.
Slagel, R. J., Jr., 6345 Dante Ave., Chicago 37, Ill.
Sreekantan, B. V. P., Sadvaidyasala Branch, Chickpet, Bangalore, Mysor State, India
Steeb, E. C., Jr., 15 Groveland Ave., Buffalo 14, N. Y.
Stormer, W. J., 1223 F St., N.E., Washington 2, D. C.
Stotler, D. E., 7034 Quince, Houston 17, Tex.
Sudla, A. T., 113 Shaw St., Garfield, N. J.
Trapp, J. A., Electrical Engineering Department, Iowa State College, Ames, Iowa
Uphoff, R. L., 812 Park Ave., Plainfield, N. J.
Veit, F., 78-08—75 St., Brooklyn 27, N. Y.
Viles, F. L., Box 1005, Route 5, Vancouver, Wash.
Volk, N., 32 E. Hill St., Baltimore 30, Md.
Walker, C. E., 165 Fountain Ave., Glendale, Ohio
Walkup, J. F., 7742 Westover Rd., Overland Park, Kan.
Ward, W. P., 1881 Cornelia St., Ridgewood, Brooklyn 27, N. Y.
Weagel, K. D., 30 N. Ogden Ave., Chicago 7, Ill.
Webster, W. M., 22 E. Stanworth Dr., Princeton, N. J.
Weeks, A. D., 818 N. Second St., Alhambra, Calif.
White, L. B., 4803 Arvilla Lane, Houston, Tex.
Whittier, M. L., 43-G. Jones Dr., Bremerton, Wash.
Willhite, G. B., 834 Arlington St., Houston 7, Tex.
Woodbrey, C. S., 18 Ridge Rd., Farmingdale, L. I., N. Y.
Woody, F., 4340 Kenmore, Apt. 22, Chicago 13, Ill.

ERRATUM

The following membership was erroneously listed and should read as follows:

Admission to Senior Member, effective as of February 1, 1948, Umpleby, K. F., Research Department-Bendix Radio Corporation, Baltimore, Md.

News—New Products

(Continued from page 26A)

New Terminal Block

A new feed-through terminal block which meets the need for subpanel and chassis construction with combination screw and soldered terminals has been developed by the **Curtis Development and Manufacturing Co.**, 1 North Crawford Ave., Chicago 24, Ill., which has its factory at Milwaukee 10, Wis. This new type of feed-through terminal block has ample clearance and leakage distances for use in circuits carrying up to 300 volts at 20 amperes.

(Continued on page 48A)

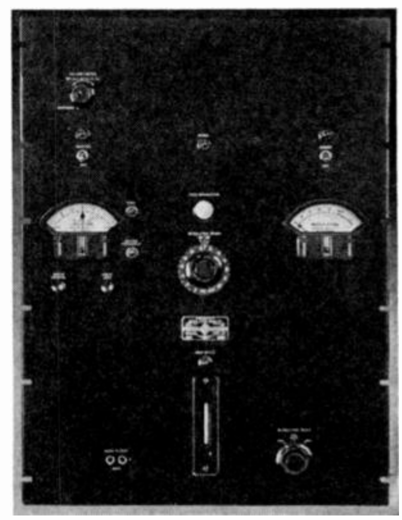
News—New Products

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

(Continued from page 46A)

New F.M. Monitor

The new f.m. monitor, manufactured by **General Radio Co.**, 275 Massachusetts Ave., Cambridge 39, Mass., is designed for monitoring f.m. broadcast and television audio transmitters. It provides a continuous indication of center frequency; a meter indication of percentage modulation, positive, negative, or peak-to-peak; and a lamp indication of peaks in excess of a predetermined percentage.



The stability of center-frequency indication is comparable with that obtained on a.m. monitors in the standard broadcast band, so that no calibration checks need to be made during the operating day, and hence a remote indicator can be used at the transmitter engineer's desk.

The Monitor uses a counter-type discriminator which not only permits the use of a low intermediate frequency, with a resulting high degree of stability, but also because of its inherent linearity keeps distortion at a minimum. Two audio output systems are provided, one for measuring distortion and noise with the Type 1932-A Distortion and Noise Meter, and the other for audio monitoring. Inherent distortion is less than 0.2 per cent and distortions as low as 0.5 per cent can be measured accurately, according to the manufacturer. The noise level is at least 75 db below 100 per cent modulation.

The panel is 19 by 26 inches, and the depth behind the panel is 13½ inches overall. The instrument weighs 88 pounds.

(Continued on page 65A)

We will be grateful if you will mention PROCEEDINGS of the I.R.E. when writing to our advertisers.

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.

ENGINEERS

Microwave engineers wanted. Laboratory experience essential (Industry or Government). Communications for video. Permanent. Salary relatively high. You are invited to visit our modern plant and talk to our engineers, or write us your job history and education. Motorola, Inc., 4545 W. Augusta Blvd., Chicago 51, Illinois, Att: Mr. E. Dyke.

SCIENTISTS AND ENGINEERS

Wanted for research and advanced development work in the fields of microwaves, radar circuits, gyroscope systems, servomechanisms and general electronics. Scientific or engineering degrees required. Salary commensurate with experience and ability. Inquiries should be directed to Manager, Engineering Personnel, Bell Aircraft Corporation, Buffalo 5, N.Y.

ENGINEERS AND PHYSICISTS

Several openings available in radar and in medium and ultra high frequency design work. Positions are in the Engineering Department of a progressive middle western manufacturer. Physicists will find unusual opportunities in the Research Department. A degree from a recognized engineering college is essential as well as good industrial experience. Address reply to Collins Radio Company, Cedar Rapids, Iowa.

ELECTRICAL ENGINEERS AND PHYSICISTS

An expanding program of teaching and research has created opportunities as instructor, assistant professor and associate professor level in a large mid-eastern college. Your inquiries are invited. Box 500.

SALES ENGINEER

Manufacturer's representative wants assistant to cover Connecticut territory. Must have sales experience and knowledge of electronic components. Salary, expense allowance and commission. Please furnish complete background information. Box 501.

MECHANICAL ENGINEER

Mechanical engineer to prepare technical manuscripts covering certain operations of the Los Alamos Laboratory. Applicant must have a B.S. degree in mechanical engineering and considerable experience in engineering and technical writing. Interested persons may write directly to Employment Director, P.O. Box 1663, Los Alamos, New Mexico.

(Continued on page 54A)

SMALLER, LIGHTER
"EVEREADY"
"A-B" BATTERY
for more compact portables!

Will outlast any other battery pack of its size!

SPECIFICATIONS:

Voltage:
"A"-9, tapped at 7½.
"B"-90.

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

Net weight:
4 lbs. 15 ozs.

The No. 753 "Eveready" combination "A-B" 90-volt battery pack provides plenty of power for the more compact "pick-up" portable radios. It will last longer than any other "A-B" pack of comparable size.

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

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

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

EVEREADY
TRADE-MARKS
MINI-MAX
RADIO BATTERIES

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

NATIONAL CARBON COMPANY, INC.

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

Unit of Union Carbide and Carbon Corporation

UCC

Positions in APPLIED RESEARCH . . . .
INSTRUMENT DESIGN . . . .
ELECTRONIC DEVELOPMENT.

are available to Physicists and EE's with Bachelor, Master or Ph.D. degrees. Assignments are in radar, transmitters, receivers, antennas, electronic instruments, and field intensity measurements.

Scientists now on our staff are interested not only in developing new devices, but also in exploring fully the principles behind them. These men, experts in their fields, have helped evolve an organization conducive to success in both undertakings.

Employment at this Laboratory provides:
- Opportunity for professional advancement.
- Optional insurance and pension plans.
- Advantageous location for personal living, with ready access to engineering school graduate work.
- Recreational programs.

Visit us at the Commodore Hotel during the IRE convention; or phone Garden City 6880; or write to: PERSONNEL DEPARTMENT,

Airborne Instruments Laboratory
INCORPORATED
160 OLD COUNTRY ROAD · MINEOLA, N.Y.

ENGINEERS . . . We have immediate openings for electrical and mechanical engineers experienced in the design, development, and research of the following:

TRANSMITTERS	SCIENTIFIC INSTRUMENTS
RECEIVERS	TUBES
RADAR	RECORDING INSTRUMENTS
MOBILE COMMUNICATIONS EQUIPMENT	TELEVISION
ALLIED ELECTRONIC FIELDS	AVIATION EQUIPMENT

Address detailed replies to National Recruiting Division, RCA Victor, Camden 4, N. J.

RCA **RADIO CORPORATION of AMERICA**

POSITIONS OPEN

(Continued from page 50A)

ACOUSTICAL ENGINEER

Acoustical engineer wanted with experience in microphone or pickup design. Must know mechanical and acoustical circuits. Write details of experience and education to Engineering Department, Electro-Voice, Inc., Buchanan, Michigan.

ENGINEERS

(1) Mid-western manufacturer has opening for electronic engineer with background in electronic circuit design and instrumentation. Experience with pulse technique, servo-systems or telemetering procedures is desirable. Unlimited opportunity in a specialized field. Submit complete résumé and salary desired.

(2) Electrical designer with drafting experience and knowledge of mechanical layout. Experience in design of automatic test equipment desirable. State salary expected.

(3) Electro-mechanical draftsman. Must know symbols and be able to make composite layouts of electrical subassemblies. State salary desired. Write Box 502.

(Continued on page 56A)

LOS ALAMOS SCIENTIFIC LABORATORY

Has present need for physicists of Ph.D. grade with research and development experience in various phases of nuclear physics, electronics, optics, and physical chemistry. Both field and laboratory experimentation involved, depending on interests and capabilities. Also included is design and development of special laboratory electronic, mechanical and optical apparatus. Interviews at project expense can be arranged for qualified applicants.

Write direct to Employment Director, P.O. Box 1663, Los Alamos, New Mexico for further particulars, giving brief résumé of education and experience.

POSITIONS OPEN

(Continued from page 54A)

ELECTRONIC ENGINEERS

Opportunity for experienced electronic engineers in established and expanding development company. We design and produce electronic controls and computor equipment. Contact Electronic Associates, Inc., Long Branch, New Jersey. Telephone: L.B. 6-1100. Att.: Mr. Arthur L. Adamson.

DEVELOPMENT ENGINEER

Development engineer needed to design and develop electronic instruments for research work. Excellent opportunity for a man with a degree and practical experience in test equipment construction. Write, giving full details of education and experience to Personnel Office, University of Chicago, 956 E. 58 Street, Chicago 37, Illinois.

INSTRUCTOR

Southwestern church-related University. Man with Master's degree and teaching experience to teach radio theory and electronics. Salary range: $3,000-$3,300 for 9 months depending upon experience. Box 505.

(Continued on page 58A)

THE FINEST CORED SOLDER IN THE WORLD

ON DISPLAY
AT THE
IRE SHOW

Our claim is a simple one. We believe that Ersin Multicore is the finest cored solder in the world. If you are not already familiar with our product, we believe it can be of special assistance to you in your soldering processes whether you are manufacturing 10,000 radio receivers or repairing one ● Ersin Multicore is solder in the form of a wire containing 3 cores of non-corrosive Ersin Flux ● You get a guarantee of flux continuity. The Multicore construction gives you extra-rapid melting. Combined with a super active Ersin Flux—**Exclusive with Multicore**—you enjoy a speedy and consistently high standard of precision soldering ● Available in 5 alloys and 9 gauges. Please write for detailed technical information and samples.

Address U.S.A. and Canadian Inquiries to:
British Industries Corp.
315 Broadway, New York 7, N.Y.
Cables: Britind, New York

ERSIN MULTICORE

Regarding other territories:
MULTICORE SOLDERS LTD.
Mellier House, Albemarle St.,
London, W.1, England
Cables: Dustickon, Piccy, London.

POSITIONS OPEN

(Continued from page 56A)

ELECTRONICS ENGINEER

Electronics engineer for carrier telephone work by a St. Louis manufacturer of public utility equipment. Several positions open. Box 506.

SENIOR ENGINEER

Fine opportunity with a large midwestern radio corporation. Must have a minimum of three years' experience in loudspeaker design and materials used in manufacturing. College education in electronics or equivalent. In replying state age, education, experience and salary requirements. Box 507.

ENGINEERS

Technical college in northern New York is adding several staff members to the Department of Electrical Engineering. Men with some experience in research or design and development in communications —electronics or physics are preferred. Salary and rank will be commensurate with training and experience. Box 508.

ELECTRONICS ENGINEER

Graduate electronics engineer, experienced in the application of electronic controls on air born equipment. Some production and development experience is essential. In reply, give training, experience and other pertinent facts. Box 510.

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.

ENGINEER

M.S.E.E. in 1947. Single. Age 28. Two years' electronic work in the Navy. Two years' teaching. B.S.E.E. in 1942. Prefer development or research. Box 133W.

(Continued on page 60A)

Positions Wanted

(Continued from page 58A)

ADMINISTRATIVE ENGINEER

Relieve top level engineering personnel of technical-administrative duties; 5 years responsible experience National Bureau of Standards; project coordination and planning; new systems development; preparation of technical reports, engineering specifications; electronics procurement; technical representative for outside contacts. Age 27. Intelligent, initiative, ability to secure cooperation of others. Box 134W.

ENGINEER

B.E.E. New York University, 1944. Age 24. Single. Ex-communications officer. Desires work as executive's assistant or sales engineering in the radio-electronics field. Interesting work and opportunity for advancement primary importance. Box 135W.

JUNIOR ENGINEER

B.S.E.E. Carnegie Tech. in September 1947. Age 22. Single. 2 years' Navy electronics experience. Desires position in electronics research design or development. Box 139W.

JUNIOR ENGINEER

Graduate of RCA Institutes. Age 27. Married. Desires work in radio, electronics, television production or development anywhere in U. S. 3 years' radio work in Army. Box 140W.

SALES ENGINEER

Sales Engineer: Broadcast engineer, past three years as Chief Engineer of local station. Interested in entering sales field as an engineer with an established manufacturer of broadcast equipment. Family man. 28 years old; prefer southwest territory. Box 144W.

ELECTRICAL ENGINEER

Electrical Engineer. Age 27. Single. B.E.E. 1949. 5 years' experience in testing, development, drafting and maintenance of electrical equipment. Wants New York sales position with electrical or electronic equipment manufacturer. Sales training must be part of long range program. Write Box 145W.

ELECTRICAL ENGINEER

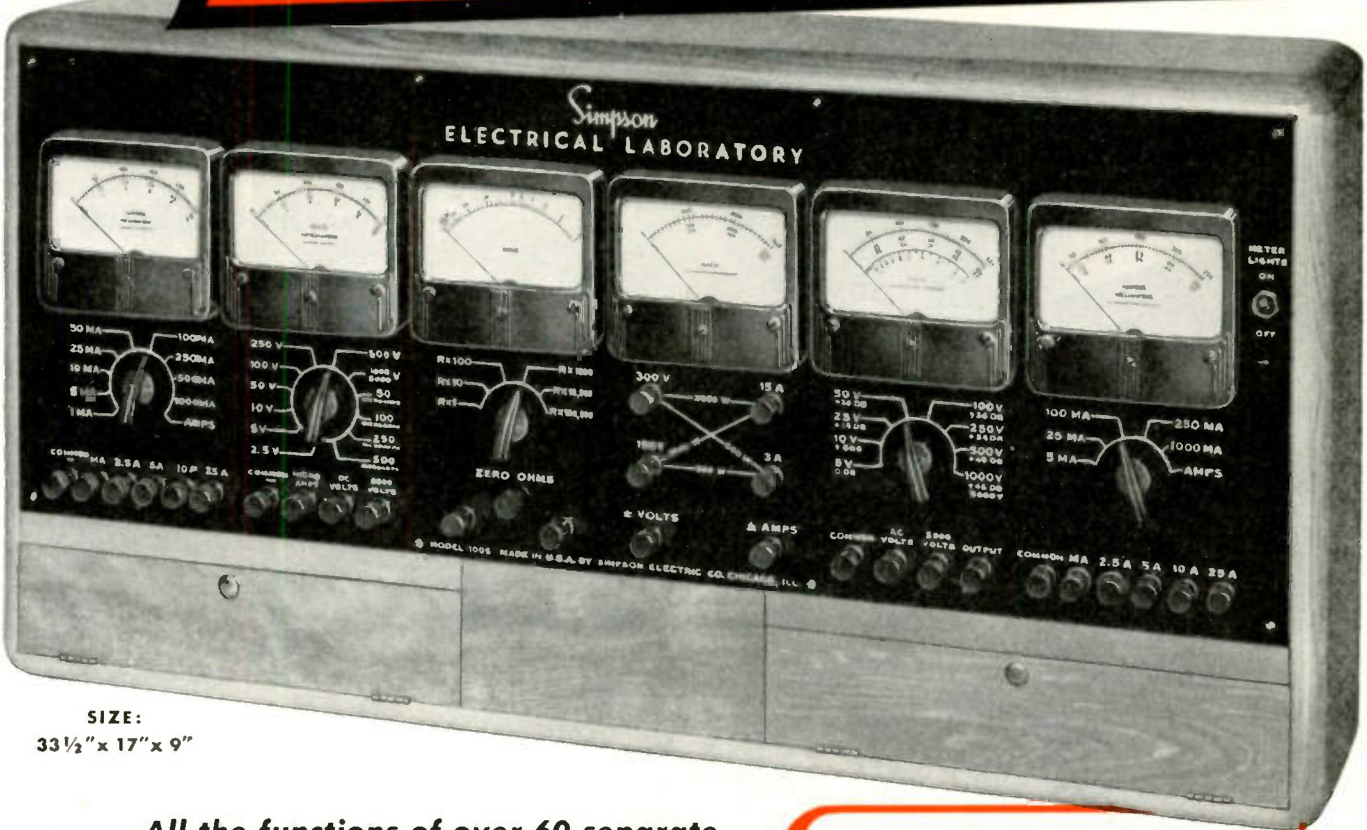
B.E.E. 1943 C.C.N.Y., M.S. 1948 Columbia University. Age 27. Married. Six years' experience on electronic research, product engineering and instructing including one year AAF service. Available June 1948. Prefer Omaha, Neb. or New York. Box 146W.

JUNIOR ENGINEER

School: RCA Institutes Inc., N.Y.C. 1948 Ex-Petty Officer Telegraphis (Royal Canadian Navy). Married. No children. Experience: Radio assembly. Operation, maintenance and installation of receivers, transmitters (LF, UHF), radar, RDF (LF, VHF), loran. Interested in engineering and sales. Box 147W.

(Continued on page 62A)

SIMPSON ELECTRICAL LABORATORY
MODEL 1005



SIZE:
33½" x 17" x 9"

All the functions of over 60 separate instruments combined in one unit!

Here is a complete test unit for use by radio, electronic, and electrical technicians in laboratories, shops, or service departments. It is adaptable to the testing of all electrical appliances, small motors, circuits, radio sets, etc. It consists of six individual 4½" rectangular instruments, indirectly illuminated, each with a complete set of ranges.

In addition to the wide variety of A.C. and D.C. voltage and current ranges, a multi-range ohmmeter and a single phase wattmeter have been incorporated. Also, to meet the need for extreme sensitivity required in testing circuits where only a small amount of current is available, an instrument is provided with a sensitivity of 50 microamperes, providing 20,000 ohms per volt on all D.C. voltage ranges. The Electrical Laboratory incorporates a rectifier type instrument for measuring A.C. voltage with a resistance of 1,000 ohms per volt on all ranges. This latter instrument also has in combination a complete coverage of DB ranges, from minus 10 to plus 54 for volume indications.

This beautiful instrument is Simpson-engineered and Simpson-built throughout for lifetime service.

Dealer's Net Price, complete with Leads and Break-in Plug, **$218.00**

SIMPSON ELECTRIC COMPANY

5200-5218 West Kinzie Street, Chicago 44, Illinois
In Canada, Bach-Simpson Ltd., London, Ont.

Ask your jobber

Simpson

INSTRUMENTS THAT STAY ACCURATE

RANGES OF MODEL 1005

Meter No. 1 (D.C. Milliammeter and Ammeter)
0-1 MA. D.C.
0-5 MA. D.C.
0-10 MA. D.C.
0-25 MA. D.C.
0-50 MA. D.C.
0-100 MA. D.C.
0-250 MA. D.C.
0-500 MA. D.C.
0-1000 MA. D.C.
0-2.5 Amps. D.C.
0-5 Amps. D.C.
0-10 Amps. D.C.
0-25 Amps. D.C.

Meter No. 2 (D.C. Microammeter and Voltmeter)
0-2.5 Volts D.C.
0-5 Volts D.C.
0-10 Volts D.C.
0-50 Volts D.C.
0-100 Volts D.C.
0-250 Volts D.C.
0-500 Volts D.C.
0-1000 Volts D.C.
0-5000 Volts D.C.
20,000 ohms per volt
0-50 Microamps
0-100 Microamps
0-250 Microamps
0-500 Microamps

Meter No. 3 (Ohmmeter)
0-500 Ohms (5 ohms center)
0-5000 Ohms (50 ohms center)
0-50,000 Ohms (500 ohms center)
0-500,000 Ohms (5,000 ohms center)
0-5 Megohms (50,000 ohms center)
0-50 Megohms (500,000 ohms center)

Meter No. 4 (Wattmeter)
0-300 Watts A.C.
0-600 Watts A.C.
0-1500 Watts A.C.
0-3000 Watts A.C.

Meter No. 5 (A.C. Voltmeter, Output and DB meter)
0-5 Volts A.C.
0-10 Volts A.C.
0-25 Volts A.C.
0-50 Volts A.C.
0-100 Volts A.C.
0-250 Volts A.C.
0-500 Volts A.C.
0-1000 Volts A.C.
0-5000 Volts A.C.
Rectifier type 1000 Ohms per volt
DB Ranges
−10 to +54
Output Ranges same as volts except 5000 Volt Range

Meter No. 6 (A.C. Milliammeter and Ammeter)
0-5 MA. A.C.
0-25 MA. A.C.
0-100 MA. A.C.
0-250 MA. A.C.
0-1000 MA. A.C.
0-2.5 Amps. A.C.
0-5 Amps. A.C.
0-10 Amps. A.C.
0-25 Amps. A.C.

Positions Wanted

(Continued from page 60A)

ELECTRICAL ENGINEER

B.E.E., R.P.I. (Communications). Age 26. Married. 1 child. Two years' Navy maintenance of radio and radar equipment. One year Civil Service testing VHF-UHF transreceivers. Current development of signal generators. Desires position as sales engineer. Box 148W.

ADMINISTRATIVE ENGINEER

Registered electrical engineer (N.Y.) FCC licensed. Eight years' experience in engineering, production, construction and administration; with power, communications and aircraft organizations. Harvard Business School graduate. Navy radar trained veteran. West coast preferred. Box 149W.

ELECTRICAL ENGINEER

B.S.E.E. Columbia, June 1948. Age 29. Married. 2½ years Naval radar; 1 year Naval Research Lab.; 1½ years' instructor in theory and shop practice; 5 years' experience in production planning and coordinator-metal and woodworking manufacturing. Tau Beta Pi. Desires position in design, development or production anywhere in the U. S. Box 150W.

ENGINEER

B.S.E.E. M.I.T., February 1943. Completing graduate work for E.E. degree (minor in Business Administration) Stanford, June 1948. Age 27. Married. One child. 3½ years' experience Naval shipboard and airborne electronics including assignments with N.R.L. and Airborne Coordinating Group (BuAero). Amateur Radio 10 years. Research or development. Box 151W.

ENGINEERING PHYSICIST

Army telephone central two years' test experience; graduate studies in higher math, physics, electronics; two credits remain to B.S. in engineering physics, Lehigh University, June, 1948; presently employed, seeking test research opportunity. Box 152W.

"THE hottest ham performance ever at this price . . ." That's the verdict of amateurs who have had a chance to try Hallicrafters new Model SX-43.

This new member of the Hallicrafters line offers continuous coverage from 540 kilocycles to 55 megacycles and has an additional band from 88 to 108 megacycles. AM reception is provided on all bands, except band 6, CW on the four lower bands and FM on frequencies above 44 megacycles. In the band of 44 to 55 Mc., wide band FM or narrow band AM just right for narrow band FM reception is provided.

One stage of high gain tuned RF and a type 7F8 dual triode converter assure an exceptionally good signal-to-noise ratio. Image ratio on the AM channel on band 5 (44 to 55 Mc.) is excellent as the receiver is used as a double superheterodyne. The new Hallicrafters dual IF transformers provide a 455 kilocycle IF channel for operating frequencies below 44 megacycles and a 10.7 megacycle IF channel for the VHF bands. Two IF stages are used on the four lower bands and a third stage is added above 44 megacycles. Switching of IF frequencies is automatic. The separate electrical bandspread dial is calibrated for the amateur 3.5, 7, 14, and 28 megacycle bands.

Every important feature for excellent communications receiver performance is included.

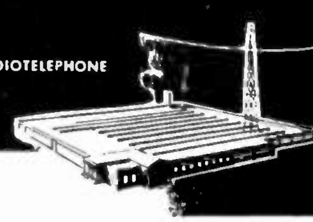
Model SX·43

$169.50

FEATURES FOUND IN NO OTHER RECEIVER AT THIS PRICE

- ALL ESSENTIAL AMATEUR FREQUENCIES FROM 540 kc to 108 MC
- AM - FM - CW RECEPTION
- IN BAND OF 44 TO 55 MC: WIDE BAND FM OR NARROW BAND AM . . . JUST RIGHT FOR NARROW BAND FM RECEPTION
- CRYSTAL FILTER AND EXPANDING IF CHANNEL PROVIDE 4 VARIATIONS OF SELECTIVITY ON LOWER BANDS
- SERIES TYPE NOISE LIMITER
- TEMPERATURE COMPENSATION FOR FREEDOM FROM DRIFT
- PERMEABILITY ADJUSTED "MICROSET" INDUCTANCES IN THE RF CIRCUITS
- SEPARATE RF AND AF GAIN CONTROLS
- EXCEPTIONALLY GOOD SIGNAL-TO-NOISE RATIO
- SEPARATE ELECTRICAL BANDSPREAD CALIBRATED FOR THE AMATEUR 3.5, 7, 14 AND 28 Mc BANDS

BUILDERS OF *Skyfone* AVIATION RADIOTELEPHONE



hallicrafters RADIO

THE HALLICRAFTERS CO., MANUFACTURERS OF RADIO AND ELECTRONIC EQUIPMENT, CHICAGO 16, U. S. A.

Sole Hallicrafters Representatives in Canada: Rogers Majestic Limited, Toronto-Montreal

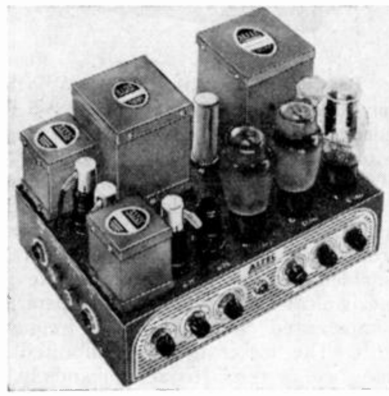
News—New Products

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

(Continued from page 48A)

Portable Amplifier

A new portable public-address amplifier engineered to provide quality consonant with the highest-priced microphones and loudspeaker systems has been brought out by **Altec Lansing Corporation,** 250 W. 57 St., New York 19, N. Y. The new amplifier is catalogued as Model A-324.



The A-324 is rated at 15 watts with a guaranteed full-power output within 1 db from 35 to 12,000 cycles. Its over-all frequency response is flat within 1 db from 20 to 20,000 cycles.

Several unique features are claimed for the new amplifier. Four inputs are provided: two of the inputs provide 95 db gain for low-impedance microphones with individual volume controls on each input for mixing purposes.

The transformer in the high-gain low-impedance microphone input circuits has 90 db shielding to guard against hum and noise pickup.

Two other high-impedance inputs provide 72 db gain for radio or phonograph pickup or high-impedance microphones; they are coupled to a dual-type volume control which allows fading smoothly from one input to the other.

Another feature is a continuously variable bass control which, at the low end, is coupled to a switch to cut in special equalization to correct the boomy reproduction. A continuously variable treble attenuator is also provided.

New Enterprise

A new company, **International Rectifier, Corp.,** announces the opening of a plant at 6809 Victoria Ave., Los Angeles, Calif. The new firm is equipped for research and manufacture in the field of colorimetric equipment, photoelectric cells, and selenium rectifiers.

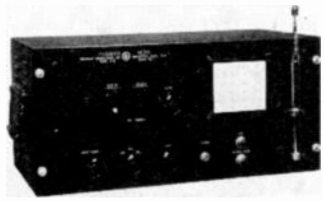
(Continued on page 66A)

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 65A)

Model S-5 Frequency Meter



Browning Laboratories, Inc., Winchester, Mass., announces the Model S-5 Frequency Meter for the accurate checking of transmitters operating between 30 and 500 Mc. A crystal standard in a temperature-controlled oven is employed with a long-time accuracy of 0.001%. The electron-coupled interpolation oscillator is assembled on an aluminum plate for mechanical stability and is temperature-compensated for minimum frequency drift. The meter is hand-calibrated for one, two, or three frequency bands in the range from 30 to 500 Mc. with an over-all accuracy of 0.0025%. High mixer sensitivity permits use of the instrument without the need of direct connection to the transmitter. A panel-mounted telescoping antenna is employed as a pickup means. The heavy steel cabinet which is provided may be removed and the instrument used in a relay rack with associated equipment. It measures 8¾ inches high, by 19 inches wide, by 9 inches deep, and weighs 35 pounds.

Recent Catalog

• • • On crystals, Bulletin 36 containing a complete listing of all types of crystals currently manufactured by **Bliley Electric Co.,** Erie, Pa., for all types of commercial applications. This catalog does not contain crystals designed specifically for amateur application, as these units are described in Bulletin 35.

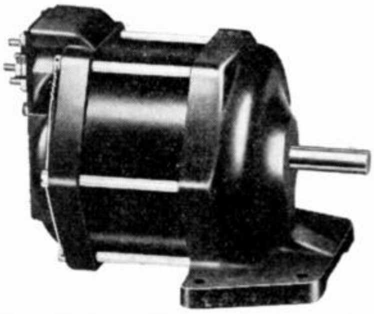
• • • On *Audigage*, a portable instrument designed to measure the thickness of a wide variety of materials, by **Branson Instruments Inc.,** Joe's Hill Rd., Danbury, Conn.

• • • On capacitors for electronic and service-replacement applications, a 24-page illustrated catalog, by **Cornell-Dubilier Electric Corp.,** So. Plainfield, N. J. Ask for Catalog No. 200.

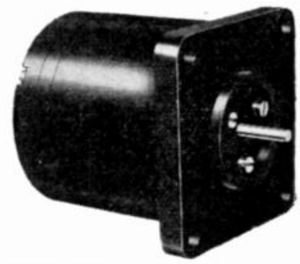
(Continued on page 68A)

For precise, positive linkage between instrumentation and control

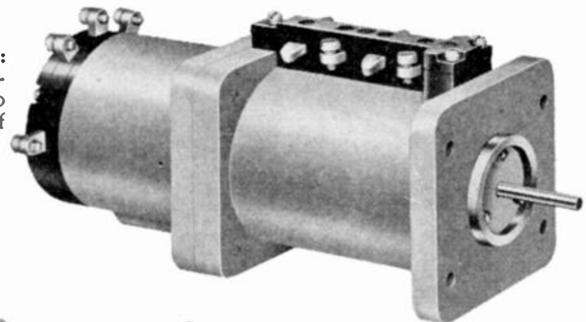
INDUCTION GENERATOR: when fed from AC source produces voltage proportional to speed of rotation. Used in circuits as velocity control component.



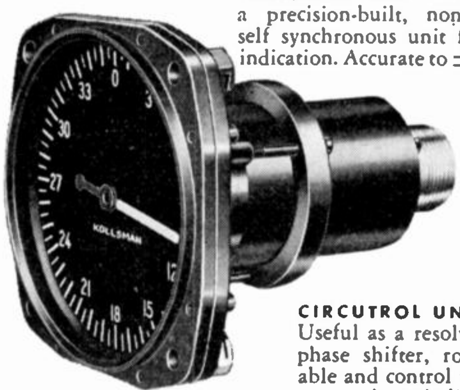
PERMANENT MAGNET GENERATOR: designed as AC potential source. Produces sinusoidal wave form with harmonic content under 2%.



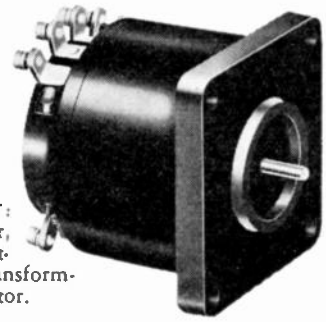
INDUCTION GENERATOR: type designed particularly for use where low residual voltage is required.



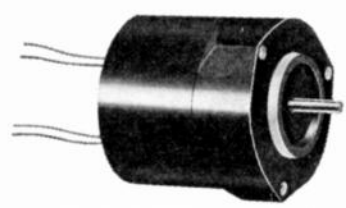
MOTOR DRIVEN INDUCTION GENERATOR: powered by 2-phase, low-inertia induction motor. Used as fast reversing servo motor where maximum stall torques of less than 7 oz. in. are required.



TELETORQUE UNIT — below left: a precision-built, non-motoring, self synchronous unit for remote indication. Accurate to ± 1 degree.



CIRCUTROL UNIT: Useful as a resolver, phase shifter, rotatable and control transformer or phase indicator.



INDUCTION MOTOR: Low inertia, two-phase squirrel cage unit for use as precision servo motor.

KOLLSMAN OFFERS A LINE OF SPECIAL PURPOSE AC UNITS

To meet the varying needs of the electronics engineer in linking instrumentation up to control, Kollsman offers a group of units with sufficiently varied functions to solve a wide range of control problems. In nearly every case, units are available for operation at various voltages and frequencies to fit widely diversified electronic control and remote indication applications. These Kollsman units are the outgrowth of long development in aircraft instrumentation and control and — more recently — Kollsman's considerable work in this field for naval and military applications. They are light in weight, compact, and highly precise, so that engineers working with exact quantities will find them reliable to a high degree. Complete data on any or all of these units may be had upon request. Kollsman Instrument Division, Square D Company, 80-08 45th Avenue, Elmhurst, N. Y.

KOLLSMAN AIRCRAFT INSTRUMENTS

PRODUCT OF

SQUARE D COMPANY

ELMHURST, NEW YORK · GLENDALE, CALIFORNIA

News—New Products

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

(Continued from page 66A)

Recent Catalogs

• • •On the reduction of precipitation static in aircraft radio, an 8-page booklet published by **Dayton Aircraft Products, Inc.,** 342 Xenia Ave., Dayton, Ohio, manufacturers of shielded antenna fittings for commercial aircraft and for the U. S. Air Force.

• • •On synchronous timing motors and timing devices, a 16-page illustrated catalog by the **Haydon Manufacturing Co., Inc.** The catalog is divided into sections for each of nine different motor series and for the various types of timing devices, such as repeat-cycle and reset timers, time-delay relays, interval timers, etc. Copies may be obtained by writing E. B. Hamlin, Haydon Mfg. Co., Inc., East Elm St., Torrington, Conn.

• • •On a new signal generator, a 4-page illustrated technical bulletin issued by **Premier Crystal Laboratories, Inc.,** 53–63 Park Row, New York 7, N. Y., giving details and specifications of the new Model 117 Crystal-Controlled High-Frequency Mini-Signal Generator.

(Continued on page 70A)

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 68A)

Recent Catalogs

• • • A new edition of the RCA Receiving Tube Manual, designated as Technical Series RC-15, is now being distributed, according to an announcement by **Radio Corporation of America,** RCA Victor Division, Tube Department, Harrison, N. J. This is the first edition of the manual since 1939. In addition to greatly expanded coverage in its regular sections, the RC-15 presents many new features.

• • • On type "B" high-voltage resistors, a new catalog sheet, which presents engineering data, issued by **Resistance Products Co.,** Division of Electronic Manufacturing Co., 714 Race St., Harrisburg, Pa.

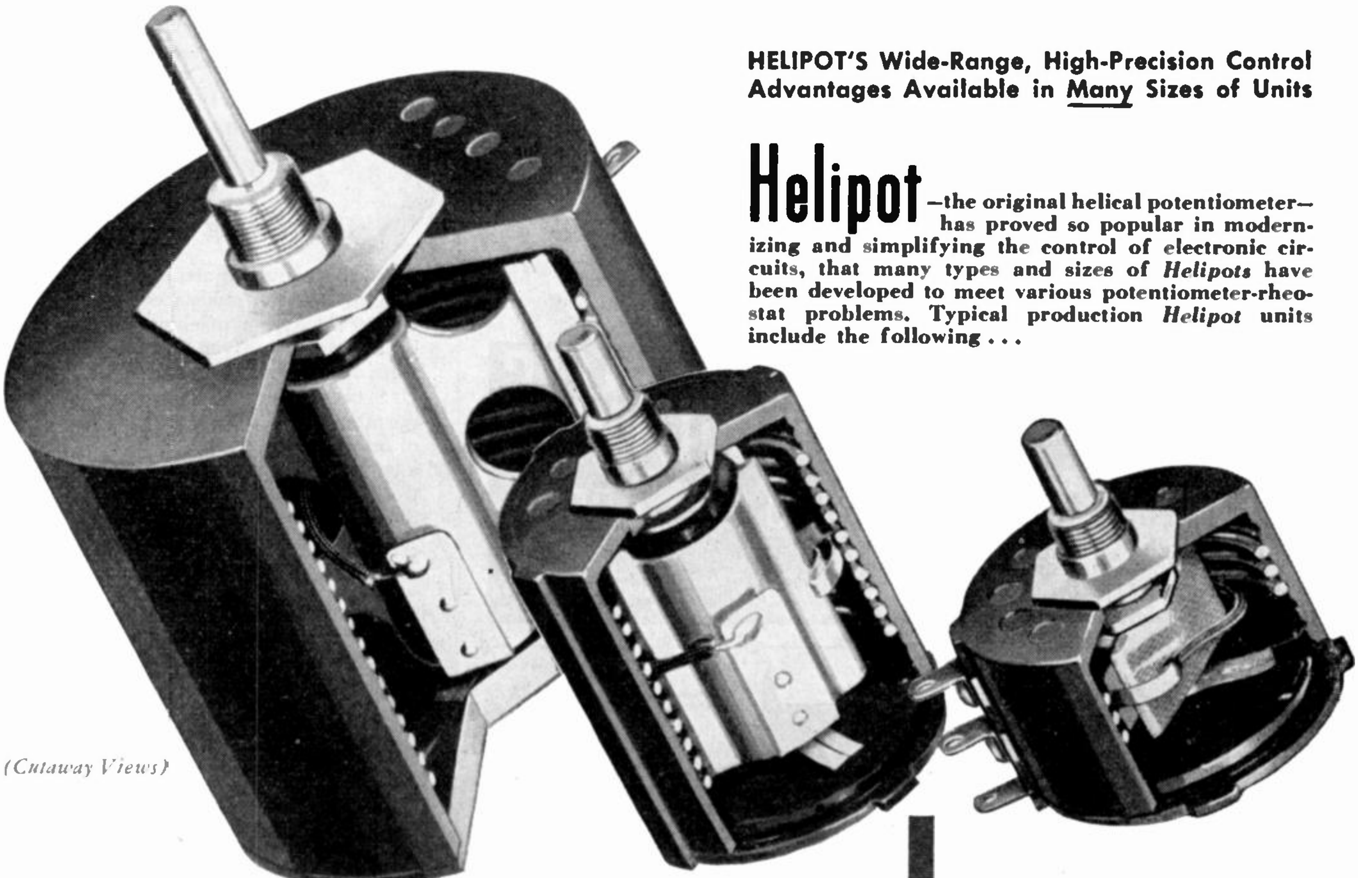
• • • On Flexible Waveguide Assemblies, an illustrated bulletin issued by **Technicraft Laboratories, Inc.,** 237 East Aurora St., Waterbury, Conn. Technicraft manufactures a complete line of rectangular flexible waveguide assemblies for use in the microwave spectrum. Three basic types are described in the bulletin: Type L, Interlocked Construction; Type S, Seamless Construction; and Type V, Vertebra Construction. Write to the manufacturer for Bulletin F-1.

(Continued on page 72A)

There's a Beckman Helipot

(Trade Mark of the HELIcal POTentiometer)

to simplify YOUR Potentiometer—Rheostat Problems!



(Cutaway Views)

HELIPOT'S Wide-Range, High-Precision Control Advantages Available in Many Sizes of Units

Helipot—the original helical potentiometer—has proved so popular in modernizing and simplifying the control of electronic circuits, that many types and sizes of *Helipots* have been developed to meet various potentiometer-rheostat problems. Typical production *Helipot* units include the following . . .

MODEL B—Case diameter—3.3"; Number of turns—15; Slide wire length—140½"; Rotation—5400°; Power rating—10 watts; Resistance ratings—50 to 200,000 ohms.

MODEL A—Case diameter—1.8"; Number of turns—10; Slide wire length—46½"; Rotation—3600°; Power rating—5 watts; Resistance ratings—10 to 50,000 ohms.

MODEL C—Case diameter—1.8"; Number of turns—3; Slide wire length—13.5"; Rotation—1080°; Power rating—3 watts; Resistance ratings—5 to 15,000 ohms.

WIDE CHOICE OF DESIGN FEATURES

Not only are *Helipots* available in a wide range of sizes and ratings, but also can be supplied with various design features to meet individual requirements . . .

- Available with special length shafts, flatted shafts, screw-driver slots, etc.
- Can be supplied with shaft extensions at each end to permit coupling to indicating instruments or other devices.
- May be provided in ganged assemblies of two or three units, all operating from a common shaft.
- Available with linearity tolerances of 0.1%—and even less.
- Models A & B can be modified to include additional taps at virtually any point on windings.

. . . and many other special features.

Investigate the many important advantages to be gained by using the *Helipot* in your electronic control applications. Write outlining your problem!

SPECIAL MODELS

In addition to the above standard *Helipot* units, special models in production include . . .

MODEL D—Similar to Model B, above, but longer and with greater length of slide wire. Case diameter—3.3"; Number of turns—25; Slide wire length—234"; Rotation—9000°; Power rating—15 watts; Resistance ratings—100 to 300,000 ohms.

MODEL E—Similar to Model B, but longer and with greater length of slide wire than Model D. Case diameter—3.3"; Number of turns—40; Slide wire length—373"; Rotation—14,400°; Power rating—20 watts; Resistance ratings—150 to 500,000 ohms.

Send for HELIPOT Literature!

THE Helipot CORPORATION
1011 Mission Street
South Pasadena 6, California

See the HELIPOT at the IRE Show, Booth 243. Also the new DUODIAL—the revolutionary multi-turn indicating knob dial!

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 70A)

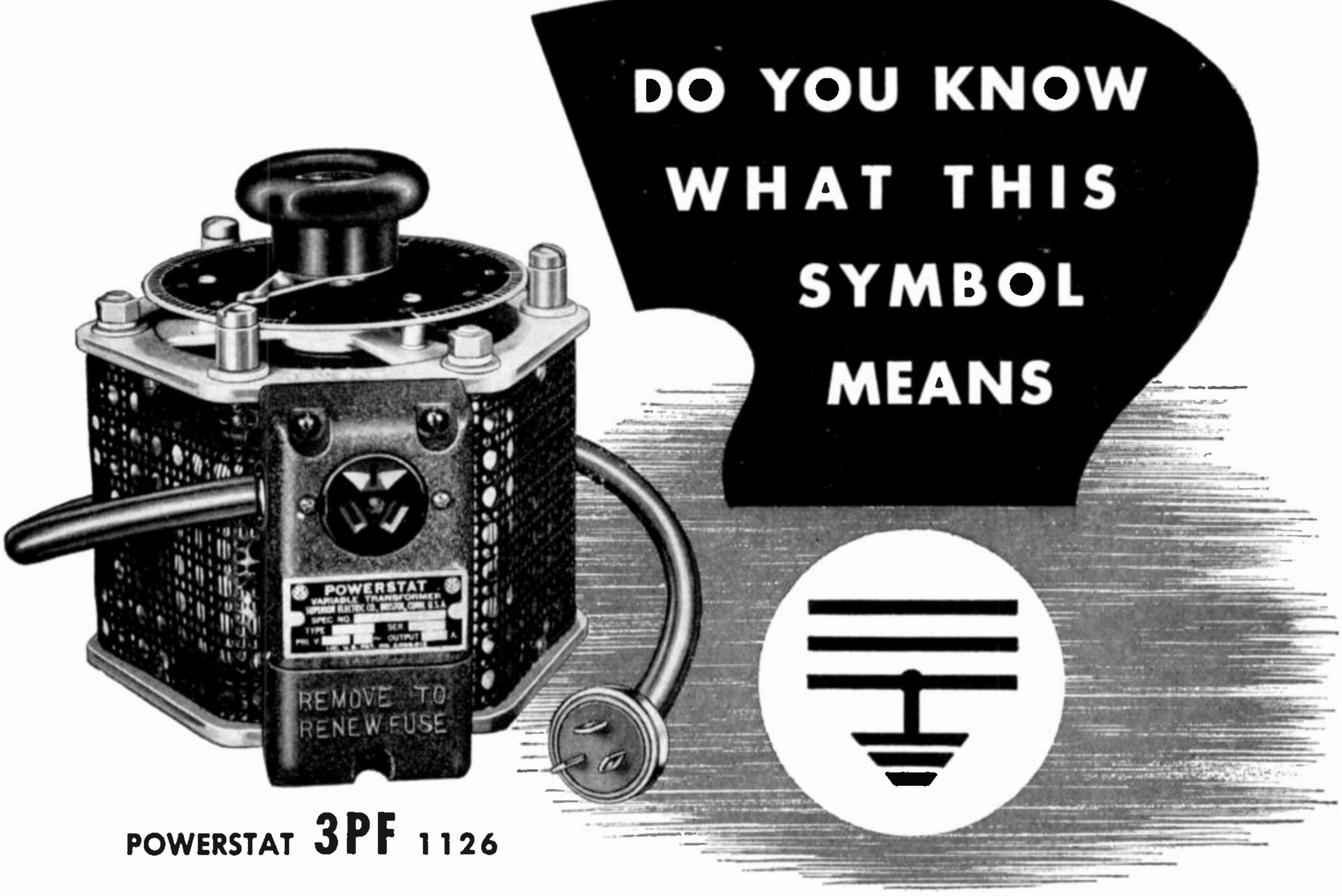
Recent Catalogs

• • •On resistors, rheostats, and relays for amateur use, a new catalog, No. D-30, published by **Ward Leonard Electric Co.,** Mount Vernon, N. Y. A copy of the new catalog may be obtained by writing to Radio and Electronic Distributor Division, Ward Leonard Electric Company, 53 W. Jackson Blvd., Chicago 4, Ill.

Interesting Abstract

• • •Several new illustrated bulletins describing various models of the *Original-Odhner* calculating machines have recently been issued by **Ivan Sorvall, Inc.,** 210 Fifth Ave., New York 10, N. Y., the sole U. S. distributor for these machines. These portable, hand-operated calculators will be of interest to the engineer as they feature a back-transfer device to speed up calculations, when series of multiplications and divisions are involved.

(Continued on page 80A)



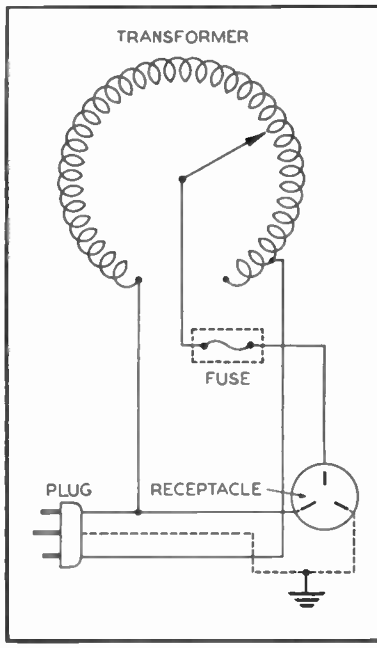
POWERSTAT 3PF 1126

It stands for a three-wire system, with the third wire grounded. It means added personal safety and insurance against shock from "hot" circuits.

A third wire grounded in a three-wire, single phase system is becoming a requirement in more and more communities . . . and POWERSTAT variable transformers are prepared for this transition. Standard models are available . . . wired for a three-wire, single phase system with one wire grounded.

Safety and versatility—two important features of The Superior Electric Company's dependable voltage control equipment has resulted in wide acceptance of these quality units for use in laboratory and industry.

POWERSTAT variable transformers are easily adapted to fit individual specifications. Let the experience of The Superior Electric Company's voltage control engineers assist in solving your specific problem. Request Bulletin 547 for complete voltage control engineering data.

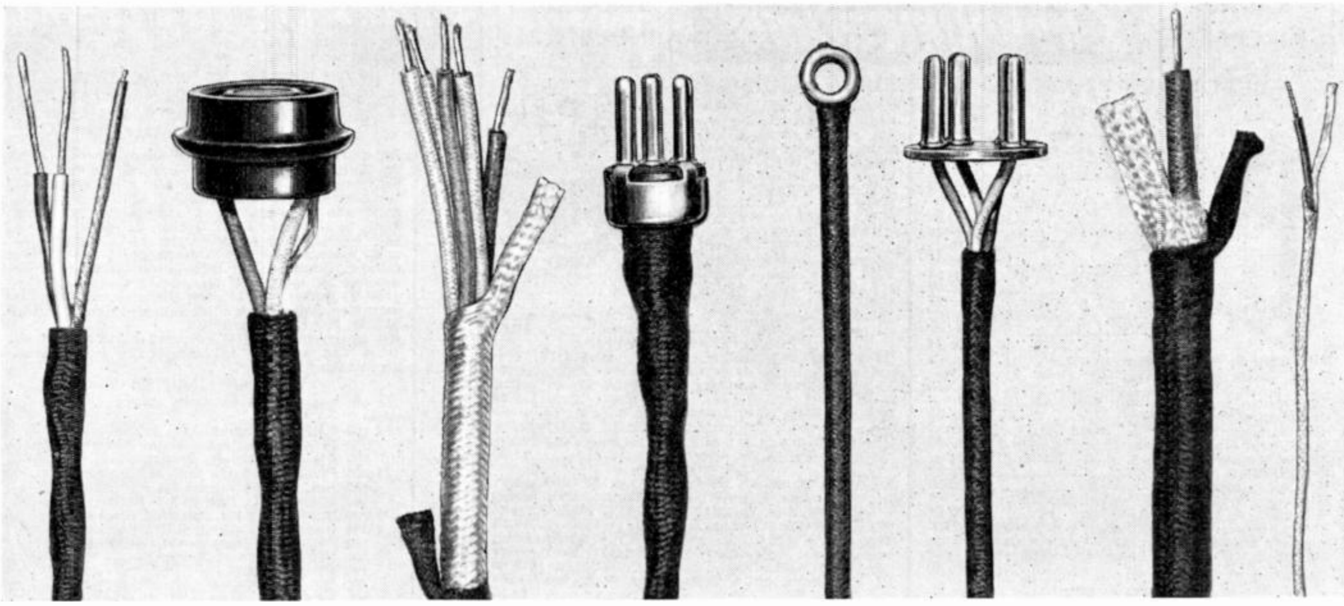


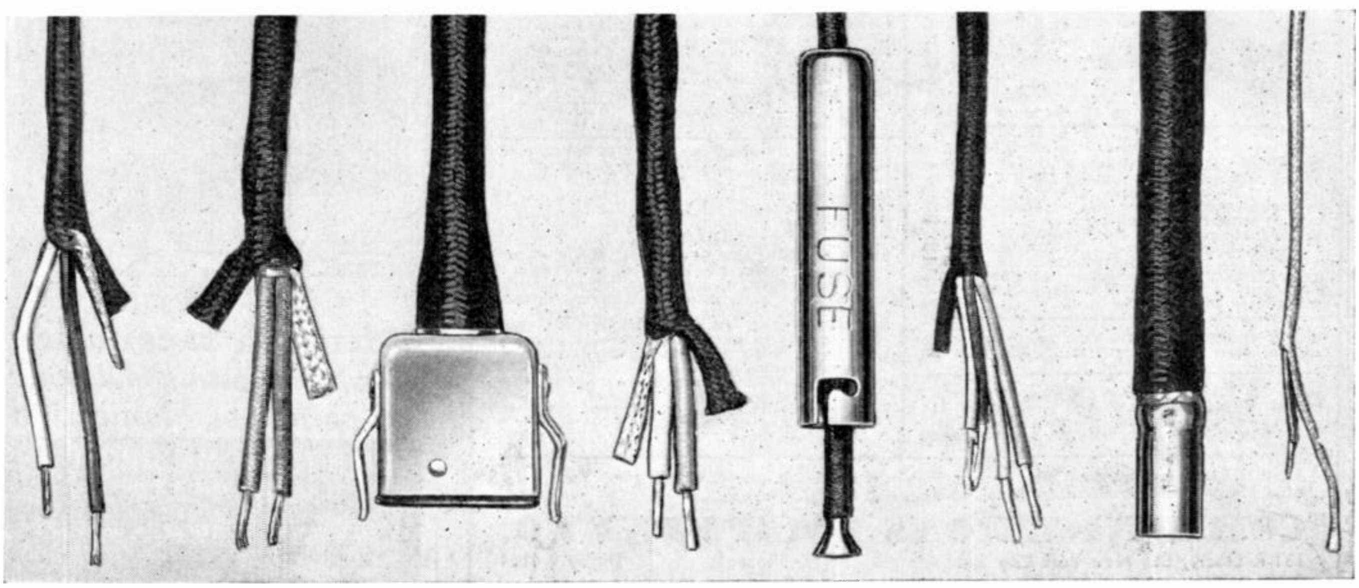
The schematic drawing shows a typical single phase, three-wire POWERSTAT variable transformer with the third wire grounded.

Write The Superior Electric Co., 803 Meadow St., Bristol, Conn.

THE SUPERIOR ELECTRIC CO.
BRISTOL, CONNECTICUT

Powerstat Variable Transformers • Voltbox A C Power Supply • Stabiline Voltage Regulators.





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TYPE	PRICE	TYPE	PRICE	TYPE	PRICE	TYPE	PRICE	TYPE	PRICE	TYPE	PRICE	TYPE	PRICE
1A3	.98	5U4	.44	6Q5G	.98	12X3	.98	85	.89	829A/B	2.95	8016	1.65
1A5	.49	5V4G	.98	6Q7	.89	14A7	1.10	100TS	3.00	830B	5.25	8020	5.95
1A7GT	1.10	5W4	.98	6R7	.98	14B7	1.10	117L7	1.89	832A	2.25	8025	2.95
1B24	2.49	5Y3	.60	6SA7	.90	14H7	1.25	117Z3	.89	833A	34.50	9001	.89
1B38	4.50	5Y4G	.59	6SC7	.85	14J7	1.25	117Z6GT	1.10	836	1.15	9002	.49
1G4	.98	5Z3	.89	6SF5	.79	14R7	1.10	121A	2.65	837	2.50	9003	.49
1G5	.44	5Z4	.89	6SG7	.79	15E	1.50	205B	4.50	838	3.75	9004	.49
1G6	.98	6A6	.75	6SH7	.39	23D4	.49	211	.98	841	.69	9005	.98
1H4G	.98	6AC7	.99	6SJ7GT	.69	23D6	.98	215A	3.00	845	3.75	9006	.49
1L4	.89	6AG5	.89	6SK7	.79	24G	.69	217C	7.50	860	3.00	EE50	.79
1R4/1294	1.29	6AG7	.99	6SL7	.89	25A6GT	.75	250TH	12.95	861	50.00	HF100	6.95
1T4	.58	6AK5	.69	6SN7GT	.69	25L6GT	.75	304TL	2.49	866A	.75	HY75	1.25
1H5	.99	6AL5	.99	6SQ7	.89	25Z5	.75	307A	6.25	872A	1.95	HY615	1.25
1N5GT	1.10	6AQ5	.98	6SR7	.89	25Z6	.98	316	.89	874	1.95	OZ4	1.25
1LN5	1.92	6AT6	.75	6SS7	.75	28D7	.75	371A	1.39	884	.75	RK60	.79
1R5	1.10	6AU6	.89	6U5	.98	30	.78	371B	3.00	923	.49	RK72	3.50
1S5	1.10	6B4	1.29	6V6GT	.99	32L7	1.50	394A	4.50	954	.49	T20	1.95
2A3	1.39	6B6G	.89	6Y6G	.89	34	.98	417A	19.95	955	.49	TZ40	2.95
2C22	.69	6B8	.99	6X4	.98	35L6GT	.75	446A	1.25	956	.75	V70D	6.90
2C26A	.75	6C4	.64	6X5	.89	35Y4	1.10	450TH	12.95	957	.49	VR78	.75
2C34	.98	6C5	.51	7AE7	.75	35W4	.69	703A	7.50	958A	.49	VR90	.75
2C40	2.60	6C6	.75	7B7	.69	35Z3	.99	705A	1.85	959	.49	VR105	.75
2C44	1.75	6C21	12.95	7C4	1.50	35Z5	.69	713A	1.65	991	.50	VR150	.69
2D21	.75	6D4	.89	7C5	.89	36	1.10	715B	4.95	1005	.39	Z225	1.95
2E22	1.50	6D6	.75	7F7	1.25	37	.69	717A	.69	1006	.39	902	2.95
2E25	3.95	6F4	1.35	7L7GT	1.39	38	.89	721A	3.95	1613	.95	2AP1	1.95
2E30	2.25	6F5	.51	10Y	.69	39/44	.59	723A/B	5.50	1614	1.75	3AP1	1.95
2J32	20.00	6F6	.79	12A6	.89	41	.69	725A	12.50	1616	1.39	3BP1	1.95
2J33	20.00	6F6G	.80	12AH7	1.10	45	.64	800	2.25	1619	.98	3CP1	1.89
2J851	4.95	6F7	.98	12AT6	1.10	46	.65	801A	1.10	1622	1.75	5AP1	2.49
2X2	.69	6F8	1.10	12BA6	.89	47	.90	802/RK25	1.49	1624	.98	5BP1	1.49
3A4	.49	6G6	1.10	12BE6	.89	50B5	.89	803	8.95	1625	.49	5BP4	4.95
3B7	.98	6H6	.49	12C8	.89	50L6GT	.75	804	6.75	1626	.49	5CP1	3.95
3B22	4.95	6J4	1.50	12H6	.44	70L7	.89	805	3.75	1629	.59	5FP7	4.50
3B24	.98	6J5	.49	12J5	.69	71A	.69	807	1.25	1631	1.49	7BP7	2.95
3D6/1299	.89	6J6	.49	12K8	1.25	75	.69	808	2.95	1641/RK60	.79	7DP4	14.95
3E29	2.95	6J7	.89	12SA7GT	.99	75T	2.39	809	1.50	1851	1.25	7EP4	17.95
3Q4	1.10	6K6	.49	12SG7	.89	76	.75	810	5.95	2050	.90	7GP4	19.40
3Q5GT	.58	6K7	.59	12SH7	.89	77	.75	811	1.95	2051	.49	9AP4	50.00
3S4	.43	6K8	1.25	12SJ7	.79	78	.75	812	3.15	5514	3.95	10BP4	29.95
4C35	7.95	6L6	1.25	12SK7	.69	79	1.10	812H	6.90	7193	.39	10FP4	42.20
4E27/257B	4.95	6L6G	1.20	12SL7	1.10	80	.53	813	5.95	8001	4.95	12JP4	49.50
5R4GY	1.15	6L7	.98	12SN7GT	.79	82	.98	814	4.39	8005	3.25	15AP4	110.00
5T4	1.25	6N7	.89	12SQ7GT	.99	83V	.89	815	2.25	8011	2.95	20AP4	270.00
		6Q5	.98	12SR7	.79	84	.75	826	1.75	8012	4.95		

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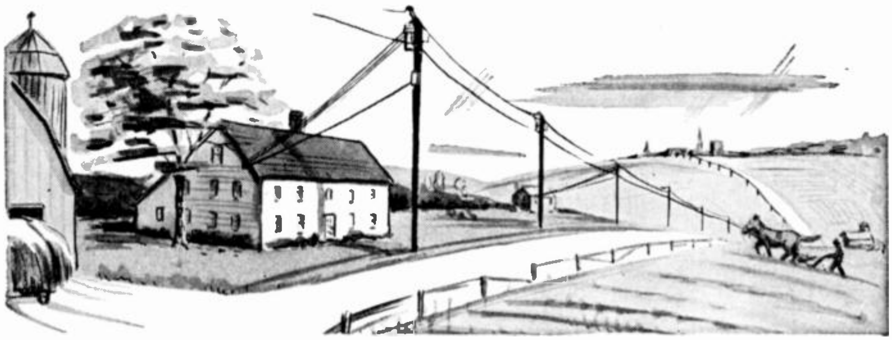
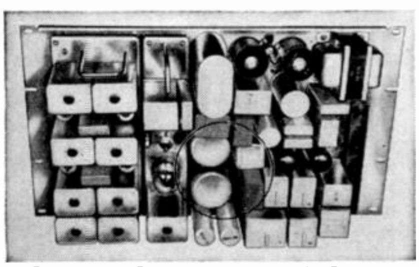
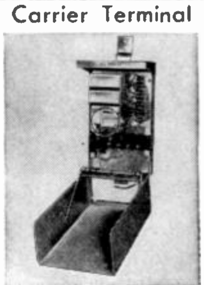
Sigma RELAYS ARE USED IN Western Electric POWER LINE CARRIER TELEPHONE SYSTEM

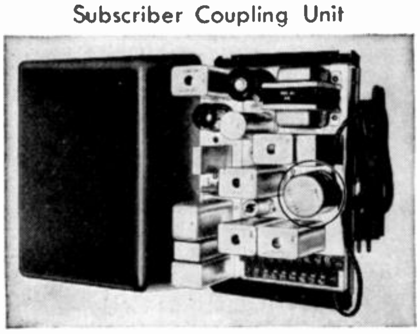




Common Carrier Terminal Chassis



Subscriber Coupling Unit



Subscriber Terminal Equipment

The Western Electric M1 Power Line Carrier Telephone System permits telephone service in thousands of farm houses having electric power service but no telephone wire line connections. It will help raise living standards in many rural areas.

Sigma Relays are used for three functions in this equipment, two of which are unusually exacting. By careful cooperative study of each application Sigma was able to work out solutions using highly refined but none the less conventional sensitive relays of standard Sigma design — available at comparatively low cost.

From vending machines to V-Bombs specialized relay design plus facility at solving problems involving circuit, relay and function enable Sigma to render valuable service.

SIGMA RELAY TYPES

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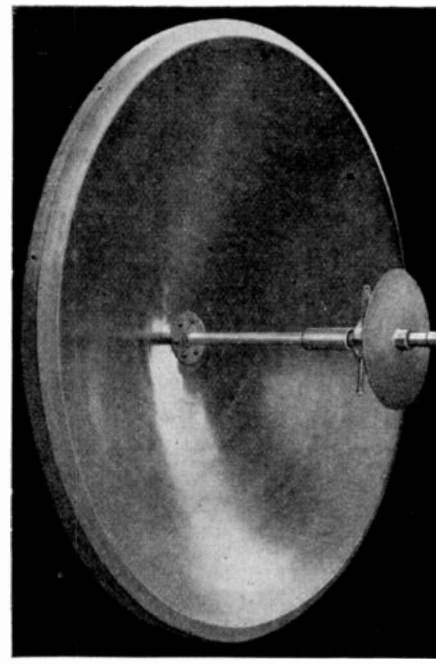
SINGLE OR MULTIPLE CIRCUIT
From 68¢ to $25.00 each!

SIGMA

Sigma Instruments, INC.

Sensitive **RELAYS**

94 CEYLON ST., BOSTON 21, MASS.



PARABOLIC ANTENNAS

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- **FM and AM Studio-to-Transmitter Link**
- **Television and Facsimile Relay Work**
- **Multi-channel Point-to-Point Relay**
- **Research and Development Laboratories**

The Workshop can supply parabolic antennas in a wide range of types, sizes and focal lengths, plus a complete production and engineering service on this type of antenna.

Workshop test equipment and measurements for the determination of antenna characteristics is outstanding in the industry. These facilities, coupled with the wartime experience of its engineers on high frequency antennas, assure exceptional performance.

PARABOLAS — Precision-formed aluminum reflectors. Can be supplied separately, if desired.

MOUNTINGS — Various types of aluminum reinforced mountings can be supplied with all antennas.

R. F. COMPONENTS — Precisions machined and heavily silver plated. Critical elements protected by low-loss plastic radome.

PATTERN AND IMPEDANCE DATA — A series of elaborate measurements of both pattern and impedance are made to adjust the settings for optimum performance. Pattern and impedance data is supplied with each antenna.

POLARIZATION — Either vertical or horizontal polarization can be obtained easily by a simple adjustment at the rear of the reflector.

SPECIAL ANTENNAS — Parabolas can be perforated to eliminate wind resistance or sectioned to produce a specified antenna pattern.

OTHER ANTENNAS — FM and television receiving antennas. A complete line of amateur antenna equipment.

Prices on Request

The Workshop invites your inquiry on any type of high frequency antenna problem — no obligation. Write, or phone Boston, BIGelow 3330.

The WORKSHOP ASSOCIATES, Inc.
66 NEEDHAM STREET
Newton Highlands, Massachusetts

NEWS—NEW PRODUCTS

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

(Continued from page 72A)

Voltage Regulator Tube

The Tube Department, **Radio Corp. of America,** Harrison, N. J., has announced production of the OB2 voltage regulator tube, which, like the OA2, is a miniature cold-cathode glow-discharge tube.

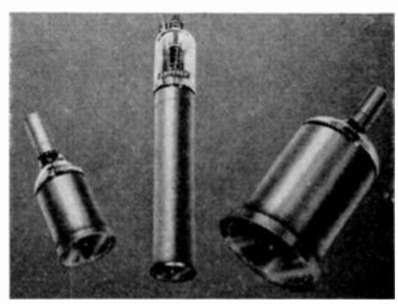
The OB2 regulates at approximately 108 volts over a current range of 5 to 30 ma., whereas the OA2 regulates at approximately 150 volts. These two types permit equipment designers to provide regulated B and C voltages in compact equipment where space heretofore precluded use of the larger voltage-regulator tubes.

New Folded Dipole

A folded dipole designed for use as a receiving or transmitting antenna in the 85- to 150-Mc. range is being introduced by the Communications Equipment Division of **Heintz and Kaufman, Ltd.,** 50 Drumm St., San Francisco, Calif.

This dipole can be accurately tuned to any frequency within this range; hence it is adaptable for f.m. reception, aviation service, the amateur two-meter band, and mobile services in the vicinity of 150 Mc. For 85 Mc. operation the dipole is extended to 65 inches; at 148 Mc., its over-all length is reduced to 37 inches.

Geiger-Mueller Counter Tubes



The illustration shows the redesigned beta, gamma and X-ray counter tubes now being produced by **Amperex Electronic Corp.,** 25 Washington St., Brooklyn 1, N. Y.

The manufacturer has long supplied these tubes as laboratory devices to supply the special needs of research physicists. Now, they are redesigned for standardized production.

Brochures describing the many counter tubes now in regular production are available upon request directed to the manufacturer.

(Continued on page 82A)

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 80A)

Interesting Abstracts

• • • Expansion of **Industrial Television, Inc.,** formerly of 36 Franklin Ave., Nutley, N. J., was indicated by recent announcements that the company has signed a long-term lease for a two-story building at 359 Lexington Ave., Clifton, N. J. and that a sales office is being opened at 354 Fourth Ave., New York, N. Y. Industrial Television specializes in the manufacture of large-screen, direct-view Teleceivers.

• • • **Leeds & Northrup Company,** Stenton Ave., Philadelphia, Pa., manufacturers of electrical measuring instruments, automatic controls, and other equipment, have announced the purchase of a two-story-and-basement building at 34 East Logan St., one block from its main plant.

• • • **The Minnesota Electronics Corp.,** Oppenheim Bldg., Sixth & Minnesota Sts., St. Paul 1, Minn., makers of the *Goodell* line of radio-phonographs, announce a new Dynamic Noise Suppressor Amplifier under a license agreement with Hermon Hosmer Scott, inventor of the dynamic noise suppressor.

• • • A comprehensive expansion program for stepping up the production of television picture tubes at the Lancaster, Pa., plant of the **Radio Corporation of America,** was announced by L. W. Teegarden, vice president in charge of the RCA Tube Department. At present the plant is turning out tubes for both the transmission and reception of television as well as power and other special types of tubes.

• • • The **Wire Recording Corporation of America,** 1331 Halsey St., Brooklyn, N. Y., recently incorporated, has taken over the assets and manufacturing facilities of the St. George Recording Equipment Co. of New York City. J. J. Sullivan, president of the newly formed corporation, announces completion of plans to manufacture the *Wireway* Wire Recorder.

(Continued on page 85A)

NOTICE

Information for our News and New Products section is warmly welcomed. News releases should be addressed to **Mrs. Harriet P. Watkins, I.R.E. Industry Research Division, Room 707, 303 West 42nd St., New York 18, N. Y.** Photographs, and electrotypes if not over 2″ wide, are helpful. Stories should pertain to products of interest specifically to radio engineers.

Attention

Associate Members!

Many Associate Members can qualify for higher membership grades and should certainly do so. Members are urged to keep membership grades up in pace with their present development.

An Associate over 24 years of age who is occupied as a radio engineer or scientist, and is in this active practice three years may qualify for Member Grade.

An Associate who has taught college radio or allied subjects for three years may qualify.

Some may possibly qualify for Senior Grade. But transfers can be made only upon your application. For fuller details request transfer application-form in writing or by using the coupon below.

I R E

Coupon

Institute of Radio Engineers
1 East 79th St.,
New York 21, N.Y. 3-48

Please send me the Transfer Application Membership-Form.

Name ....

Address ....

Place ....

State ....

Present Grade ....

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 82A)

Midget Wire-Wound Resistors

Handy, inexpensive, ceramic-cased midget wire-wound resistors for tight spots and for facilitating point-to-point wiring, known as Greenohm Juniors, are announced by **Clarostat Mfg. Co., Inc.**, 130 Clinton St., Brooklyn, N. Y. These resistors take the place of more cumbersome and costly bracket-mounted units, especially where space is at a premium.

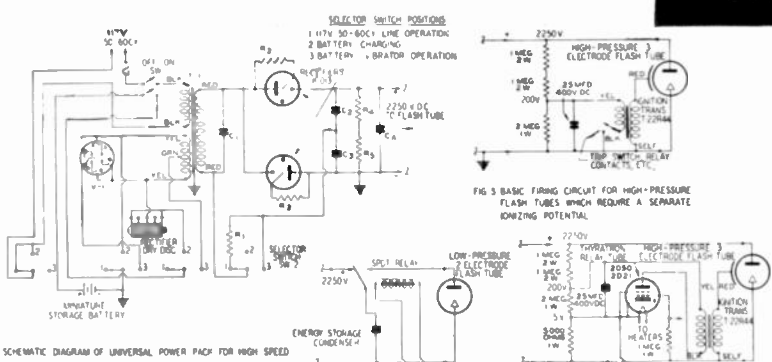
This "junior" version of the well-known Greenohm power resistors features a wire winding on a fiber-glass core, with axial bare pigtail leads clinched to the ends, placed in a steatite tube and thoroughly filled and sealed with cold-setting inorganic cement. Since there is no organic material in this resistor, it will not blister, crack, or change shape.

INDEX AND DISPLAY ADVERTISERS

DISPLAY ADVERTISERS

THORDARSON

DEPENDABILITY AND VERSATILITY FOR INDUSTRY



SCHEMATIC DIAGRAM OF UNIVERSAL POWER PACK FOR HIGH SPEED PHOTO-FLASH TUBES

FIG 4 FIRING CIRCUIT FOR LOW-PRESSURE FLASH TUBES

FIG 5 BASIC FIRING CIRCUIT FOR HIGH-PRESSURE FLASH TUBES WHICH REQUIRE A SEPARATE IONIZING POTENTIAL

FIG 6 IMPROVED FIRING CIRCUIT USING THYRATRON RELAY TUBE TO CONTROL IONIZING POTENTIAL

PHOTO FLASH POWER SUPPLY

Here is another Thordarson FIRST . . . a typical example of Thordarson engineering skill that has helped established leadership in the field. This circuit features:

- **A.C. Line or Portable Battery Operation**
- **Charging Time — 10 to 15 Seconds**
- **A.C. Line Battery Recharge Feature**
- **Light Compact Low Drain Power Transformer**
- **Power Supply Output — 2250 V.O.C.**
- **Storage Condenser Delivers 75 Watt-Sec. Energy Element**
- **Adaptable Trigger Circuits for 2 or 3 Tubes**
- **Cold Cathode Rectifiers Employed in a Voltage Doubling Circuit**

OUR ENGINEERING STAFF IS AVAILABLE TO SOLVE YOUR PROBLEMS FOR YOU UPON REQUEST

Whatever your position in the field of electronics Thordarson can serve you better. Our large variety of stock types fill almost every need. For extraordinary conditions, send us your problems and our engineering staff will come up with the right answers.

1. LOW VOLTAGE - LOW CURRENT
2. LOW VOLTAGE - HIGH CURRENT
3. HIGH VOLTAGE - HIGH CURRENT
4. HIGH VOLTAGE - LOW CURRENT

WHERE QUALITY IS A NECESSITY...

Thordarson has the answers to many electrical problems that daily confront industry. With a background of over 25,000 active specifications in their files, built up over 53 years of leadership in the field, Thordarson supplies the leaders with their large variety of stock types of transformers as well as the hundreds of special types built to customer specifications or resulting from recommendations of our own engineers upon studying the various requirements submitted to them by industry.

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Export — Scheel International Inc.

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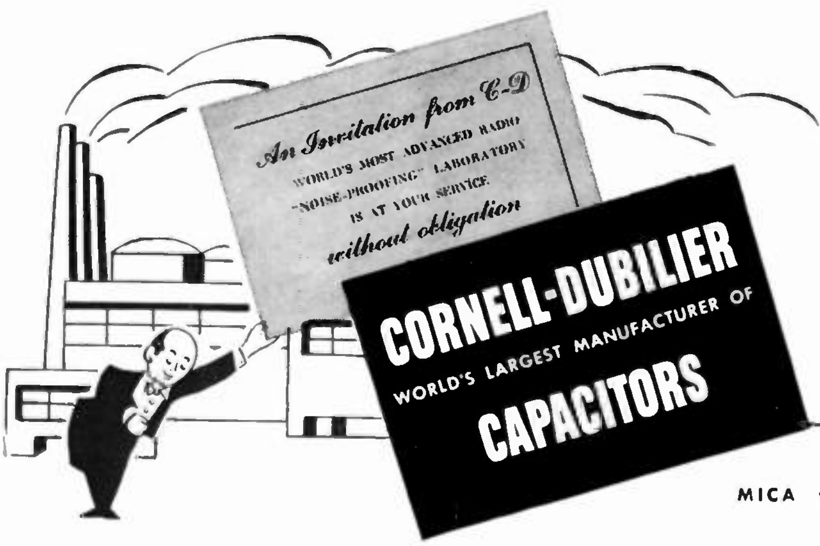
You may build the best appliance of its kind on the market — but if it sets up local radio interference—you'll have tough sledding against today's keen competition. Your customers are *demanding* radio noise-free performance in the electrical equipment they buy.

The answer, of course, is to equip *your* products with C-D Quietones. Why Quietones? First, because they're the best-engineered noise filters — second, because they guard your product's reputation by giving long trouble-free service — third, because they're designed and built to meet manufacturers' specific needs — efficiently and economically.

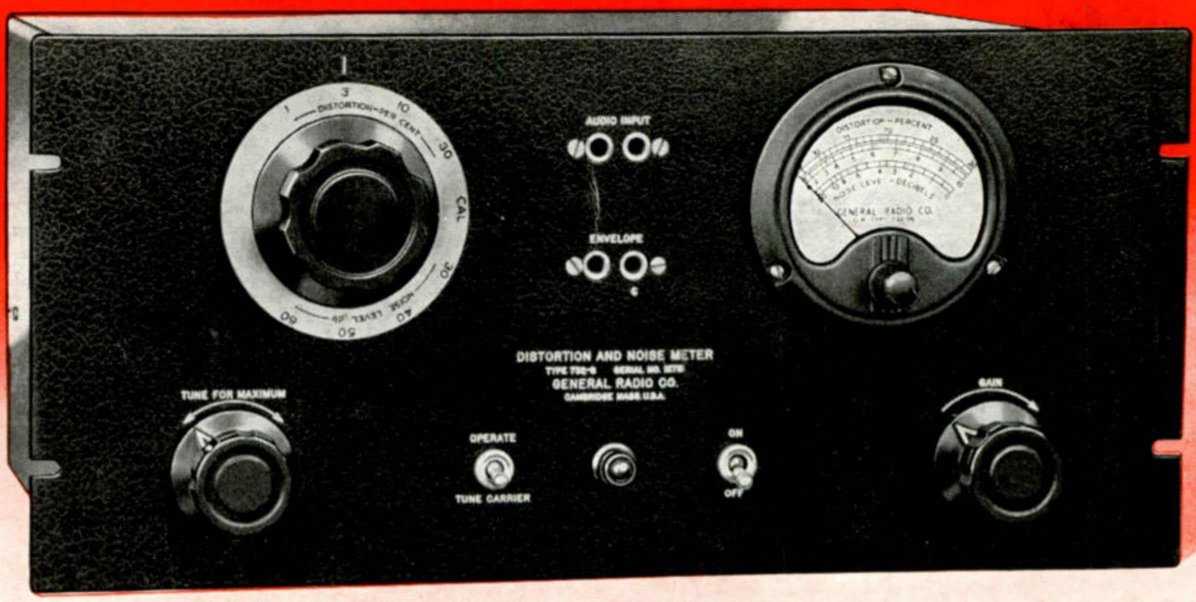
Speed up sales — build prestige — boost profits with C-D Quietones. Your inquiries are invited. Cornell-Dubilier Electric Corporation, Dept. M-3, South Plainfield, New Jersey. Other large plants in New Bedford, Brookline and Worcester, Mass., and Providence, Rhode Island.

Make Your Product More Saleable with C-D Quietone Radio Noise Filters and Spark Suppressors



MICA • DYKANOL • PAPER • ELECTROLYTIC

for A-F DISTORTION and NOISE MEASUREMENTS



. . . in Sound on Film and Disc Recordings

. . . in Production Tests on Radio Transmitters & Receivers

FROM necessity, because of war production, the pre-war very popular Type 732-B Distortion & Noise Meter was dropped from the G-R line. It is now in production again to meet an insistent demand for a meter to supplement the new Type 1932-A which is designed primarily for broadcast and communication applications.

The Type 732-B is equipped with a 400-cycle high-pass L-C filter so that harmonic content measurements of a 400-cycle signal can be made rapidly. Because of the width of the pass band, unsteady signals, "wows" and other irregularities do not affect the accuracy of measurement.

The ease with which accurate measurements can be made over the distortion range of 0.25 to 30% and noise range of 30 to 70 db below 100% modulation, make it very valuable in these types of production testing:

ON RADIO TRANSMITTERS

- **Signal-to-noise ratio**
- **Distortion vs** { **power**, **r-f levels**, **frequency**, **percentage modulation** }
- **A-F response**
- **Noise vs carrier level**
- **Hum modulation**
- **Hum level**

ON RADIO RECEIVERS

- **Distortion & noise vs a-f output**
- **Whistle output at 2nd and 3rd harmonic of i.f.**
- **Two-signal cross-talk**

The broad pass band characteristic of this meter is particulary useful when making distortion measurements on sound on film or on disc recordings where the fundamental frequency is not constant.

The Type 732-P1 Range Extension Filter is available as an auxiliary unit so that measurements at additional frequencies of 50, 100, 1000, 5000 and 7500 cycles can be made.

TYPE 732-B DISTORTION and NOISE METER $374.00
(For either 0.5 to 8 Mc or 3 to 60 Mc carrier range, specify which)

TYPE 732-P1 RANGE EXTENSION FILTER $209.00

WE HAVE A FEW IN STOCK. ORDER NOW FOR PROMPT SHIPMENT

GENERAL RADIO COMPANY Cambridge 39, Massachusetts

90 West St., New York 6 — 920 S. Michigan Ave., Chicago 5 — 950 N. Highland Ave., Los Angeles 38