Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

January, 1951

Volume 39 Number 1

Bell Telephone Laboratories, Inc.

RADIAL-AND-LINEAR-LARGE-SCALE SCANNING

The sound field produced by an acoustic lens is visually plotted by a neon lamp associated with a minute microphone, both systematically passing through the field.

The following IRE Standards appear in this issue: Television, Methods of Measurement of Electronically Regulated Power Supplies; and Circuits, Definitions of Terms in Network Topology.

PROCEEDINGS OF THE I.R.E.

TABLE OF CONTENTS, INDICATED BY BLACK-AND-WHITE MARGIN, FOLLOWS PAGE 32A

The Institute of Radio Engineers

COMING

17,000 Engineers and Businessmen

MEET

at the IRE Convention and Radio Engineering SHOW

March 19-22, 1951 in New York City

at the Waldorf-Astoria Hotel and Grand Central Palace

"Advance with Radio-Electronics"

is the theme of the 1951 Convention and Show. Already 212 top-rate papers patterning the "Advance" of Radio-Electronic engineering have been submitted. 26 great technical sessions will organize these papers, plus 11 Professional Group Symposia.

256 Manufacturers and three Armed Forces Departments will present their own 1951 "Advance" in this vital, fast-moving field, where to miss a single year is to lose step with progress. Advancements in every division of electronics from television to components will be sorted out for your swift discernment by these far-sighted manufacturers.

Time spent at the Radio Engineering Show is time saved in procurement. In no other way can you see so much, as easily and quickly—or get the direct answers to your product questions. This national event has become a necessity to the practical engineer.

11 Professional Groups!

The explosive new development of "Professional Groups" in the IRE finds full expression in the 1951 Convention. Eleven of these groups have organized symposia which offer the attending engineer all the advantages of the big meetings and general interests plus the special interest coverage at his own professional level.

Tuesday is "Broadcast Day" for the station engineer, climaxed with the Tuesday Evening Television Symposium. Instrumentation, Audio, Nucleonics, Quality Control, are a few of the special professional interests that will be covered.

And to further the social contacts (as well as the technical aspects) of the engineers in these Professional Groups, tables at the Presidents' Luncheon on Tuesday will be identified to these groups so that you may lunch and talk with fellow engineers of your own interests. It is often as important to know men as to know facts!

"IRE Meetings and Shows Accelerate Electronic Progress"

STRAW VOTE—Help Us!

In order to estimate how many tables each Professional Group will need at The Presidents' Luncheon, please fill out this vote coupon—No obligation to come, of course.

Mail Early to:
Mr. J. Harold Moore, Chr. Presidents' Luncheon
American Telephone & Telegraph Co.
195 Broadway, New York 7, N.Y.
If I am able to attend the Presidents' Luncheon, I would request assignment to a "Group" table, or General Table as indicated. I think it is a fine idea.

"Interest Evaluation Coupon"

Group: ....

....

Name: ....

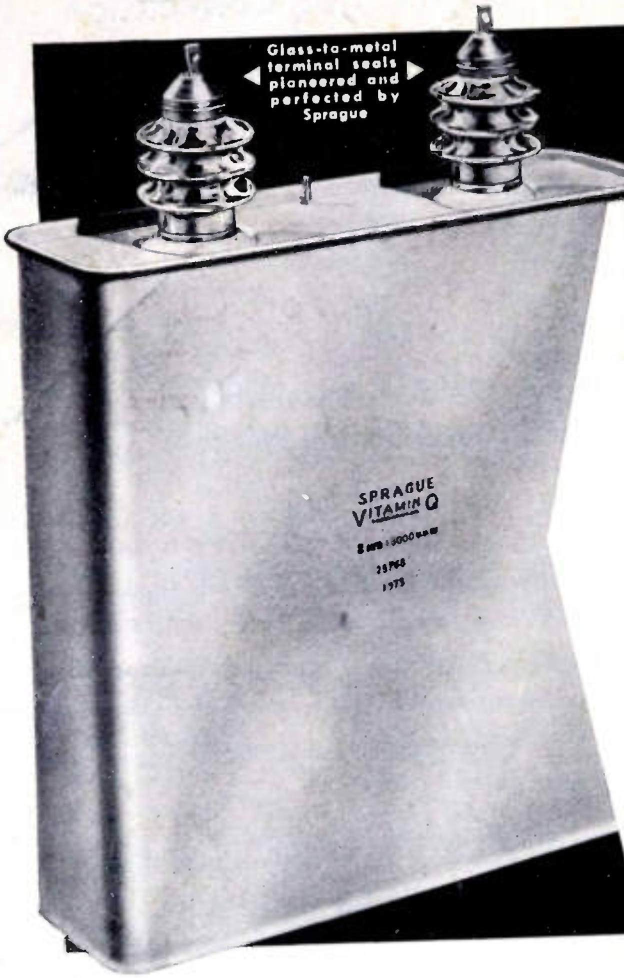
....

PROCEEDINGS OF THE I.R.E. January, 1951, Vol. 39, No. 1. Published monthly by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price per copy: members of the Institute of Radio Engineers $1.00; non-members $2.25. Yearly subscription price: to members $9.00; to non-members in United States, Canada and U.S. Possessions $18.00; to non-members in foreign countries $19.00. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

Table of Contents will be found following page 32A

SMALLER, LIGHTER, LESS EXPENSIVE

HIGH VOLTAGE D-C CAPACITORS



NOTE THE SIZES AND RATINGS!

Standard Vitamin Q Capacitors. Many special sizes and ratings also available.

Rated Cap Mfds.	D-C† Rated Voltage	DIMENSIONS Width	Depth	Can Height	Terminal Height	Cat. No.
2.0	8000	8 1/8	4 1/8	6	2 1/4	25P51
4.0	8000	8 1/8	4 1/8	9 1/4	2 1/4	25P52
6.0	8000	8 1/8	4 1/8	13	2 1/4	25P53
10.0	8000	13 1/2	4 1/4	13 1/8	2 1/4	25P54
12.0	8000	13 1/2	5 1/4	12 1/4	2 1/4	25P55
1.0	10000	8 1/8	4 1/8	5 1/2	3 11/16	25P56
2.0	10000	8 1/8	4 1/8	8 1/2	3 11/16	25P57
4.0	10000	13 1/2	4 1/4	9 1/4	3 11/16	25P58
6.0	10000	13 1/2	4 1/4	13 1/8	3 11/16	25P59
8.0	10000	13 1/2	5 1/4	12 7/8	3 11/16	25P60
1.0	12500	8 1/8	4 1/8	7 1/2	3 11/16	25P61
2.0	12500	8 1/8	4 1/8	12 1/4	3 11/16	25P62
4.0	12500	13 1/2	5 1/4	11 1/2	3 11/16	25P63
5.0	12500	13 1/2	5 1/4	13 3/4	3 11/16	25P64
1.0	16000	8 1/8	4 1/8	10 1/2	4 11/16	25P65
2.0	16000	13 1/2	4 1/4	12 1/4	4 11/16	25P66
3.0	16000	13 1/2	5 1/4	13 3/4	4 11/16	25P67
1.0	20000	13 1/2	4 1/4	11	4 11/16	25P68
1.5	20000	13 1/2	5 1/4	12 1/4	4 11/16	25P69

†Capacitors with voltage ratings above 10 KV are recommended for upright mounting only. For mounting in other positions, please supply complete application data for recommendation by Sprague engineers.

USE an ordinary capacitor rated for 40°C. operation on a high-voltage d-c filtering circuit and chances are the higher temperatures encountered will necessitate a serious de-rating. In other words, you will have to buy a larger, heavier and costlier capacitor than you actually need.

Standard Sprague high-voltage capacitors impregnated with Vitamin Q, however, are rated conservatively for operation at 85°C. *They require no de-rating up to this temperature.* Special units can be supplied for continuous use up to 105°C.

These capacitors are consistently superior in their ability to maintain a high degree of capacitance-temperature stability. Power factor is outstandingly low over a wide temperature range; d-c insulation resistance is notably high; and a-c ripple voltage at audio frequencies falls well within permissible limits. Equally important, Vitamin Q impregnated capacitors have a high safety factor at all temperatures, thus assuring long life.

Write for Sprague Engineering Bulletin 20

SPRAGUE VITAMIN Q* CAPACITORS

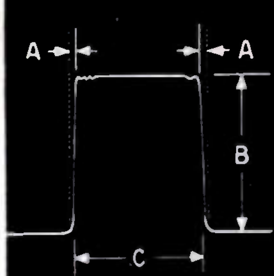
*Reg. U.S. Pat. Off.

SPRAGUE ELECTRIC COMPANY • NORTH ADAMS, MASS.

NEW hp GENERAL PURPOSE PULSE GENERATOR

-hp- MODEL 212A

TYPICAL 1 MICROSECOND PULSE INTO 50-OHM LOAD



A. 0.02 µsec rise and decay time. Minimum overshoot.

B. 50 watt peak power. (50 v. to 50 Ω load.)

C. Pulse length variable 0.07 to 10 µsec.

SPECIFICATIONS

PULSE LENGTH:
Continuously variable, 0.07 to 10 µsec. Direct reading panel control.

PULSE AMPLITUDE:
50 v. into 50 Ω load. Pos. & neg. pulses. 100 v. open circuit.

AMPLITUDE CONTROL:
Continuous control throughout range. 50 db in 10 db steps. 10 db fine adjustment.

INTERNAL IMPEDANCE:
50 Ω or less.

PULSE SHAPE:
Rise and decay time approx. 0.02 µsec. (10% to 90% amplitude.)

REPETITION RATE:
50 pps to 5,000 pps. Internally or externally controlled.

SYNC IN:
May be triggered by pos. or neg. pulse of 5 v. at rates up to 5,000 pps.

SYNC OUT:
50 v. into 200 Ω load. Approx. 2 µsec long. Approx. 0.25 µsec rise time.

PULSE DELAY:
Main pulse delayable 0 to 100 µsec from sync output pulse.

PULSE ADVANCE:
Main pulse can be advanced 0 to 10 µsec from sync output pulse.

POWER SUPPLY:
110/220 v; 50/60 cps.

SIZE:
Panel 10½" high, 19" wide. Depth 12".

PRICE:
$550.00 f.o.b. Palo Alto.

Data Subject to Change Without Notice

CONTINUOUSLY VARIABLE, HIGH POWER PULSES OF SUPERIOR WAVE FORM!

THIS NEW -hp- 212A PULSE GENERATOR saves you time and work testing "fast" circuits as well as making everyday laboratory checks of other generators, rf circuits, peak-measuring equipment, etc. It is the first commercial pulse generator to successfully combine broad laboratory usefulness with the fast rise time, high power, variable pulsing and other features demanded in radar, television and nuclear work.

ACCURATE PULSES AT END OF LONG TRANSMISSION LINE

The pulse length is continuously variable from 0.07 µsec to 10 µsec, and is varied by a direct reading panel control. Extremely fast rise and decay time, together with freedom from ringing or overshoot provide a virtually distortion-free pulse. A low internal impedance (50 ohms or less) insures a pulse shape virtually independent of load. This low impedance also makes it possible to deliver accurate pulses at a distance from the instrument, if the transmission lines are correctly terminated.

The Model 212A's repetition rate is continuously variable from 50 to 5,000 pps. It can be controlled internally, or from an external synchronizing source. Synchronizing pulses are available from the instrument either in advance of or following the output pulse. An amplifier-attenuator output system gives a low source impedance, and makes possible continuously variable pulse amplitude, positive or negative.

Brief specifications of this new -hp- instrument are shown in the adjoining column. For complete details... see your local -hp- representative... or write to the factory.

HEWLETT-PACKARD COMPANY
2040D Page Mill Road • Palo Alto, California
Export: FRAZAR & HANSEN, Ltd., 301 Clay St., San Francisco, Calif., U. S. A. Offices: New York, N.Y. and Los Angeles, Calif.

2040

hp laboratory instruments
FOR SPEED AND ACCURACY

Another reason why your telephone gives so much for so little

Studying punched card record of dial system operation. Each card (top) can report 1080 items

In a large, modern dial telephone office, 2,000,000 switch contacts await the orders of your dial—and 10,000 of them may be needed to clear a path for your voice when you make a single telephone call. Within this maze of signal paths, faults—though infrequent—must be detected and fixed before they can impair telephone service.

The latest system developed by Bell Telephone Laboratories automatically detects its own faults, detours calls around them without delay—then makes out a "written" report on what happened.

The fault may be a broken wire, or high resistance caused by specks of dirt on switch contacts. In one second, the trouble recorder punches out a card, noting in detail the circuits involved and the stage in the switching operation where the fault appeared.

Maintenance men examine the reports at intervals and learn what needs attention. Between times they go about their own duties in keeping service moving.

This is another example of how research at Bell Laboratories helps your telephone system operate at top efficiency, so the cost to you stays low.

BELL TELEPHONE LABORATORIES

AMERICAN TELEPHONE & TELEGRAPH CO. AND ASSOCIATED COMPANIES — BELL SYSTEM

WORKING CONTINUALLY TO KEEP YOUR TELEPHONE SERVICE BIG IN VALUE AND LOW IN COST

PRODUCED FOR YOUR REQUIREMENTS
- by the top specialists in the ceramic field

HI-Q

CERAMIC PLATE CAPACITORS

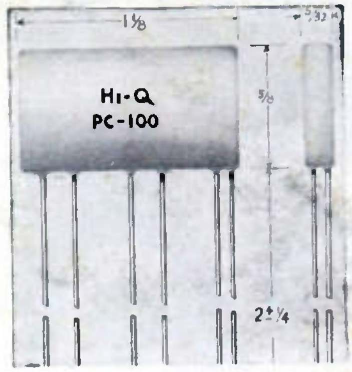
Essentially similar, except in shape, to **HI-Q** Disk Capacitors except that in the multiple units they do NOT have to have a common ground as is the case with disks. These **HI-Q** Plates can be produced in an unlimited range of capacities, the number on a plate being limited only by the K of the material and the physical size of the unit. They offer the greatest available capacity per unit volume of any type condenser on the market.

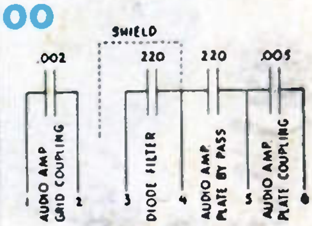
Guaranteed minimum values of capacity up to 33,000 mmf per sq. in. are available. This is based on the use of Body 41 ceramic having 3000 as a dielectric constant "K" and .020 in. thickness and the formula:

$$C\ (\text{mmf}) = \frac{.224\ K\ A\ (\text{Sq. in.})}{D\ \text{in.}}$$

If temperature compensating ceramics are used, the capacity will be considerably lower. Typical circuits are shown here, but almost any combination can be produced for your specific needs. Consult our engineers for complete details. Write for new **HI-Q** datalog.

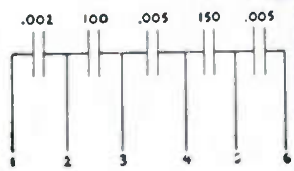
PC-100





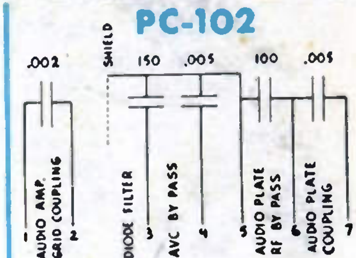
1-2	.002	+75 −25 %	mf
3-4	220	+50 −50 %	mmf
4-5	220	+50 −50 %	mmf
5-6	.005	+75 −25 %	mf

PC-101



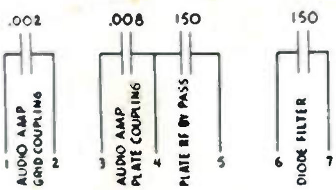
1-2	.002	+75 −25 %	mf
2-3	100	+50 −50 %	mmf
3-4	.005	+75 −25 %	mf
4-5	150	+50 −50 %	mmf
5-6	.005	+75 −25 %	mf

PC-102



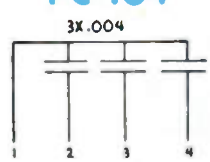
1-2	.002	+75 −25 %	mf
3-5	150	+100 −0 %	mmf
4-5	.005	+100 −0 %	mf
5-6	100	+75 −25 %	mmf
6-7	.005	+75 −25 %	mf

PC-103



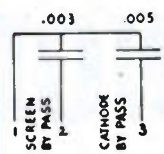
1-2	.002	+75 −25 %	mf
3-4	.008	Min %	mf
4-5	150	+50 −50 %	mmf
6-7	150	+50 −50 %	mmf

PC-104



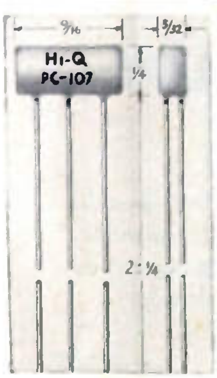
3X .004 Min mf

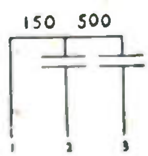
PC-105



1-2	.003	+75 −25 %	mf
1-3	.005	+75 −25 %	mf

PC-107





1-2	150	+75 −25 %	mmf
1-3	500	+75 −25 %	mmf

HI-Q COMPONENTS

Capacitors
Trimmers • Choke Coils
Wire Wound Resistors

BETTER 4 WAYS

- ✓ PRECISION
- ✓ UNIFORMITY
- ✓ DEPENDABILITY
- ✓ MINIATURIZATION

JOBBERS—ADDRESS: 740 Belleville Ave., New Bedford, Mass.

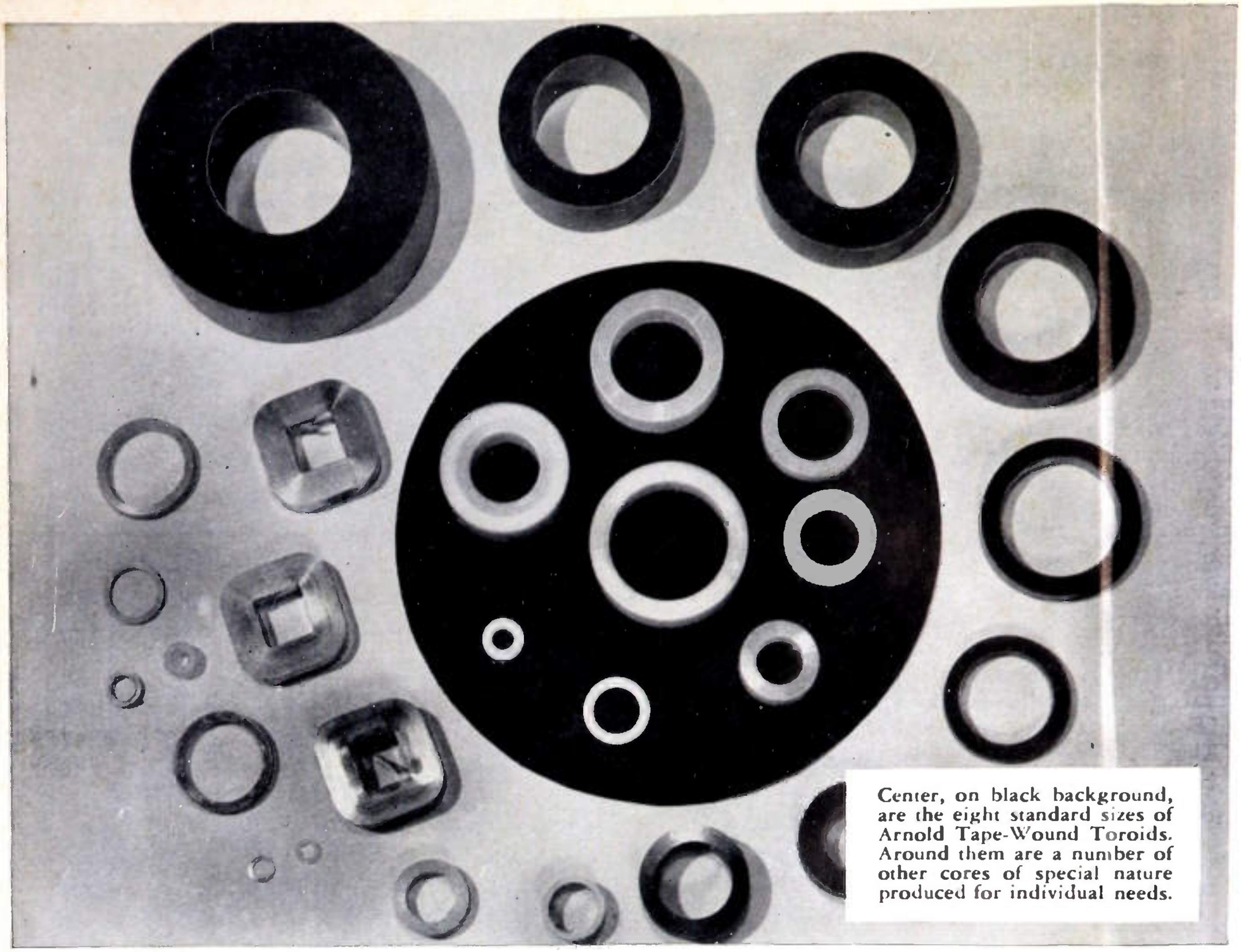
HI-Q

Electrical Reactance Corp.

OLEAN, N. Y.

SALES OFFICES: New York, Philadelphia, Detroit, Chicago, Los Angeles

PLANTS: Olean, N. Y., Franklinville, N. Y. Jessup, Pa., Myrtle Beach, S. C.

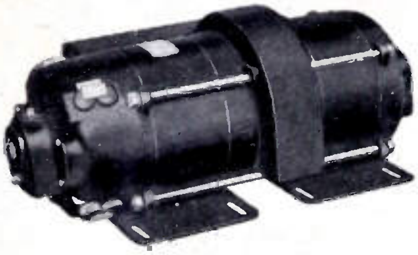


Center, on black background, are the eight standard sizes of Arnold Tape-Wound Toroids. Around them are a number of other cores of special nature produced for individual needs.

ELECTRONICS

Designers

AMPLIDYNES—500 to 25,000 WATTS



▲ 1-kw Amplidyne motor-generator set

▼ 5-kw Amplidyne generator

◀ 3-kw Amplidyne motor-generator set

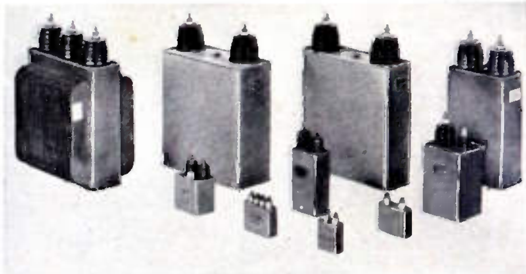
PRECISE CONTROL OF position · torque · speed tension · power factor · voltage · current

The General Electric amplidyne is a simple d-c generator which, through the arrangement of field and armature circuits, possesses extremely high speed of response and amplification.

First used in radar and fire-control apparatus, it now has many new jobs. That's why G-E amplidyne generators and motor-generator sets are made in a wide variety of sizes and frames with output ratings from 500 watts to 25 kilowatts.

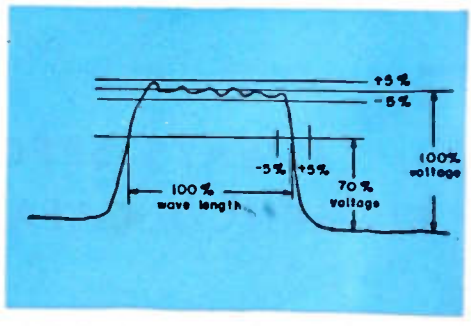
What are your requirements? For further data, write, giving complete details, to *Electrical Industries Sec., Resale Industries Div., Apparatus Dept., General Electric Co., Schenectady 5, N. Y.*

A G-E 25-KW AMPLIDYNE AMPLIFIES A 9/10 WATT INPUT 22,200 TIMES



PULSES —MADE TO ORDER

Specially designed General Electric Type-E networks will generate pulses within ±5 per cent of any length you require from 0.1 to 40 microseconds. These networks consist of capacitor and coil sections adjusted to close tolerances and hermetically sealed in single metal containers. In the last war G-E Type-E networks were produced on a large scale to meet radar demands. Now they are available for commercial or military use in a wide variety of designs, impedances, ratings, and sizes. See Bulletin GEA-4996.

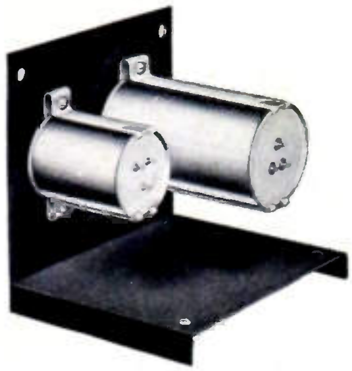


Typical design and operational limits of G E Type-E pulse-forming network: Ripple at top of pulse ±5%; Wave length ±5% measured at 70% amplitude; Capacitance tolerance ±10%

GENERAL ELECTRIC

Digest

TIMELY HIGHLIGHTS ON G-E COMPONENTS



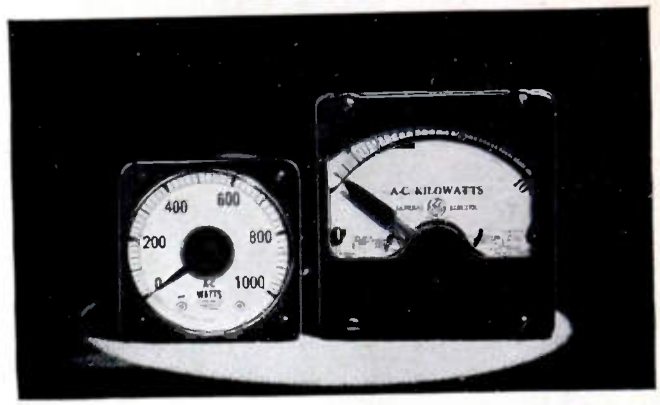
OIL-IMMERSED SELENIUM RECTIFIERS —use them "ANYWHERE"!

G-E hermetically sealed, oil-immersed selenium rectifier stacks make it possible for you to design metallic rectifiers into equipment that will be subjected to corrosive fumes, salt air, dust, fungus, or other atmospheric conditions. Because they're immersed in oil, these stacks will stand higher current drains than equivalent-size "open" units. Available in single- or full-wave circuits. Ratings: from 12 to 180 d-c volts output, 15.5 to 270 a-c volts input, .25 to 27.0 d-c amps. Write for complete data on ratings and dimensions to *Electrical Industries Section, Resale Industries Division, Apparatus Dept., General Electric Co., Schenectady 5, N. Y.*

PUSH-BUTTON STATIONS—make your selection from the COMPLETE G-E LINE

There's a General Electric push-button station or unit for virtually any electronic application. The complete line includes dozens of types. All stations have sturdy nonbreakable steel frames and covers with ample clearance between terminals. G-E units for built-in applications have terminals anchored to a molded base for firm support. Contact maintenance on all stations and units is virtually unnecessary because large fine-silver double-break contacts are used. For full data, check Bulletin GEA-3469.

NEW! SHADOW-PROOF DIALS MAKE SWITCHBOARD INSTRUMENTS EASIER-TO-READ



Here is a new switchboard instrument that can be read easily—anytime. Its dial can be clearly illuminated from almost any angle because it is set forward flush with the front of the case. A protruding anti-glare convex-type glass front prevents reflections. The new meter is available in 4¼- or 8¾-inch models, both with long 250-degree scales. D-c ammeters, volt-ammeters; a-c ammeters, voltmeters, wattmeters, frequency and power-factor meters, temperature indicators, and synchroscopes. Send for Bulletin GEC-218.

General Electric Company, Section A 667-10
Apparatus Department, Schenectady 5, N. Y.

Please send me the following bulletins:

(√) Indicate for reference only
(X) for planning an immediate project

☐ GEA-3469 Push-button stations
☐ GEA-4996 Pulse-forming networks
☐ GEC-218 Switchboard instruments

Name ____________________

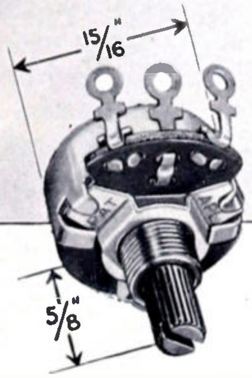
Company ____________________

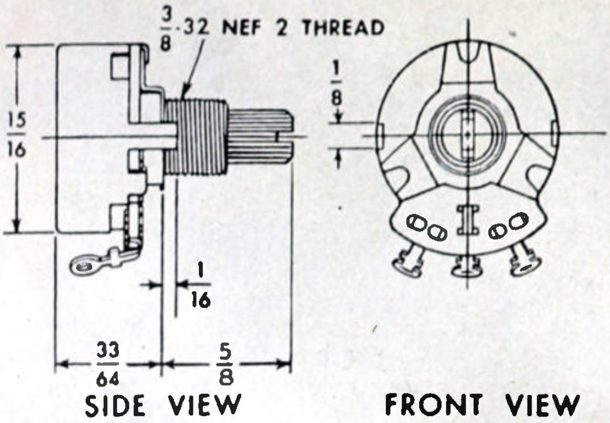
Address ____________________

City ____________ State ____________


OHMITE
Has Just the Resistor You Need
Ohmite offers fixed, adjustable, tapped, non-inductive, and precision resistors in more than 60 sizes and 18 types of terminals, in a wide range of wattages and resistances.
These rugged resistors have proved their dependability under the toughest conditions. Write on company letterhead for Catalog 40.
Be Right with
OHMITE
Reg. U. S. Pat. Off.
RHEOSTATS · RESISTORS · TAP SWITCHES
OHMITE MANUFACTURING COMPANY
4860 Flournoy St., Chicago 44, Ill.

Service Beyond The Sale!





Design · Standardization By Mallory

effects real customer savings!

15/16" MALLORY MIDGETROL*

Electrical characteristics specially designed for critical applications in television, radio and other circuits. Insulated shafts are knurled for ease in adjustment. Current-carrying parts provide 1500 volt insulation . . . 15/16" diameter saves space . . . phenolic material eliminates mechanical noise. Precision-controlled carbon element provides smooth tapers, quiet operation, accurate resistance values, less drift in television applications.

**Trade Mark*

Mallory goes beyond the basic research and development work which results in totally new products . . . and utilizes every opportunity to pass on to customers the savings effected by product standardization.

Such is the case with the well known Mallory Midgetrol. Standardizing its diameter and shaft design resulted in cost reductions for radio and television manufacturers. The standard molded phenolic shaft provides a combination hand knurl and screwdriver slot adjustment at no extra cost . . . and it is available with either 1/4" or 3/8" bushing length. In addition, the Mallory Midgetrol occupies less space than larger controls, with no sacrifice in wattage rating.

That's service beyond the sale!

Mallory's electronic component know-how is at your disposal. What Mallory has done for others can be done for you!

Television Tuners, Special Switches, Controls and Resistors

P. R. MALLORY & CO., Inc.

MALLORY

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

SERVING INDUSTRY WITH

Electromechanical Products
Resistors — Switches — TV Tuners — Vibrators

Electrochemical Products
Capacitors — Rectifiers — Mercury Dry Batteries

Metallurgical Products
Contacts — Special Metals — Welding Materials

OPPONENT: TIME

A top bobsled team—athletes trained to work as one in handling their balanced, ruggedly built sled—can consistently defeat the clock in negotiating the treacherous, one-mile Mt. Van Hoevenberg run.

Likewise, T. E. I. engineers and production personnel work as a unit with a similar goal—the conquering of TIME's deteriorating effects through the building of an ever-stronger picture tube. For Thomas' highly trained personnel, specially designed equipment, and efficient production techniques are consistently increasing the life of Thomas tubes in the contest with TIME.

For the greatest value in today's television picture tubes—for top operating efficiency and truly LONGER life—specify T.E.I. In all popular rectangular sizes, black-face.

THOMAS ELECTRONICS, Inc.
118 Ninth Street Passaic, New Jersey

New television microphone, developed at RCA Laboratories, virtually vanishes when in active use.

Vanishing Microphone lets the stars shine

Now you see it, now you don't! RCA's new "vanishing microphone" is plainly visible when standing alone—but let a television performer stand before it and it seems to disappear.

Called the "Starmaker," this RCA microphone is little larger than a big fountain pen . . . and principles of design based on modern camouflage blend it with an artist's clothing. There's no clumsy "mike" to distract your attention from the artist—and it's also a superbly sensitive instrument.

Through research carried out at RCA Laboratories, the "Starmaker" microphone picks up sound from all directions—hears and transmits every sound the human ear can detect. It's not only small and almost invisible, but it's also one of the most efficient microphones ever devised.

* * *

See the latest wonders of radio, television, and electronics at RCA Exhibition Hall, 36 West 49th Street, New York. Admission is free. Radio Corporation of America, RCA Building, Radio City, New York 20, New York.

Known for brilliant pictures, RCA Victor's 1951 home television receivers also have the finest of sound systems—RCA Victor's "Golden Throat."

RADIO CORPORATION of AMERICA

World Leader in Radio — First in Television

SEGMENTED DEFLECTION YOKE CORES

This popular 4-segment design is highly efficient. It is easy to handle in TV production work and assures a minimum of breakage. 2-segment types are also available.

STACKPOLE *Ceramag*® ... THE CERAMIC CORES THAT SET THE QUALITY STANDARDS

The tremendous advance in the use of metallic oxide (non-metallic) cores has been due in large part to Stackpole powder molding experience which paved the way to fully dependable units in production quantities. Stackpole Ceramag Cores assure lower losses with higher operating efficiency, lower operating temperatures, lighter weight, smaller sizes, maximum permeability, less corona effect and minimum cost. Ceramag cores are made in two grades for high and low flux densities.

"U" and "E" CORES FOR FLYBACK TRANSFORMERS

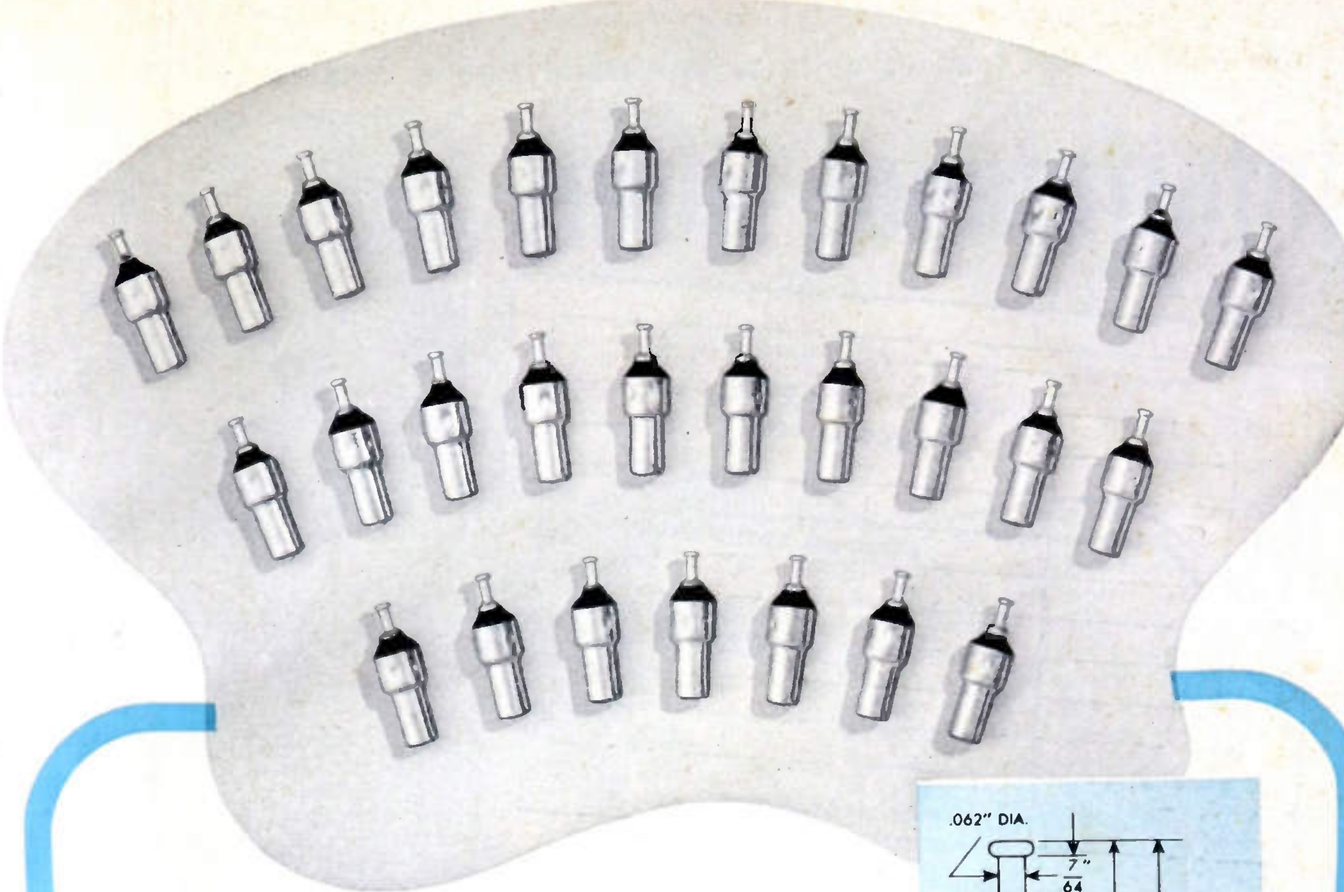
Permeability of these Stackpole Ceramag Cores is of the order of 10 to 1 by comparison with conventional iron cores. They are materially smaller, have higher resistance and operate much cooler due to the absence of eddy current losses. Many special types are regularly produced.

TELEVISION IMAGE W-I-D-T-H CONTROL TYPES

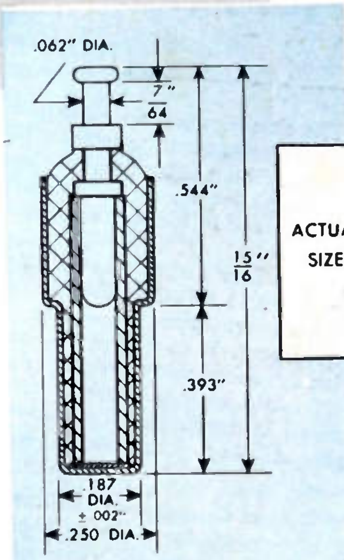
These Stackpole Ceramag Cores assure remarkably higher standards of efficiency for TV horizontal image deflection circuits. In areas where there is a low line voltage, they give ratios of from 1 to 8 or more compared with 1-5 for previous high permeability types.

Electronic Components Division

STACKPOLE CARBON COMPANY, St. Marys, Pa.



ERIE STYLE 325 . . . *A Rugged Stand-Off . . . Quickly Installed . . . Makes a Neater Chassis . . . Provides Uniform, Efficient By-passing*



Manufacturers of TV and other high frequency receivers have welcomed the Erie Type 325 Ceramicon because its distinctive features give the answer to many production problems. Consider these advantages:

1. For the first time in a hermetically sealed case, a by-pass to ground is provided through the shortest possible path, by taking advantage of the concentric cylindrical electrode configuration and making connection to the outer shell at the plane of the chassis.
2. The design provides extremely low and uniform series inductance and effective v.h.f. by-pass.
3. In assembly operations terminals and lead lengths are fixed, resulting in better mechanical uniformity.
4. High speed assembly is facilitated through use of a standard push-on clip. For more critical applications shell may be soldered directly into a hole in the chassis.
5. Post terminal matches tube socket terminal height to maintain uniform short leads, and provides a sturdy tie point for several connections.
6. Unusual mechanical ruggedness minimizes danger of breakage in installation and in use.

Available in 10, 33, 47, 68, 82, 100, 680, 1,000 and 1,500 MMF capacity. 500 Volts D. C. working.

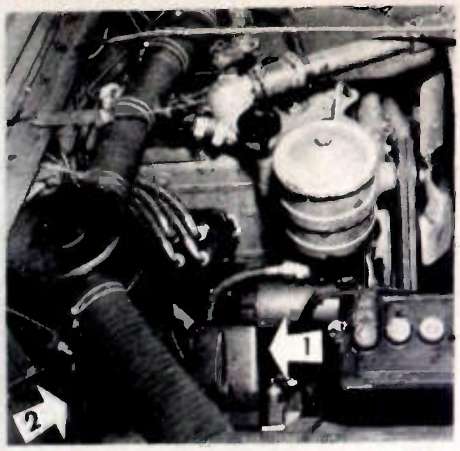
Electronics Division

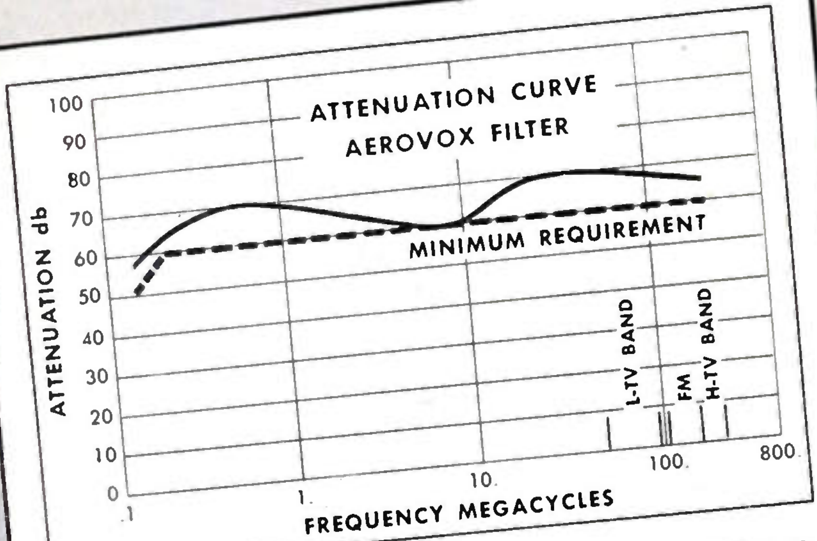
ERIE RESISTOR CORP., ERIE, PA.

LONDON, ENGLAND • • TORONTO, CANADA

Uncle Sam's latest jeep as quiet as proverbial mouse, because of AEROVOX

Interference Filters





● The chart sums it up. Note how radio interference generated by the ignition system and other electrical equipment is suppressed well in excess of requirements.

Uncle Sam's new jeep includes The Electric Auto-Lite Company's 24-volt waterproof electrical equipment. It must operate efficiently even under water. And radio interference must be minimized in the interests of dependable military communications.

Long hours of cooperative research and engineering were spent on this noise-suppression problem. The main considerations were filters to minimize interference originating with the voltage regulator, the generator and the ignition system. Aerovox finalized the complete answer based on the three major units here presented.

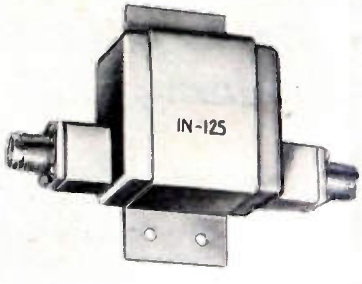
• • •

And *thoroughly waterproof, weatherproof and shockproof,* of course.

● *Capacitance applications such as this are all in the day's work for Aerovox engineers. Whatever your capacitance problems and requirements may be, Aerovox will fit the right capacitors to the right applications. Address Dept. FI.*

Aerovox Type 89ZAY using a metallized paper capacitor and mounting inside voltage regulator case to work in conjunction with IN-127.

Aerovox Type IN-127 mounted inside voltage regulator (Arrow No. 1) and acting as interference eliminator for voltage regulator and generator systems.



Aerovox IN-125 which mounts on bulkhead of jeep (Arrow No. 2) and suppresses interference originating in ignition system.

AEROVOX capacitors

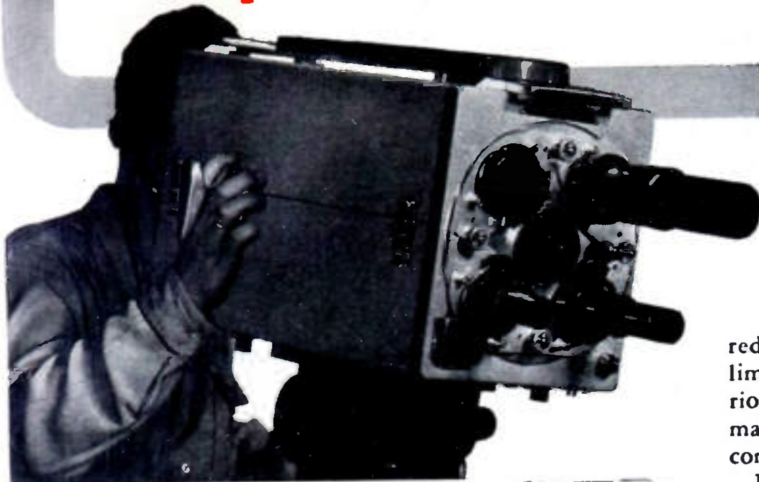
FOR RADIO-ELECTRONIC AND INDUSTRIAL APPLICATIONS

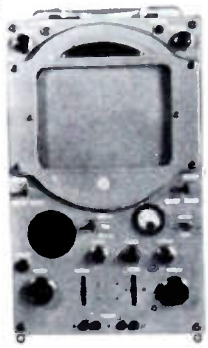
AEROVOX CORPORATION, NEW BEDFORD, MASS., U. S. A.

Sales Offices in All Principal Cities • Export: 41 E. 42nd St., New York 17, N. Y.

Cable: AEROCAP, N. Y. • In Canada: AEROVOX CANADA LTD., Hamilton, Ont.

GPL Introduces First TV Camera Chain Designed from Start to Finish for Compactness and Ease of Operation





Compact GPL camera and control unit have been "human engineered" for easy, efficient use. Camera provides uniform focus adjustment for all lenses; iris is motor-controlled from rear of camera or from control unit, with lens opening shown on dials at both locations. Control unit has 8½" monitor tube.

IMPROVED SYNC GENERATOR

The sync generator, with its power supply, is a single unit, packaged for field use. Because binary counting circuits are used, and pulse width is controlled by delay lines, it provides circuit reliability better than present studio equipment. With this circuitry, all operator adjustments are eliminated.

Built with the compact precision which distinguishes a quality watch from an alarm clock, GPL's new image orthicon camera chain is smaller, lighter, easier to use. It is the first camera chain that has been "human engineered" – designed from motion studies of cameramen and control personnel. It is the first with type and location of controls based on minimum movement and maximum ease and efficiency.

This simplification, together with size and weight reduction has been accomplished without any sacrifice or limitation whatever in performance or accessibility. Superior GPL circuit design provides a better picture than normally obtainable with image orthicon equipment. Complete control is provided for every studio or field requirement.

Logical components have been combined . . . fewer units make up a chain. A single chain consists of only 4 units; a triple chain, 12 including switching unit and master monitor. The camera, *with integral view finder*, is only 10¾" x 12½" x 22", weighs 75 lbs. instead of 100-105 lbs. The sync generator is a single portable unit including its own power supply. It may be easily removed from its case to go into a standard relay rack.

SIMPLIFIED CONTROL

All controls are at the finger-tips of cameramen and camera control operators. Focus adjustment of all lenses is uniform; a given rotation of focus control produces the same shift in plane of focus for all lenses. The iris is motor-controlled, either from the rear of the camera or from the camera control unit. Dials on both camera and control unit indicate the lens opening. Negative feedback is used to stabilize video frequency response, eliminating an adjustment. Target and beam are controlled by thumbwheels next to convenient knobs for pedestal and gain.

READILY ADAPTABLE

GPL Camera Chains completely meet all studio and field requirements or may be readily adapted to supplement existing installations. *Before you make any camera chain investment, get all the facts on this new addition to GPL's outstanding line of TV studio equipment.*

Write, Wire or Phone for Details

TV Camera Chains • TV Film Chains
TV Field and Studio Equipment
Theatre TV Equipment

GENERAL PRECISION LABORATORY INCORPORATED

General Precision Laboratory
INCORPORATED
PLEASANTVILLE NEW YORK

DU MONT TYPE 280-A

High-voltage operation, high sensitivity, and a 10-megacycle bandwidth are provided in the Type 280-A along with the calibrated sweep delay, versatile time-base synchronization, and sweep durations continuously variable from one to 15000 microseconds. These and many other precision features make the Type 280-A important quantitative equipment in industrial applications where detailed study of high-speed phenomena contributes to the evaluation of the quality and design of products.

Supplementing the usual sweep-control circuits of the cathode-ray oscillograph, the Type 280-A also contains a specialized video synchronizer unit. Used in conjunction with the calibrated sweep delay, it will select and display any line or fraction of a line of the standard RTMA television signal.

FEATURES

- Type 5XP- high sensitivity Cathode-ray Tube.
- 10-megacycle Video Amplifier.
- Pulse Response 0.01 μsec. rise time reproduced as a rise time of 0.04 μsec. or less.
- Video Synchronizer Unit—Field and line selector for a study of any portion of the composite television signal.
- Sweep Calibration — 10-, 1-, 0.2-microsecond intervals.
- Calibrated Sweep Delay Ranges—100 μsec., 1000 μsec.; Accuracy ±0.1% full scale.
- Internal Trigger Generator—120 to 2500 cps.; can be synchronized to external signals from 120 cps. to 50,000 cps. Sweep may be triggered externally.
- Output Trigger—0 or 25 μsec. delay with respect to calibrated delay; 50 volts positive; rise time, 0.5 microseconds or less.
- Output Test Pulse—variable to 20 volts positive; duration, 1 μsec.; rise time, less than 0.01 μsec.; 25-usec. delay with respect to the start of the calibrated delay.

CALIBRATED SWEEP DELAY

measures fractions of ONE MICROSECOND!

The precision sweep delays of the Du Mont Types 280-A and 256-D Cathode-ray Oscillographs provide for the accurate measurement of small time intervals.

provides DIRECT READINGS in Du Mont precision cathode-ray oscillographs

DU MONT TYPE 256-D

The Type 256-D is designed as a highly precise time-measuring device. Its versatility serves many of the general-purpose functions in laboratory work.

The calibrated sweep delay of the Type 256-D will measure time intervals up to 1000 microseconds with an accuracy of ±0.1% of the full-scale ranges of 100 μsec. or 1000 usec. With both delayed and undelayed sweeps, a movable marker in the Type 256-D will indicate that portion of the sweep which is expanded on the shorter delayed sweeps. Delayed sweeps are of 4-, 10-, and 25-microsecond durations. Undelayed sweeps are available in six ranges from 4 to 4500 microseconds.

Response of the video amplifier of the Type 256-D is within ±1 db at 20 cps; down no more than 3 db at 8 megacycles, no more than 6 db at 11 megacycles. Sensitivity is 0.7 peak-to-peak volt per inch. Pulse response is such that a rise time of 0.01 microsecond will be reproduced as a rise time of 0.04 microsecond or less.

Crystal-controlled timing markers are provided in the Type 256-D for calibration of the delay circuit sweeps. Both delayed and undelayed sweeps may be started by external trigger pulses of either polarity or by a built-in trigger generator which provides 1-microsecond pulses of either polarity, having a rise time of 0.3 microsecond and amplitude greater than 100 volts. Trigger repetition rates up to 2000 P.P.S. are usable.

Please address all inquiries concerning these precision cathode-ray oscillographs to:

ALLEN B. DU MONT LABORATORIES, INC., INSTRUMENT DIVISION, 1000 MAIN AVENUE, CLIFTON, NEW JERSEY

A page from the note-book of Sylvania Research

Color Television Research Accelerated by Study of Color Phosphors in Demountable Vacuum System

Sylvania's physicists, chemists and engineers have carried out fundamental research and development on color television tubes for a considerable period of time.

These basic investigations into practical color television tubes for home and industry have been conducted at Sylvania's Research Laboratories at Towanda, Pennsylvania, and Bayside, New York. Typical of the exceptional facilities and advanced research techniques used at the Sylvania laboratories is the demountable vacuum system shown in the photo.

Developmental samples of phosphor screens for color television tubes may be placed in the tube for study and test. Similarly developmental models of three-color electron guns may be placed in the neck of the demountable evacuated envelope and characteristics studied.

This continual basic research has enabled Sylvania to pioneer in providing the television tube industry with a complete selection of phosphors for color picture tubes. It has also paved the way toward more advanced research into the design and construction of color television tubes and circuits.

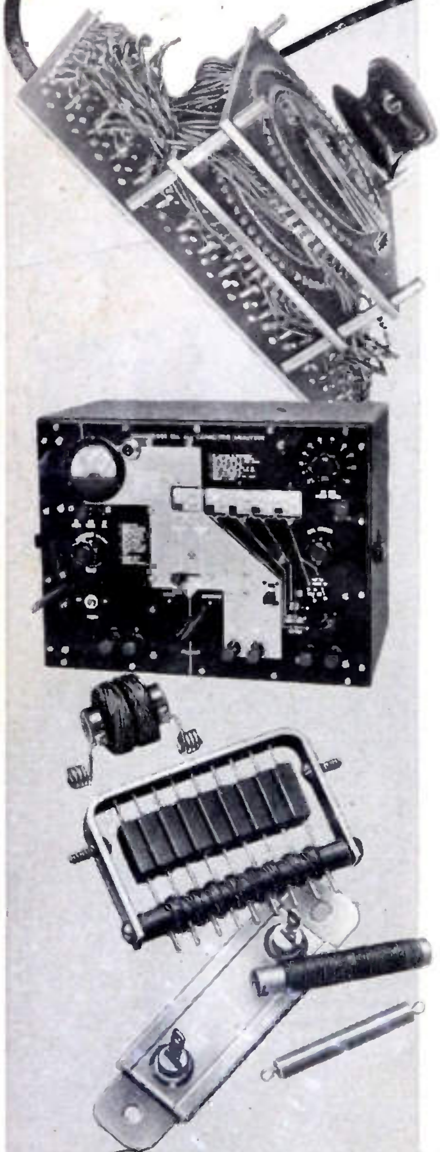
It is typical of Sylvania's never-ceasing emphasis on *productive basic research.*

The operator is observing a spot of light on the phosphor screen of a developmental color television tube. By varying voltages applied to a three-color electron gun inserted in the demountable vacuum system, she can obtain data useful in improving both phosphors and gun designs for color television tubes.

SYLVANIA ELECTRIC

RADIO TUBES; TELEVISION PICTURE TUBES; ELECTRONIC PRODUCTS; ELECTRONIC TEST EQUIPMENT; FLUORESCENT TUBES, FIXTURES, SIGN TUBING, WIRING DEVICES; LIGHT BULBS; PHOTOLAMPS; TELEVISION SETS

A RELIABLE SOURCE FOR YOUR *Custom Built* Electric-Electronic Specialties



Leading equipment manufacturers find that it pays to turn specialized assignments over to Shallcross for development, design or production . . .

From critical components to sub-assemblies and instruments, Shallcross' broad experience and precision facilities assure better results . . .

Often, they assure an appreciable cost saving as well.

- A capable staff of electrical, electronic, mechanical, chemical and instrumentation engineers . . .
- A fully equipped plant . . .
- Plus over 20 years of specialization in high quality products for military, industrial and public utility use . . . are here at your disposal.

AMONG RECENT SHALLCROSS CUSTOM-BUILT ASSIGNMENTS HAVE BEEN:

ROTARY SWITCHES
POTTED AND THERMALLY-CONTROLLED R-C NETWORKS
PRECISE DECADES AND NETWORKS FOR COMPUTER DEVICES
CALIBRATING INSTRUMENTS FOR STRAIN GAUGE BRIDGES
HIGH RESISTANCE STANDARDS
CRITICAL COIL ASSEMBLIES
HERMETICALLY SEALED CHOKES
HIGH-VOLTAGE MEASURING EQUIPMENT, ETC.

SHALLCROSS

ENGINEERS
DESIGNERS
MANUFACTURERS

SHALLCROSS MANUFACTURING CO.
COLLINGDALE, PA.

Something New

NEW OVAL SELECTOR SWITCHES

Several new oval rotary selector switches are described in Bulletin L13 just issued by the Shallcross Manufacturing Co., Collingdale, Pa. Six basic plates and three rotor types produce switches having from one to three poles per deck or gang and with other desired mechanical and electrical details. As many as 18, 9 or 6 positions may be obtained in single-, double-, or triple-pole types respectively. These may be single-, double-, or triple-pole decks exclusively or a combination of different types.

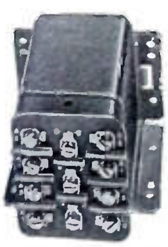
VERTICAL STYLE PRECISION RESISTORS FOR JAN USES

Improved vertical style precision wire-bound resistors for use where mounting requirements make it desirable to have both terminals at the same end of the resistor have been introduced by the Shallcross Manufacturing Co., Collingdale, Pa. These units provide a longer leakage path from the mounting screws to the terminals. Known as Shallcross Types BX120, BX140, and BX160, they are designed to meet JAN requirements for styles RB40B, RB41B and RB42B respectively. For commercial uses, the resistors carry somewhat higher ratings than for JAN applications. Wire leads instead of terminals can be furnished if desired. Complete details will gladly be sent on request to the manufacturer.

FLAT, METAL-ENCASED WIRE-WOUND RESISTORS

Flat, metal-encased, Type 265A wire-wound power resistors introduced by the Shallcross Manufacturing Company, Collingdale, Pa., are space wound, have mica insulation, and are encased in aluminum for mounting flat against a metal chassis. At 175° C. continuous use they are conservatively rated for 7½ watts in still air and 15 watts when mounted on a metal chassis. Write for Bulletin 122.

RELAYS BY GUARDIAN



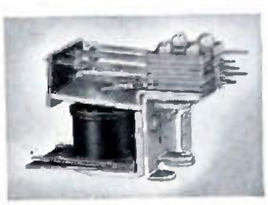
Screw Terminal
Hermetic Housing

RESOLVE IN 1951 to improve your products . . . especially radio, armament and power units coming under government contract. Let Guardian, No. 1 supplier of *improved* electrical controls, produce single units or complete control assemblies for you . . . in quantity. Guardian GOVERNMENT APPROVED CONTROLS will insure smooth, satisfactory performance.

The Series 335 D. C. Relay shown above, is typical of Guardian units built to the rigorous standards of aviation relays. Furnished open, or with a variety of HERMETICALLY SEALED housings.

New Hermetically Sealed Relay catalog 5-H is yours for the asking.

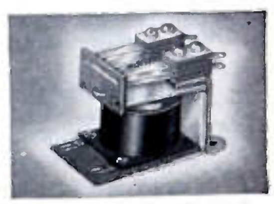
Series 595 D.C.



Series 210 A.C.—215 D.C.

Series 220 A.C.

Series 695 D.C.



Series 610 A.C.—615 D.C.

Write— ASK US TO MAKE SPECIFIC RECOMMENDATIONS. NO OBLIGATION.

GUARDIAN G ELECTRIC

1628-A W. WALNUT STREET CHICAGO 12, ILLINOIS

A COMPLETE LINE OF RELAYS SERVING AMERICAN INDUSTRY

Another Masterpiece of Truscon Engineering Skill

TRUSCON "G-W"

UNIFORM CROSS SECTION GUYED RADIO TOWER

HERE again is another example of Truscon leadership in antenna design—another example of skill gained by nearly a half-century of experience in the fabrication of structural steel products.

Truscon Type "G-W" Radio Towers are of particular interest to the buyer who is desirous of obtaining the utmost in antenna quality and strength at a cost representing only a nominal premium above the cost of secondary types of construction. These guyed towers are available in shop-welded unit lengths for tower heights up to 528 feet, and include these features:

TRIANGULAR—because this design resists distortion with greater efficiency than any other form and is so recognized by the engineering profession.

UNIFORM IN CROSS SECTION—because radio engineers proclaim this feature a distinct asset in broadcasting.

STRONG—because these towers and all component parts are designed to resist a minimum wind load of 30 lbs. per square foot which is accepted as a design adequate for most geographical areas not subject to frequent cyclone visitation.

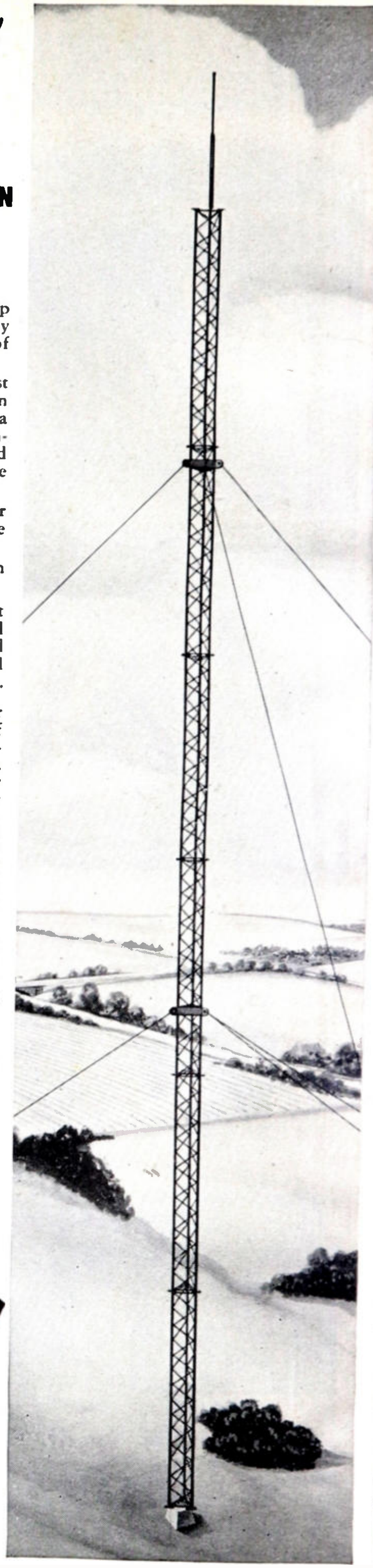
The Type "G-W" guyed tower can be adapted to a number of services. When base and guy insulated, it is an ideal antenna tower. It can also simultaneously support one or more cables or co-axial transmission lines having 3⅛" aggregate diameter and one or more whip-type UHF antennas or a side-mounted FM antenna, with some applications requiring nominal height reduction.

Although the Type "G-W" tower is rated to resist 30 pound per square foot minimum wind pressure, under certain conditions, such as an AM radiator not supporting superstructure for other services, it may be capable of safely resisting 40 or more pounds per square foot of wind pressure.

When non-insulated, the Type "G-W" tower is suitable for a number of services, such as an FM or UHF Antenna Support for Railroads; Public Utilities; Industry; Municipal, County, and State Police; and Communications Networks.

Typical central pier arrangement for non-insulated tower. Other arrangements are possible to meet specific conditions.

Base insulated central pier showing "Mast-Base" or "Pivot" type base insulator with Spark-Gap.



FREE CATALOG

Write for illustrated literature giving complete description, mechanical details, installation photos, and other important information on the new Truscon "G-W" Guyed Radio Towers.

TRUSCON SELF-SUPPORTING AND UNIFORM CROSS SECTION GUYED TOWERS
TRUSCON COPPER MESH GROUND SCREEN

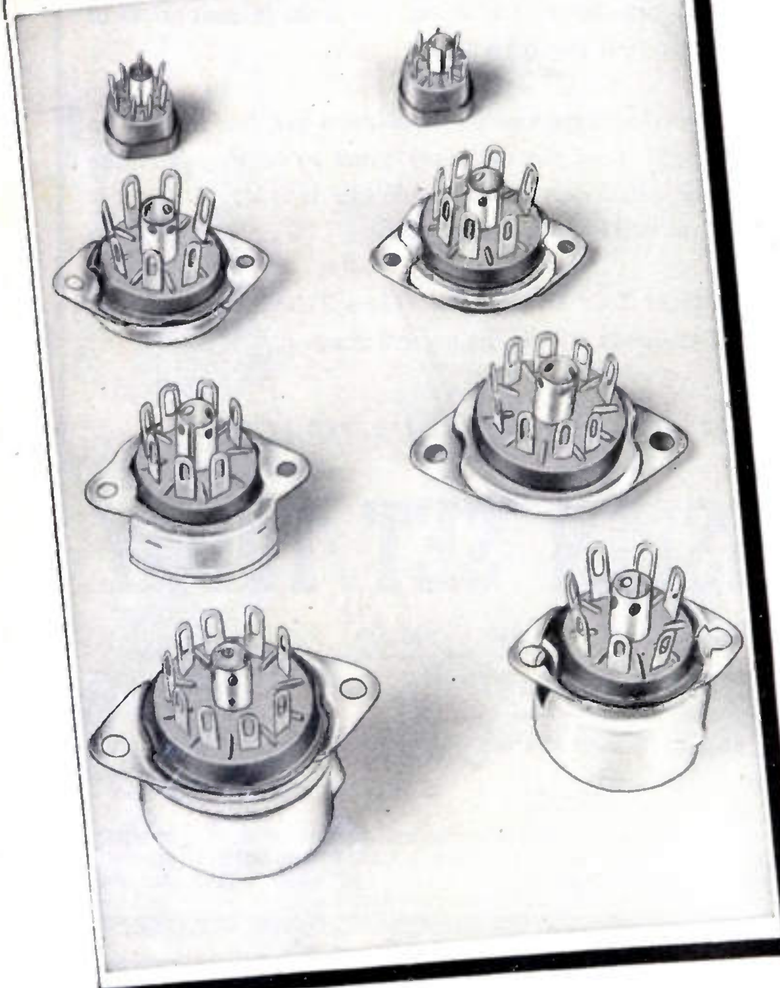
TRUSCON STEEL COMPANY
YOUNGSTOWN 1, OHIO
Subsidiary of Republic Steel Corporation

MYCALEX

MINIATURE TUBE SOCKETS

7-PIN and 9-PIN . . . and SUBMINIATURES

PREMIUM INSULATION PRICED COMPETITIVELY



Now MYCALEX offers both 7-pin and 9-pin miniature tube sockets . . . with superior low loss insulating properties, *at new low prices* that offer ceramic quality for the cost of phenolics.

MYCALEX miniature tube sockets are injection molded with precision that affords uniformity and extremely close tolerances. MYCALEX insulation has high dielectric strength, very low dielectric loss, high arc resistance and great dimensional stability.

Produced in two grades: MYCALEX 410 conforms to Grade L4 specifications, having a loss factor of only .015 at 1 MC. It is priced comparably with mica filled phenolics.

MYCALEX 410X is for applications where low cost of parts is vital. It has a loss factor only one-fourth that of "everyday" quality insulating materials, and a cost no greater.

Prices gladly quoted on your specific requirements. Samples and data sheets by return mail. Our engineers will cooperate in solving your problems of design and cost.

Mycalex Tube Socket Corporation

"Under Exclusive License of Mycalex Corporation of America"

30 Rockefeller Plaza, New York 20, N.Y.

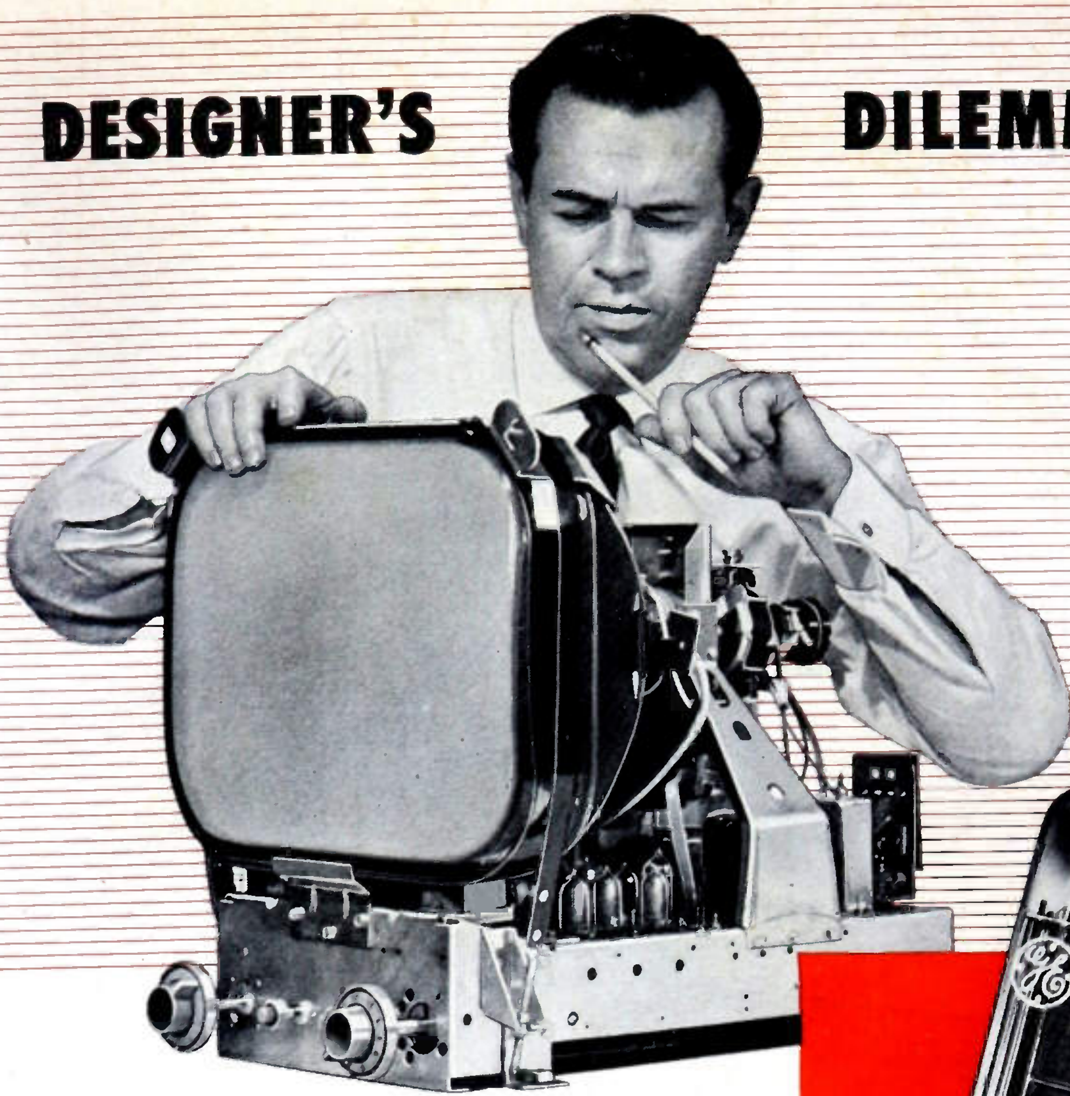
SINCE 1919 MYCALEX THE INSULATOR TRADE MARK REG. U.S. PAT. OFF.

MYCALEX CORP. OF AMERICA

"Owners of *'MYCALEX' Patents*"

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

DESIGNER'S DILEMMA



"How can I save, in order to trim the price of this set another $2.50?"

THE answer, Mr. TV-set Designer, is simple: *G.E.'s 6BN6 gated-beam tube.* It replaces three tubes and associated components, serving as a combined limiter, discriminator, and audio-amplifier.

6BN6 cost is right in line with other receiving types. You get three tubes' performance, yet you pay for only one!

Ask for Bulletin ET-B28, which tells the full story of this amazing G-E *economy* tube, also charts its performance. Or if you prefer to discuss the 6BN6 in person, an experienced G-E tube engineer gladly will call on you. Wire or write *Section 12, Electronics Department, General Electric Company, Schenectady 5, New York.*

6BN6

GATED-BEAM MINIATURE

Typical Operating Conditions, TV Application, 4.5 mc

Plate supply voltage	270 v
Plate load resistance	.33 megohms
Accelerator voltage	100 v
Cathode resistance	200 to 400 ohms
Min signal voltage for limiting action	1.25 v RMS
Audio output voltage	12.5 v RMS
AM rejection, with 2-v input signal	25 db

GENERAL ELECTRIC

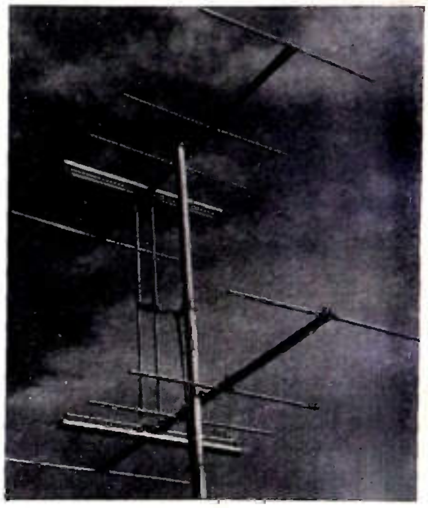
181-K1

News—New Products

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

5-Element Yagi For Weak Signal Areas

A new 5-element Yagi antenna tuned for any one of the low-band (2–6) or high-band (7–13) channels is announced by **Technical Appliance Corp.,** Sherburne, N. Y., manufacturers of antenna systems.



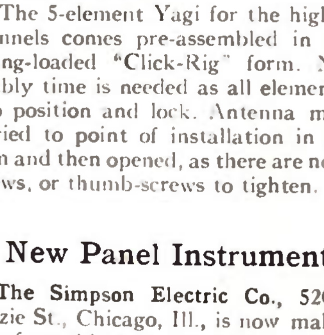
Extremely high front-to-back ratio and pinpoint directivity inherent in the Yagi design minimize ghost effects caused by reflected signals. It delivers 11-db gain by actual field measurement.

The 5-element Yagi for low-band channels is assembled by means of the "Jiffy-Rig" type of construction saving installation time and costly call-backs due to mechanical failures.

The 5-element Yagi for the high-band channels comes pre-assembled in a new spring-loaded "Click-Rig" form. No assembly time is needed as all elements flip into position and lock. Antenna may be carried to point of installation in folded form and then opened, as there are no nuts, screws, or thumb-screws to tighten.

New Panel Instruments

The Simpson Electric Co., 5200 W. Kinzie St., Chicago, Ill., is now making a line of panel instruments in three different



(Continued on page 60A)


PACKAGED Functional Electronics
Components for Control Systems
mecha* products
Servomechanisms Inc. designs, engineers and manufactures electronic control devices for military and commercial applications.
These devices are composed of a number of functional electronic units, which are miniaturized and packaged as plug-in components.
Control equipment is readily synthesized from these small component packages which solve the important system requirements of:
• Spatial adaptability
• Instant maintainability
• Training simplicity
• Ease of manufacture
SERVOmechaNISMS inc.*
Old Country & Glen Cove Roads, Mineola, N. Y.
Imperial Hway. & Sepulveda Blvd., Los Angeles, Calif.
*Trademark

Another RCA FIRST!

. . . the RCA-17CP4 metal shell rectangular kinescope

As pioneers in the development of the kinescope, RCA leads again with a new and advanced type of metal-shell *rectangular* kinescope . . . destined to become the industry's leading large-picture tube. The new RCA-17CP4 has a picture area of 14⅜" x 11", and offers designers the following notable advantages . . .

Use of the metal shell not only makes practical a construction which weighs less than a similar all-glass tube, but also makes practical the use of a higher-quality face plate than is commonly used on all-glass tubes.

The rectangular shape, which allows reproduction of the transmitted picture without waste of screen area, permits use of a cabinet having about 20 per cent less height than is required for a round-face tube having the same picture width. In addition, the chassis need not be depressed or cut out under the face of the tube and controls can be located as desired beneath the tube.

The 17CP4 with its design-center maximum anode-voltage rating of 16 kilovolts, provides pictures having high brightness and good uniformity of focus over the whole picture area. It has a high-efficiency, white fluorescent screen on a relatively flat, high-quality faceplate made of frosted Filterglass to prevent reflection of bright objects in the room and to provide increased picture contrast.

Employing magnetic focus and magnetic deflection, the 17CP4 features an improved design of funnel-to-neck section which facilitates centering of the yoke on the neck and, in combination with better centering of the beam inside the neck, contributes to the good uniformity of focus over the entire picture area. The diagonal deflection angle is 70° and the horizontal deflection angle is 66°.

Other features incorporated in the 17CP4 are short overall length and an ion-trap gun which requires only a single-field, external magnet.

RCA Application Engineers are ready to co-operate with you in applying the 17CP4 and associated components to your specific designs. For further information write RCA, Commercial Engineering, Section 47AR, Harrison, N. J.

Another RCA-developed tube

Designed for Radiosonde Service, the RCA-5794 employs two resonators integral with the tube. The output resonator is tuned to 1680 Mc by means of an adjusting screw. Useful power output is 500 milliwatts.

The Fountainhead of Modern Tube Development is RCA

RADIO CORPORATION of AMERICA
ELECTRON TUBES
HARRISON, N. J.

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Ivan S. Coggeshall

PRESIDENT, 1951

Ivan S. Coggeshall, general traffic manager of Western Union Telegraph Company's overseas communications, was born on September 30, 1896, in Newport, R. I. In 1911 he held one of the early wireless amateur licenses issued by the Government, and in 1912, he became a telegraph operator for the Postal Telegraph Cable Company.

Following a three-year course of study in electrical engineering at Worcester Polytechnic Institute, Mr. Coggeshall joined the Western Union Telegraph Company in 1917 as an engineering apprentice at Boston, Mass. He rose successively to engineering assistant in 1920, general inspector in 1922, general traffic supervisor in 1927, staff assistant on submarine cable matters to the vice-president in charge of traffic in 1936, and was appointed general traffic manager in 1946.

For the past 14 years Mr. Coggeshall has specialized in ocean cables, and much of his work has been devoted to the adoption of electronic method and devices in the telegraph and submarine cable field. The most recent of these devices is a traffic capacity-doubling amplifier nov being inserted in transatlantic cables and sunk on th ocean bottom. He was on active duty with the Unite States Navy in both World Wars. During the secon World War he represented his company on the cabl committee of the Board of War Communications.

Mr. Coggeshall joined the Institute as an Associat in 1926, transferred to Member in 1929, and in 1942 re ceived the Fellow award for his services to the Institut and to the engineering profession. He served as a Direc tor of the Institute from 1941 to 1944, and has been member of the following IRE committees: Annual Re view, Building Fund, Constitution and Laws, Execu tive, Finance, Membership, National Convention, Nev York Program, Nominations, Papers, Papers Procure ment, Papers Review, Professional Groups, Publi Relations, Sections, and Standards.

Mr. Coggeshall is a member of Tau Beta Pi and th American Institute of Electrical Engineers. He is registered professional engineer in New York State.

Symbol of the IRE

AT THE time that The Institute of Radio Engineers was formed, it was without an emblem or symbol. It had pride of ancestry since its predecessor and component societies—The Society of the Wireless Telegraph Engineers, of Boston, and The Wireless Institute, of New York—were active and enterprising organizations. But neither had an emblem which seemed readily adaptable for the newly born Institute of Radio Engineers.

The selection of a symbol for a learned society is not an altogether simple matter. There is the temptation to use something that would be readily understood by everyone active in the field of that society at the moment. Had this rather easy method been adopted, undoubtedly the Institute would have been saddled for many years with an emblem consisting of a spark gap functioning in the center of a dipole, surrounded by a circular resonator provided with a micrometer gap for reception. Fortunately the temptation to use a "current-events" emblem was sternly resisted.

And it was well that this was so. Since those early days sparks, arcs, quenched sparks, high-frequency alternators, and a multitude of types of vacuum-tube oscillators have in succession taken the center of the radio communications scene of action. Had any specific symbol of the spark-gap type been adopted, the Institute would have been forced to utilize an obsolete emblem or, alternatively, to change its symbol every few years.

With these inconveniences in view, it was decided to devise a far more general and, in a sense, perpetual symbol. It was realized that The Institute of Radio Engineers would always deal with electrical communications—that is, with electrical energy guided by conductors or passing through space. In either case, the distinguishing character of the transmission process was the existence of electrical forces and of their correlative magnetic forces. Such forces exist both in wire transmission and in space transmission using the usual electromagnetic waves.

Fortunately enough, as the scope of the industry gradually expanded and finally covered the field of electronics, it became clear that the ubiquitous electrical and magnetic forces were the omnipresent factors in that field as well. It was therefore wise on an immediate as well as a long-term basis to select a representation of electrical and magnetic forces as the emblem of the IRE.

Accordingly the electrical force was represented by a vertical arrow, and the magnetic force by a circular arrow surrounding the electric force and in the conventional relationship to it. The shape of the resulting drawing obviously lent itself to a triangular placement of the letters I.R.E. This in turn led to the selection of an attractive triangular emblem.

The symbol of the Institute seems to have evoked little or no comment, whether favorable or adverse, during the decades following its adoption. It might therefore be assumed that the membership is reasonably well satisfied with that symbol.

More important than the symbol itself is the Institute for which it stands. From a few tens of members, the Institute has grown to tens of thousands of members. The importance of the contributions of its membership to the advancement of the communications and electronics field is outstanding and unchallenged. The dignity and constructiveness of its operations are obvious. Its publications have included substantially all major advances in the field of the Institute. The standards originated in the I.R.E. and now published in its PROCEEDINGS have greatly advanced the art. Its Conventions, Section meetings, Regional Conventions, and other forms of assemblage, as well as the operations of its Professional Groups, show a long record of accomplishment and promise of still further advances.

Most hopeful of all is the fact that the Institute and its symbol represent a progressive and evolving situation, and not a mere static and fixed condition. The activities of the Institute, whether carried out by its Technical Committees, its meetings, or its publications, are always subject to critical scrutiny, to change, and to improvement. It is a privilege to be associated with others in a professional society justly proud of its past accomplishments and energetically determined far to surpass these in the future.—*The Editor.*

Video Utilization—An Introduction and Definition of the Art*

R. L. GARMAN†, ASSOCIATE, IRE

Summary—**Some of the technical and artistic problems relating to video recording have been introduced and defined for further study and analysis.**

THE 23.2 Subcommittee on Video Utilization, as set up by the Chairman of the Video Techniques Committee, has the assignment of devising methods of measurement and test for the video utilization techniques which are currently used by the major telecasters of this country. The phase of video utilization which now occupies the major share of attention is video recording. While the scope of the subcommittee is not necessarily limited to this phase, the subject has attracted the interest of a majority of the members and has, therefore, become a matter of prime consideration.

Video recording is both a science and an art, since it involves the presentation of entertainment material through the medium of photographic film. As a science, video recording is as old as television itself. The engineers working in this field from its inception had to devise some method for retaining permanent records of their results. For the most part, however, a still picture of several frames, a single frame, or even of a single field, was sufficient. The art phase was unimportant, since the subject matter usually consisted of charts and diagrams which were fitted to the immediate apparatus at hand. A glossy wood figurine of Felix the Cat was a familiar subject for early recording. Today, however, video recording is expected to convey faithfully all of the program content of a studio performance, including all of the shades and moods which a skilled director can create by various lighting and camera effects. The final product is, therefore, a motion picture in almost every sense of the word, although it does not necessarily have to be recorded in accordance with the physical standards of the motion picture industry.

For commercial reasons, most of the recording today is done on 16-mm film, at the motion picture standard rate of 24 frames per second. A frame rate conversion to television standards must therefore be effected, resulting in some loss of television picture material, as well as the possibility that evidence of this conversion might appear in the final product. It can be appreciated that this conversion presents technical difficulties for the designer of the recording instrument, and also complicates both the nature and extent of the methods of measurement and test which must be devised in the field of video recording. Since the motion picture frame rate is accepted in present practice, the product can be used for direct exhibition to large or small audiences, for retransmission on the same or another television station or network, or for other special purposes peculiar to either the motion picture or television industry. The end use does not seriously influence the methods of measurement and test, but nevertheless consideration must be given in detail to the various techniques involved in picture taking and processing so that commercial standards for the end use of the product will be satisfied. The final product must, in any case, be of the highest quality, regardless of the application.

Video recording incorporates the combined contributions of electronic, mechanical, and photographic arts. The subject not only creates new problems in all of these fields, but also places a new emphasis on existing problems. Some of these have perhaps never before received the full degree of consideration which now becomes necessary before measurement and test methods can be devised. To permit orderly analysis, the work of the subcommittee has been directed along three main subject divisions. The main subject divisions are: (a) electrical constants of the process, (b) film constants of the process, and (c) physical constants and parameters of the equipment used. Each of these may, in turn, be subdivided into various parameters for the major and minor components used in video recording systems. Some of the topics which require consideration are discussed in the following paragraphs.

Since the object photographed is a cathode-ray tube, it is necessary to outline special properties for this unit. Video recording places stringent demands on the cathode-ray tube and its associated electrical circuits. Good contrast, good resolution, small change of spot size with change in intensity, and good focus all over the face of the tube are only a few of the requirements. Special attention must also be given to linearity of scan, especially if the subject material will be used for retransmission, where four scanning operations are involved. Since, in each scanning operation, nonlinearity may be expected in the same part of the picture, any cumulative defect could well lead to disastrous results. Steadiness of interlace is of great importance, not only in the usual sense of providing maximum resolution, but also in a new sense. Unsteadiness introduces defects in the splicing of consecutive television fields, and such splicing is inevitably necessary when converting from television to motion picture frame rate standards. The recording cathode-ray tube and its associated circuits are, however, only a means toward an end. The photog-

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† General Precision Laboratory, Inc., Pleasantville, N. Y.

raphy at the cathode-ray tube is only one step in the process, but standards must be determined as though the photography were actually from the studio scene itself. In this respect, much of the work may be beyond the scope of the subcommittee.

The physical and chemical constants of the film play a very vital part in video recording, and must therefore be understood by the technicians and engineers working in this field. Some of these constants are: the resolution of the film, the speed of the film, the spectral sensitivity, and the nature and extent of the toe and shoulder of the *h* and *d* curve. The processing of the film, including the initial exposure and the conditions under which it was made, must be understood. Such factors as diffuse density, specular density, and gamma must be considered in relation to the final product. The differences in gross contrast and detailed contrast for both the film and the cathode-ray tube must be properly evaluated and interrelated if proper results are to be obtained. The gamma of the picture must be controlled in order to produce a film in which the final density and gamma are correct for the end use. This gamma may of course be controlled during several stages of the process, i.e., in the original scene, in the pickup camera, in the various line amplifiers or possibly in a special line amplifier, in the film processing, or in the retransmission apparatus. Moreover, processing considerations are complicated by the fact that at some point a sound track must be added, either in the original recording or in a later part of the process, and it may therefore be necessary to effect compromises of picture quality, as well as sound quality in the final commercial product. Since 16-mm film is in common commercial use today, this last factor and some of the preceding factors are doubly important.

The conversion from 30 frames per second to 24 frames per second may be accomplished by either a mechanical shutter or an electronic shutter, or by some combination of both. Each method presents its own problems and produces certain obvious difficulties for which standards of measurement and test must be devised. The recording cathode-ray tube may present either a negative or positive picture, depending upon whether the user desires a direct negative or a direct positive as the product produced by his equipment. Testing procedures for these different pictures may vary considerably due to the nonlinear properties of the transducers and the film used in present-day equipment.

The problem of vibration, which is important in the design of an ordinary motion picture camera, assumes really major proportions in the design of a video recording camera. In motion picture photography, when vibration occurs during a small portion of the exposure interval, only a slight loss of detail is noticed and all parts of the picture are equally affected. This, however, is not the case in video recording. The subject is a moving spot, forming lines which must be interlaced at one rate and spliced together at another rate to effect the desired record. Vibration at either rate causes a very apparent defect. The same may be said about film motion at the beginning and end of the exposure cycle. Other physical factors which are important are the size of the picture in relation to motion picture camera aperture size, drift of the cathode-ray tube pattern, and preservation of the aspect ratio, especially under the conditions where there is some nonlinearity of scan of the horizontal and vertical sweeps.

The foregoing discussion, covering only a partial statement of the new problems and corresponding new methods which can be anticipated in this field, illustrates the magnitude of the problem encountered in devising methods of measurement and test. Considerable advance preparation must therefore be completed before standards can be proposed and implemented for use by the industry. The immediate tasks which face the subcommittee are consolidation of all subject material, and organization of this material under the three main subject headings which have been adopted. Some of the material must be assembled from fields which are related to electronics, but which have not heretofore been extensively treated in the literature.

In this new and somewhat fluid field, the task of devising methods of measurement and test is indeed a huge one. Additionally, the practicability of any test devised as a result of this work must be demonstrated before these methods can be stated with authority. Since the task cannot be completed at an early date, a recommendation has been made that the findings of the subcommittee be presented to the Institute members in the form of a series of papers, the first of which appears in this issue. Other papers which will be presented to the Editorial Board of IRE for consideration for publication in forthcoming issues are:

- Some Film Characteristics in Video Recording
- Calibration Methods in Video Recording
- The Characteristics of Cathode-Ray Tubes in Relation to Video Recording
- Some Dimensional Aspects of the Apparatus Used in Video Recording.

These papers are designed to lay the groundwork in the video recording field by covering the present state of development of the various techniques. They will be written, for the most part, in the language of the practical engineer, and will give specific instructions and references where possible.

Some Comparative Factors of Picture Resolution in Television and Film Industries*

H. J. SCHLAFLY†, SENIOR MEMBER, IRE

Summary—**This paper reviews and compares the quantitative meaning of the term "resolution" as commonly used by the television industry and the film industry. The danger of using values of limiting resolution as the sole measure of picture quality is discussed. Conversion equations are developed and a table listing numerically equivalent values of resolution is provided.**

Introduction

THE MERGER of electronics and photography into the corporate function of television recording has resulted in a unique situation. It is a situation which is logical and natural, but which, nevertheless, has caused misunderstandings, delays, and even exasperation. The problem is simply one wherein two sciences that have hitherto been comparatively independent of each other suddenly find that they define and describe certain phenomena in terms which are not identical, but which are similar enough to be thoroughly confusing.

The ultimate objective of both television and photography is the faithful reproduction of an original scene. But while the beginning and end products are the same, the medium and methods are widely different. Thus, it is little wonder that there are few, if any, existing experts who are so thoroughly familiar with the terminology and techniques of both sciences that they can point out in advance the areas of confusion or misunderstanding. This paper will attempt to deal with only one "area of confusion," the meaning of picture resolution as defined by terminology in current use.

General

The resolving power of a medium or a device is a measure of the ability of that device to convert, transmit, or reproduce details of the original scene. Detail, of course, is a "separately considered particular,"[1] which contributes to or is part of the whole. A device which is capable of handling more or finer detail is said to have the greater resolving power and the resulting picture has more resolution. Lack of picture resolution not only results in the subordination or complete loss of parts of the original, but also in a loss of "edge sharpness" which gives the picture a "soft" or, more correctly, a diffuse quality. The accepted method of determining resolution is to provide a scale or chart having calibrated points or steps of increasing fineness of detail and to determine the point at which the device under test breaks down in the performance of its function. The human eye itself has a certain resolving capability which is influenced by the portion of the retina being used, the spectral content of the light and the absolute value of the light energy, as well as by the optical characteristics of the lens. Each technical device which precedes the seeing process of the eye has its own resolution characteristic and contributes its part to the degradation of the original scene.

In general, the deterioration contributed by any physical device is evidenced by a gradual reduction in contrast ratio with increasing detail until a point is reached where there is no distinction between two adjacent points which did have some quality of distinction in the original. Whether this contrast ratio is measured in light energy, grains of silver deposit per area, potential difference, or whatever, is immaterial. A notable exception to this gradual deterioration of resolution is the "sharp cutoff" voltage amplifier which might maintain constant amplification with increasing frequency (detail) until a certain critical or cutoff point is reached, and thereafter drop sharply towards zero output.

Comparative physical sizes play a large part in determining the point where "signal attenuation" begins to occur. Thus, so-called "aperture size," or the area within which there can be no differentiation—such as a single nerve ending in the eye, the focused scanning spot in a cathode-ray device, or the grain size in a photographic emulsion—is a major contributor to the limitation of resolution. But there are many other contributing causes which do not necessarily deal with physical size, such as electrical time constants; aberrations in optical devices; phase shift in amplifiers; spectral sensitivity of emulsions, photocathodes, and lenses; and, unfortunately, others.

Today both the photographic and television industries speak of the absolute limit of resolution as a measure of picture quality. Actually the evaluation of quality is so complex that measurement of one of the contributing factors is not adequate to describe the end result. Much work has been done and is being done to determine all of the factors involved.[2,3] In particular, analytical attention is being given to detail contrast ratio, random noise, brightness, and tone reproduction, as well as to limiting resolution. The paragraphs which follow deal only with definitions and conversion factors

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† Twentieth Century-Fox Film Corporation, New York, N. Y.

[1] Funk and Wagnalls *New College Standard Dictionary*.

[2] M. W. Baldwin, Jr., "The subjective sharpness of simulated television images," *Bell Sys. Tech. Jour.*, vol. 19, p. 563; October, 1940.

[3] Otto H. Schade, "Electro optical characteristics of television systems," *RCA Rev.*, vol. 9, p. 5, p. 245, p. 491, p. 653; March, June, September, and December, 1948.

for the resolution terminology in current use. The paper should definitely not be considered as a treatment of the measure of picture quality.

Terms

One is likely to assume that the use of the common term "lines" permits a basis for comparison between photography and television picture resolution. Such is not the case. Each industry has independently arrived at a definition in language best suited to its own measurement technique and, as a result, numerical values which are not apparently related might refer to the same degree of "absolute" resolution in a television picture and in a photograph.

The *film industry* defines resolution in terms of lines per millimeter of film surface. Typical test charts are provided by the National Bureau of Standards (shown in Fig. 1) and by the American Standards Association. Such charts usually consist of a series of blocks or squares of parallel black lines separated by clear spaces of the same width. Each block represents a given number of black lines per millimeter of film surface when the chart is photographically reproduced on the film emulsion. For determining resolution values given in film specification sheets, the contrast ratio between the

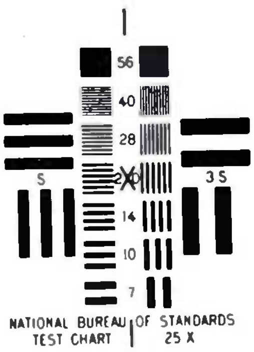

Fig. 1—National Bureau of Standards test chart.

black lines and the clear spaces on the original chart is held at 30:1. Incidentally, this is about the highest value which can be obtained on a printed chart. Transmission type charts, used in some resolution measurements, can provide contrast ratios of 100:1 or 1000:1.

The resolving power of a given film emulsion is determined by photographing a test chart using the optimum exposure, processing the film by recommended methods and examining the image under a microscope. The maximum number of black lines per millimeter just resolved, not lost as an indistinguishable grey mass, is the value used to indicate the resolving power of that particular film. In practice, the resolving power values of commercial films varies from about 55 lines per millimeter for negative film to as high as 150 lines per millimeter for fine grain sound recording films.

Of course, the figures given in the above paragraph do not necessarily represent the end product of film resolution as seen on the screen of a motion picture theater. In February, 1946, a portion of the Television Committee of the Society of Motion Picture Engineers made observations of screen resolution of a special test film projected in a group of leading New York theaters. These data were not published because the tests were not sufficiently extensive to permit definite conclusions. In the words of the Committee report: "The influence of many individual factors has not been determined, but it is believed that the results . . . are broadly representative of present motion picture practice . . . ." The conclusion reached in the same report stated: "In general, it can be concluded from theater projection of the two test films specially prepared for the use of this Committee that projection in first run theaters shows resolution of 28 lines per millimeter on 35-mm film where the test object includes pictorial subject matter, and 40 lines per millimeter where the test card alone was photographed."

In the *television industry* picture resolution is usually measured with the aid of a test pattern, such as the RMA Resolution Chart 1946. This chart follows the practice of using horizontal and vertical wedges, rather than a series of parallel lines. The pattern is composed of a given number of alternate black-and-white lines of equal width which continuously converge from the wide to the narrow end of the wedge. Thus, the chart is provided with a continuously variable resolution pattern, numerically calibrated by indexing various points along the wedge. Each black and each white line is counted as an individual line, whereas in the film industry each black line only is counted as an individual line.

The resolution of the television picture is indicated by a value which represents the limiting number of black-and-white lines identifiable as such, not lost in an indistinguishable grey, in a vertical or a horizontal dimension equivalent to the picture height. For the purpose of assigning this value, it is assumed that the resolution of any and every point in the picture is equal to that observed at the wedge. Such an assumption is, of course, not true, but it is a convention which provides a numerical value of resolution accepted throughout the industry. Degradation at the corners of the picture sometimes is identified by the term "corner resolution" and is evaluated by the same process—interpreting the resolution of a wedge located in the corner in terms of the full dimension of picture height.

It is a common error to confuse the number of horizontal scanning lines as set by the television standards with the figure for picture resolution. The television standards in this country specify 525 horizontal scanning lines per picture frame. Only 92 to 95 per cent of these are active scanning lines, the remainder being blanked out during the vertical sweep retrace. But even the remaining 480 some odd lines do not specify the limit of vertical resolution. There is an additional loss in vertical resolution inherent in the television dissecting process which provides a second factor, even when there is perfect interlace of the alternate scanning fields. This effect is illustrated by integration of those portions of black-and-white resolution lines within the width of the scanning line and point-by-point comparison of the resulting half-tone with the original.[4] Fig. 2 plots such information for a range of scanning line factors, showing the percentage of possible scanning-line-resolution-line phasing for which the integrated half-tone will be within 20 per cent of the original. Choice of a scanning line factor may be a matter of individual preference, but a value of 0.75 is commonly accepted and is the value used in the equation included in this paper.

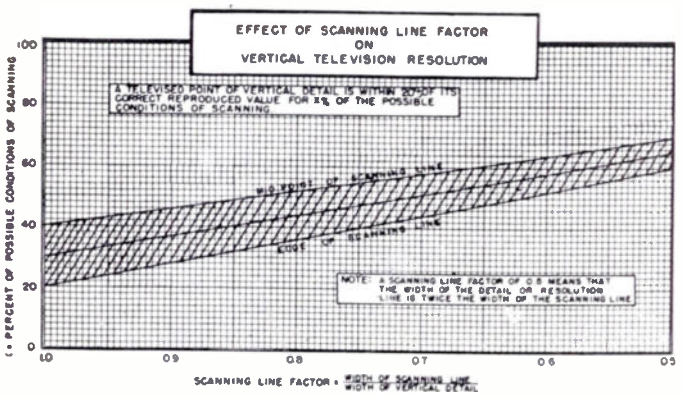

Fig. 2—Integration effect of scanning line width.

Using the above factors, present-day standards impose a limitation on vertical resolution of the television picture of approximately 360 lines.

Television picture resolution, by virtue of common usage among electronic personnel, has also come to be identified in terms of bandpass, or maximum pass frequency of the video circuits. Such usage has meaning only when applied to horizontal resolution, and then only if a definite horizontal scanning period is specified. One cycle of video during active horizontal scanning represents one dark and one light picture element on a particular scanning line. The higher the video frequency, the greater number of picture elements that can be theoretically squeezed into one line. Ideally the one cycle which supplies the light and the dark picture element should contain sufficient harmonics to resemble a square corner pulse; actually, a sinusoidal wave form is considered sufficient for the limiting condition, sacrificing "edge sharpness" between picture elements. It will be realized that a longer scanning period would permit more cycles of video signal to be included in one scanning line, and thus the value of horizontal resolution would be increased. The scanning period is set by the horizontal scanning frequency, or, by the combination of picture frames per second and scanning lines per frame. Fig. 3 indicates the relationship between video bandpass and horizontal resolution for several values of scanning lines per frame.

It is interesting to note that the video bandpass of 4.5 Mc, the nominal television broadcast standard, results in a horizontal resolution of approximately 360 lines. Thus, it is seen that the present standards provide about the same picture resolution in the vertical and horizontal co-ordinates.

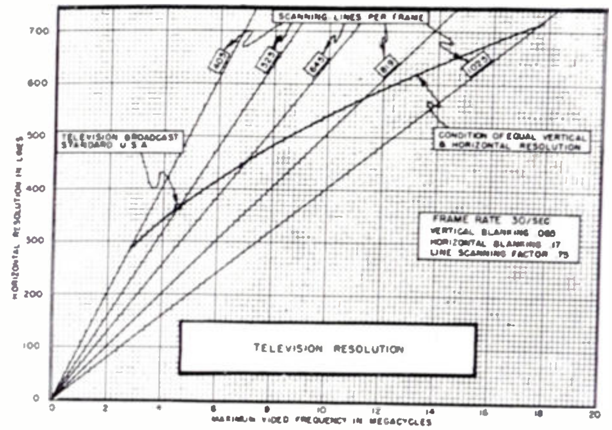

Fig. 3—Television resolution, frame rate 30/second, vertical blanking, 0.065; horizontal blanking, 0.17; line scanning factor, 0.75.

Conversion Factors

A. Conversion of Film Resolution in Lines per Millimeter to Television Resolution in Lines

$$R_t = 2H_f R_f$$

where

R_t = television resolution in lines per picture height
R_f = film resolution in lines per millimeter
H_f = height of standard motion picture projector aperture in millimeters.

For 35-mm sound film

$$H_f = 15.25 \text{ mm}$$
$$R_t = 30.5 R_f.$$

For 16-mm film

$$H_f = 7.21 \text{ mm}$$
$$R_t = 14.42 R_f.$$

B. Conversion of Television Scanning Lines Per Frame to Lines of Vertical Resolution (Television)

$$R_{tv} = b(aL)$$

where

R_{tv} = vertical resolution (television) in lines
a = vertical blanking factor
b = line scanning factor
L = total number of scanning lines per television frame.

Substituting present Standards:

$a = 0.92$ min., 0.95 max., 0.935 average
$b = 0.75$ (representative)
$L = 525$ lines
$R_{tv} = 0.701L = 0.701 \times 525 = 368$ lines.

[4] A detailed analysis of the effect of finite scanning apertures can be found in P. Mertz and F. Gray, "Theory of scanning and its relation to the characteristics of the transmitted signal in telephotography and television," *Bell. Sys. Tech. Jour.*, vol. 8, p. 464; July, 1934.

C. Conversion of Maximum Video Pass Frequency to Lines of Horizontal Resolution (Television)

$$R_{th} = 2 f_{max} T_h / A$$

where

R_{th} = horizontal resolution (television) in lines

f_{max} = maximum video pass frequency in megacycles

T_h = active time (unblanked) of horizontal sweep in microseconds

$$= C \frac{1}{L(F_r)} 10^6 \ \mu\text{sec}$$

where

C = horizontal blanking factor

F_r = frames per second (television)

A = television aspect ratio.

Substituting present standards:

$$f_{max} = 4.5 \text{ megacycles}$$

$$C = 0.82 \text{ min., } 0.84 \text{ max., } 0.83 \text{ average}$$

$$F_r = 30 \text{ frames per second}$$

$$T_h = 0.83 \frac{10^6}{525 \times 30} = 52.7 \text{ microseconds}$$

$$A = 4/3.$$

Then

$R_{th} = 79 f_{max} = 356$ lines.

D. General Conversion Formulas for Equal Resolving Power Between Film and Television

1. Television scanning lines per frame in terms of film resolution (required for equal vertical resolution)

$$L = (2/ab) H_f \times R_f = \frac{2}{0.935 \times 0.75} H_f \times R_f$$

$$L = 43.5 R_f \quad \text{for} \quad 35\text{-mm film}$$

$$L = 20.6 R_f \quad \text{for} \quad 16\text{-mm film.}$$

2. Maximum video frequency in terms of film resolution (required for equal horizontal resolution), in 525-line, 30-frame television system.

$$f_{max} = \left(\frac{A}{C} \times L \times F_r \times H_f \times 10^{-6}\right) R_f$$

$$= \left(\frac{4}{3 \times 0.83} \times 525 \times 30 \times 10^{-6} \times H_f\right) R_f$$

$$f_{max} = 0.386 R_f \text{ megacycles for 35-mm film}$$

$$= 0.182 R_f \text{ megacycles for 16-mm film.}$$

3. Maximum video frequency in terms of film resolution (required for equal horizontal resolution in a 30 frame television picture), if the number of scanning lines in that picture has been chosen to give equal vertical resolution

$$f_{max} = \left(\frac{A}{CF_r}\left(\frac{2}{ab} H_f\right) H_f \times 10^{-6}\right) R_f^2$$

$$= \left(\frac{2A}{abC} F_r \times H_f^2 \times 10^{-6}\right) R_f^2$$

$$f_{max} = 0.032 R_f^2 \text{ megacycles for 35-mm film}$$

$$= 0.00715 R_f^2 \text{ megacycles for 16-mm film.}$$

The above equations have been applied to several values of film resolution for both 35-mm and 16-mm sound film and the results have been tabulated in Table I and Table II, respectively, in the Appendix. These tables list numerical equivalent values of resolution and the corresponding television standards which would be necessary to realize such a value first, of horizontal resolution (with 525 scanning lines per television frame, 30 frames per second) and second, of both vertical and horizontal resolution (with 30 frames per second).

These tables could be interpreted to say that, provided *all other factors affecting picture quality are equal*, a television picture having a limiting resolution of 360 lines (the approximate capabilities of the existing television broadcast standard in the United States) is equivalent to a 35-mm sound motion picture film having a limiting resolution of about 12 lines per mm; or, to a 16-mm sound motion picture film having a resolution of about 25 lines per mm. In actual practice film resolution having a limiting value of 30 to 40 lines per mm is not difficult to achieve. On the other hand, the reproduced film picture is not able to maintain the contrast ratio that can be realized in a reproduced television picture as detail approaches the limit of resolution. Some workers in the field believe that the "other factors affecting picture quality" mentioned above may eventually be so improved in the television system that existing standards will permit a television picture quality closely approximating that of the present-day 35-mm motion picture film in spite of wide differences in the limiting value of picture resolution.

It must be emphasized again that the tables in the Appendix provide numerically equivalent values of resolution. They do not in themselves permit a comparison of picture quality. They in no way indicate the film resolution that is required when a film is to be reproduced over a television system or when a television picture is to be reproduced on film. It is obvious that when a film is reproduced by a television system, or vice versa, the end result will contain the defects of both. For best results, therefore, both systems should be operated as close as possible to their limit of perfection, or, in some cases, be controlled to compensate for defects or limitations of the other.[5]

Summation

Picture quality and picture resolution are not necessarily synonymous. A figure indicating picture resolution is generally a numerical measure of the limit of

[5] Charles L. Townsend, "Specifications for motion picture films intended for television transmission," *Jour. Soc. Mot. Pict. & Telev. Eng.*, vol. 55, pp. 147–157; August, 1950.

detail distinction. Picture quality is a function not only of the limit of detail distinction, but also of the attenuation characteristic which accompanies the reproduction of increasing detail, and numerous other factors of reproduction.

The film industry speaks of resolution as a figure indicating the maximum number of black lines, separated by white spaces of equal width, which can be identified in a dimension equal to one millimeter of film surface.

TABLE I

35-mm Sound Film

Numerically Equivalent Values of Resolution		Minimum Television Standards Required for This Resolution		
Film (lines per mm.)	Television (lines)	Horizontally* (video freq.)	Vertically & Horizontally (lines/frame)	Vertically & Horizontally (video freq.)
90	2740	35 Mc	3900	260 Mc
40	1220	15	1700	51
28	850	11	1200	25
17	520	6.5	750	9.3
11	335	4.2	475	3.9

The television industry speaks of resolution as a figure indicating the maximum number of alternate black-and-white lines of equal width, which can be identified in a dimension equal to the picture height.

Appendix

The following tables list comparative figures obtained by applying the equations given above, for a representative number of resolution values.

TABLE II

16-mm Sound Film

90	1300	16 Mc	1850	58 Mc
40	580	7.3	820	11
28	400	5.1	580	5.6
17	250	3.1	350	2.1
11	160	2.0	230	0.9

* Provided the standard of 525 scanning lines per frame is retained.

Note: When transcribing film to television or television to film degradation factors of each system are cumulative. To minimize over-all degradation, the resolution capabilities of one system should substantially exceed that of the other. The magnitude of this "safety factor" is governed by operational techniques.

Correction of Deflection Defocusing in Cathode-Ray Tubes*

JENNY E. ROSENTHAL†

Summary—**This investigation deals with the problem of minimizing spot distortion by suitable design of the deflecting plates in a cathode-ray tube with electrostatic deflection. The treatment is mathematical throughout. A number of the results obtained may be applicable to a much wider range of problems. A number of special design formulas are given which solve the problem satisfactorily.**

Introduction

The behavior of an electron beam inside a cathode-ray tube is well known experimentally.[1] It is described very briefly here in order to give a clear picture of the mathematical problem studied. For the sake of simplicity, we restrict ourselves to one set of deflecting plates D' and D'' (see Fig. 1). An electrostatic lens F focuses the electron beam so that in the absence of a deflecting potential it forms a circular spot

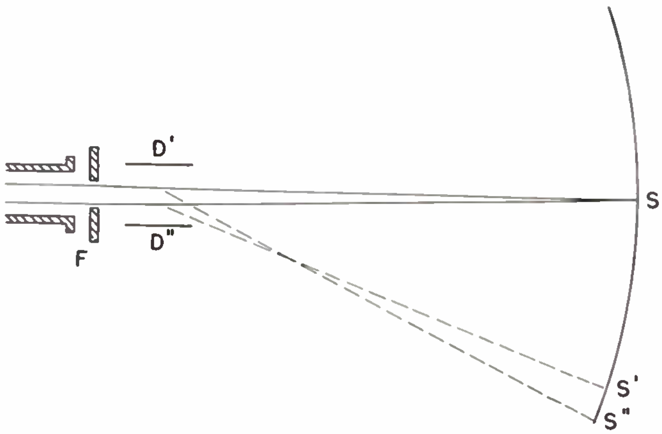

Fig. 1—Schematic illustration of spot distortion showing electrostatic lens F and deflecting plates D' and D''. The solid lines represent the undeflected electron beam while the broken lines show its behavior in the presence of the deflecting field.

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[1] See, for example, J. de Gier and A. P. van Rooy, "Improvements in the construction of cathode ray tubes," *Philips Tech. Rev.*, vol. 9, pp. 180–184; June, 1947.

image S on the screen. If we apply to the plate D' a potential which is negative and to the plate D'' a potential which is positive with respect to the final anode potential, then the electrons on the side D' are retarded while those on the side D'' are accelerated. As a result, the upper portion of the beam in Fig. 1 is deflected more than the lower one. This causes a crossover inside the beam and a subsequent spreading out. The circular spot S is distorted to an oval image $S'S''$. This is known as spot distortion.

It is well known that deflection defocusing may be corrected by suitable circuit design.[2] However, since the method requires extra equipment and is thus too cumbersome to be of much value in practical applications, we present here a method for correcting for spot distortion by suitable plate design exclusively.

General Theory

In the mathematical treatment of the problem the following simplified experimental arrangement is assumed as a first approximation: a monochromatic beam of parallel electrons leaves the electron gun with a velocity corresponding to U volts. The electrons enter a system of two deflecting plates charged to the potentials $-V_0$ and $+V_0$, respectively, with reference to the potential U. The plate system has two planes of symmetry perpendicular to each other. The electron beam is also symmetrical with respect to one of these planes. The problem may thus be treated in two dimensions. No additional restriction is imposed on the shapes of the plates, which in the two-dimensional problem are referred to as the deflecting curves. Space charge is neglected. While the velocity of the electron beam entering the deflecting system depends on the voltage and the shape of the plates as well as on the potential U, this dependence will be taken into account only to a first approximation, all other end effects being disregarded. As shown in a subsequent paper, fringing cannot be determined by standard mathematical techniques even for the simplest case, and the solution given in the literature[3] for fringing in the parallel-plate problem is incorrect.

The following notation is used (see Fig. 2): The x axis is the axis of symmetry. The deflecting-curve system extends from $x=0$ to $x=1$. (The choice of the projection of the deflecting curves on the x axis as unit of length is a matter of convenience.) The electron beam entering the deflecting system is parallel to the x axis. Its bounding rays intersect the y axis at the points A and B, respectively. The width of the beam is $AB=2\sigma$. We shall denote by the superscripts ^{A} and ^{B} all quantities referring to the bounding rays A and B, respectively. The spacing between the deflecting curves is $2g$ at $x=0$ and $2h$ at $x=1$.

[2] W. F. Schreiber, "A dynamic focus for electrostatic CRT," *Sylvania Technologist*, vol. 2, pp. 16–18; April, 1949.
[3] J. H. Jeans, "The Mathematical Theory of Electricity and Magnetism," 5th ed., Cambridge University Press, London, England, pp. 274–275; 1941.

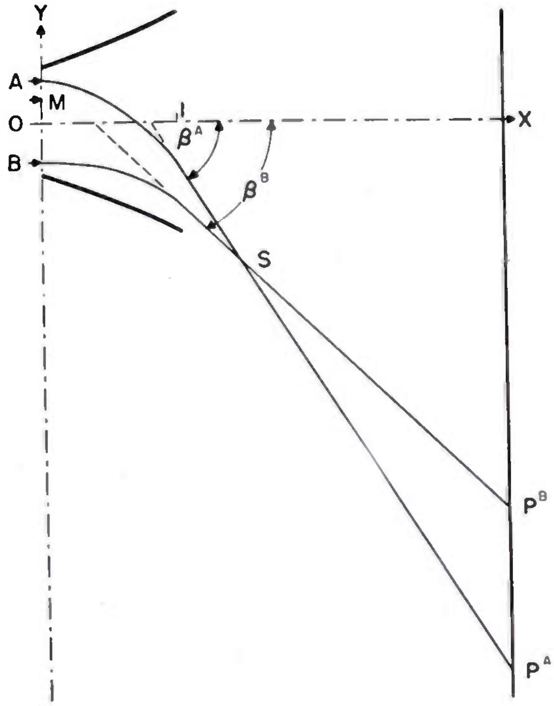

Fig. 2—Diagrammatic cross section of deflecting system in cathode-ray tube illustrating notation.

Before proceeding with the calculations, we shall describe qualitatively the behavior of an electron. Inside the deflecting system it describes a trajectory depending on the shape of the system and the voltage applied. After emerging from the deflecting system, it follows the tangent to its trajectory at the point of emergence. Let β be the angle the tangent makes with the x axis and let the subscript $_k$ denote the value of any quantity at the point of emergence. The equation of the tangent is

$$y = (x - 1)\tan\beta + y_k. \tag{1}$$

If the system were designed so as to eliminate spot distortion, the tangents followed by the bounding electrons A and B would be parallel lines. However, as mentioned in the introductory statement, they intersect in general at a point S with the co-ordinates

$$\begin{aligned} x_S &= 1 + (y_k^A - y_k^B)/(\tan\beta^B - \tan\beta^A) \\ y_S &= y_k^B + (x_S - 1)\tan\beta^B. \end{aligned} \tag{2}$$

After crossing over at S, the rays diverge until they impinge upon the screen at the points P^A and P^B, respectively. It should be pointed out in this connection that though the bounding rays intersect at S, all other rays do not. Nevertheless, the area surrounding S represents the narrowest portion of the deflected beam.

All calculations are made for an electron entering the deflecting system at a point M with ordinate $OM=\eta$; by definition $\eta^A=\sigma$ and $\eta^B=-\sigma$.

The first step toward finding the deflection of the electrons is to determine the electric field at every point, i.e., solve Laplace's equation

$$\frac{\partial^2 V}{\partial x^2} + \frac{\partial^2 V}{\partial y^2} = 0 \tag{3}$$

with the boundary conditions $V = -V_0$ and $V = +V_0$ at the upper and lower curves, respectively. Since no reference was found in the literature to an exact solution of this problem with arbitrarily shaped curves, the following procedure was devised.

The form of the boundary conditions postulates that the solution $V(x, y)$ of (3) must be an odd function of y. It is known that if

$$z = x + iy \quad \text{and} \quad z^* = x - iy,$$

then any analytic function $f(z)$ as well as $f(z^*)$ are solutions of Laplace's equation.[4] If the expansion of $f(z)$ in powers of z involves real constants only, then $f(z^*)$ is the complex conjugate of $f(z)$, $f(z) - f(z^*)$ is an odd function of y and a pure imaginary. If we impose this restriction on $f(z)$, then

$$V(x, y) = \tfrac{1}{2} i V_0 [f(z) - f(z^*)] \tag{4}$$

is a solution of (3) satisfying the boundary conditions

$$V = \mp V_0$$

over the curves

$$f(z) - f(z^*) = \pm 2i; \tag{5}$$

the minus sign in the boundary condition corresponds to the plus sign in (5) and vice versa. The factor $\frac{1}{2}$ in (4) is introduced merely as a matter of convenience in algebraic computation.

A mathematical investigation is in progress to determine how closely a limited portion of an arbitrary continuous curve may be approximated by an equation of type (5). The approximation should be sufficiently close for all practical purposes.

It may be interesting to point out that the equation for the equipotential curve $V = -rV_0(-1 \leq r \leq 1)$ has the form

$$f(z) - f(z^*) = 2ir; \tag{6}$$

its intersection η with the y axis is found by solving

$$f(i\eta) - f(-i\eta) = 2ir. \tag{7}$$

The equations for the lines of force have the form

$$f(z) + f(z^*) = 2K \tag{8}$$

where the constant K is real. Equation (8) is listed only for the sake of completeness.

To find the trajectory of an electron after it enters the deflecting field we have to solve Lagrange's equations of motion

$$m \frac{d^2 x}{dt^2} - e \frac{\partial V(x, y)}{\partial x} = 0,$$

$$m \frac{d^2 y}{dt^2} - e \frac{\partial V(x, y)}{\partial y} = 0. \tag{9}$$

with $V(x, y)$ given by (4).[5]

The initial conditions to be satisfied by the solutions of (9) are found from the following considerations.

If end effects were completely ignored, then the initial velocity of the electron, i.e., its velocity at the instant it enters the deflecting system, would be independent of the deflecting field. The beam would remain monochromatic. However, since this simplification leads to a discontinuity in the initial velocity, we replace it by the following approximation due to Deserno.[6]

The effective potential $\overline{U}$ at any point η on the y axis is given by

$$\overline{U} = U + V(0, \eta) = U + \tfrac{1}{2} i V_0 [f(i\eta) - f(-i\eta)]$$
$$= U - rV_0. \tag{10}$$

The initial velocity at that point is found from the relation

$$\tfrac{1}{2} m v^2 = e\overline{U} = e(U - rV_0). \tag{11}$$

It is now assumed that the deflecting field does not change the original direction of the electron beam, i.e., that the initial velocity given by (11) is entirely x-directed. In the absence of any mathematical knowledge of end effects, we cannot make any improvement on Deserno's[6] approximation.

Let $t = 0$ at the instant the electron enters the deflecting system. If $x(t)$ and $y(t)$ are the co-ordinates of the electron at the time t, then in view of (11), the initial conditions may be written as follows:

$$x(0) = 0, \quad \dot{x}(0) = v = [(2e/m)(U - rV_0)]^{1/2},$$
$$y(0) = \eta, \quad \dot{y}(0) = 0. \tag{12}$$

Since r, the ratio of entering-point deflecting potential to plate potential V_0, is a function of η given by (7), the electron entering at any point η will enter with a definite value of v.

The form of (4) for the potential V indicates the convenience of using z and z^* as the dependent variables in place of x and y. With this transformation we find that (9) is replaced by

$$m \frac{d^2 z}{dt^2} + ieV_0 \frac{df(z^*)}{dz^*} = 0 \tag{13}$$

and its complex conjugate. Henceforth, we shall restrict the discussion to the differential equation having z as the dependent variable since the corresponding results for z^* follow directly.

[4] E. T. Whittaker and G. N. Watson, "A Course of Modern Analysis," American Ed., The Macmillan Company, New York, N. Y., p. 390; 1948.

[5] Thus the present treatment is more general than that given by Hutter, who expands the potential function. See R. G. E. Hutter, "The Deflection of Beams of Charged Particles," "Advances in Electronics," vol. I, Academic Press Inc., New York, N. Y. pp. 167–218; 1948.

[6] P. Deserno, "Deflection of the electron beam and spot distortion in a cathode ray tube," *Arch. fur Electrotech.*, vol. 29, pp. 139–148; 1935.

In actual tube operation, the deflection potential V_0 is small as compared to U. To take advantage of this fact in solving (13) we introduce

$$u = vt \tag{14}$$

as the new independent variable and substitute for v the value (12). Thus U is a measure of time, but its proportionality with time is different for each point of entry. Thus we obtain

$$2(U - rV_0)\frac{d^2z}{du^2} + iV_0\frac{df(z^*)}{dz^*} = 0, \tag{15}$$

or letting

$$\tfrac{1}{2}V_0/U = \delta, \tag{16}$$

the final form of the equation

$$(1 - 2r\delta)\frac{d^2z}{du^2} + i\delta\frac{df(z^*)}{dz^*} = 0 \tag{17}$$

with

$$z(0) = i\eta, \qquad \left(\frac{dz}{du}\right)_{u=0} = 1 \tag{18}$$

as the boundary conditions. The solution of this equation involves as parameters the ratio r of entering-point deflecting potential to plate potential V_0 and the ratio 2δ of plate potential V_0 to beam voltage U. The first parameter is fixed for any particular electron, while the second one is a constant for any given problem.

Since δ is a small quantity, z may be expanded in powers of δ as follows:

$$z = z_0(u) + \sum_{n=1}^{\infty}\delta^n z_n(u) \tag{19}$$

Here $z_0(u)$ is the solution of the problem in the absence of deflecting potential. To satisfy boundary conditions (18) we must have

$$z_0 = u + i\eta, \quad \sum_{n=1}^{\infty}\delta^n z_n(0) = 0, \quad \sum_{n=1}^{\infty}\delta^n\left(\frac{dz_n}{du}\right)_{u=0} = 0. \tag{20}$$

As is customary in perturbation calculations, we assume that conditions (20) hold independently for every power of δ, i.e., that

$$z_n(0) = 0, \quad \left(\frac{dz_n}{du}\right)_{u=0} = 0, \quad (n = 1, 2, \cdots). \tag{21}$$

The solution of the differential equation (17) is obtained by substituting expansion (19) for z in the differential equation and equating successive powers of δ to 0. In order to be able to carry it out we must write $df(z^*)/dz^*$ also as a series in powers of δ. This is done by means of a Taylor's expansion in the neighborhood of $z^* = z_0^*$. Since z_0^* differs from u only by an additive constant, d/dz_0^* may be replaced by d/du so that the expansion is written as

$$\frac{df(z^*)}{dz^*} = \frac{df(z_0^*)}{du} + \delta z_1^*\frac{d^2f(z_0^*)}{du^2} + \delta^2\left[\frac{1}{2}z_1^{*2}\frac{d^3f(z_0^*)}{du^3} + z_2^*\frac{d^2f(z_0^*)}{du^2}\right]. \tag{22}$$

Setting the coefficients of various powers of δ equal to zero gives rise to the following equations:

$$\frac{d^2z_1}{du^2} + i\frac{df(z_0^*)}{du} = 0 \tag{23}$$

and

$$\frac{d^2z_2}{du^2} - 2r\frac{d^2z_1}{du^2} + iz_1^*\frac{d^2f(z_0^*)}{du^2} = 0. \tag{24}$$

We stop at the second-order approximation since the neglect of space charge and end effects makes a third-order calculation meaningless in our case. However, in general, the method can be carried to as many approximations as desired.

As shown by the forms of (23) and (24), the problem of solving the complicated differential equation (13) has been reduced to the problem of carrying out integrations. Thus the first integral of (23) is

$$\frac{dz_1}{du} = -i\int_0^u\frac{df(z_0^*)}{du}du = -if(z_0^*) + if(-i\eta) \tag{25}$$

and

$$z_1 = iuf(-i\eta) - i\int_0^u f(z_0^*)du. \tag{26}$$

Similarly after replacing $2ir$ by its value (6) we obtain as the first integral of (24)

$$\frac{dz_2}{du} = f(-i\eta)[f(z_0^*) - f(-i\eta)] - uf(i\eta)\frac{df(z_0^*)}{du} + \frac{df(z_0^*)}{du}\int_0^u f(z_0)du - \int_0^u f(z_0)\frac{df(z_0^*)}{du}du, \tag{27}$$

and after further integration

$$\begin{aligned} z_2 = {} & f(-i\eta)\int_0^u f(z_0^*)du - uf^2(-i\eta) \\ & - f(i\eta)\left[uf(z_0^*) - \int_0^u f(z_0^*)du\right] \\ & + f(z_0^*)\int_0^u f(z_0)du - \int_0^u f(z_0)f(z_0^*)du \\ & - \int_0^u du\left[\int_0^u f(z_0)\frac{df(z_0^*)}{du}du\right]. \end{aligned} \tag{28}$$

This completes the determination of the electron trajectory inside the deflecting plate system. The electron emerges from the deflecting system at $x=1$; the tangent to its trajectory at that point is determined by

$$\tan\beta = (dy/dx)_{x=1}. \tag{29}$$

In terms of the variables u, z, and z^* we may write

$$\tan\beta = \left(\frac{dy}{du}\Big/\frac{dx}{du}\right)_{u=u_k}$$

$$= i\left[\frac{d(z^*-z)}{du}\Big/\frac{d(z^*+z)}{du}\right]_{u=u_k}$$

$$= \frac{1}{2}i\delta\left\{\frac{d(z_1^*-z_1)}{du}+\delta\left[\frac{d(z_2^*-z_2)}{du}\right.\right.$$

$$\left.\left.-\frac{1}{2}\frac{d(z_1^*+z_1)}{du}\frac{d(z_1^*-z_1)}{du}\right]\right\}_{u=u_k} \quad (30)$$

where u_k by definition is the value of u corresponding to $x=1$. To find u_k we therefore solve the equation

$$x(u_k) = \tfrac{1}{2}[z(u_k)+z^*(u_k)] \quad (31)$$

which can be done by Newton's method to any desired degree of approximation. Since $\tan\beta$ contains no terms independent of δ, and since we are only interested in getting the final answer to the order δ^2, we need find u_k only to the order δ. Thus (31) simplifies to

$$x(u_k) = u_k + \tfrac{1}{2}\delta[z_1(u_k)+z_1^*(u_k)] = 1 \quad (32)$$

so that

$$u_k = 1 - \tfrac{1}{2}\delta[z_1(1)+z_1^*(1)]. \quad (33)$$

After all the operations indicated have been performed, the final answer takes the form

$$\tan\beta = -\tfrac{1}{2}\delta[f(1+i\eta)+f(1-i\eta)-f(i\eta)-f(-i\eta)]$$
$$+\tfrac{1}{4}i\delta^2\Big\{f^2(-i\eta)-f^2(i\eta)+f^2(1+i\eta)$$
$$-f^2(1-i\eta)+[f'(1+i\eta)-f'(1-i\eta)]$$
$$\cdot\left[\int_0^1(f(z_0)+f(z_0^*))du-f(i\eta)-f(-i\eta)\right]$$
$$+2\int_0^1 du\left[f(z_0)\frac{df(z_0^*)}{du}\right.$$
$$\left.\left.-f(z_0^*)\frac{df(z_0)}{du}\right]\right\}. \quad (34)$$

Similarly

$$y_k = \eta+\tfrac{1}{2}i\delta[z_1^*(u_k)-z_1(u_k)+\tfrac{1}{2}i\delta^2[z_2^*(1)-z_2(1)]$$
$$= \eta+\tfrac{1}{2}\delta\left\{f(i\eta)+f(-i\eta)-\int_0^1[f(z_0)+f(z_0^*)]du\right\}$$
$$+\tfrac{1}{4}i\delta^2\Big\{f^2(-i\eta)-f^2(i\eta)$$
$$+[f(i\eta)+f(-i\eta)]\int_0^1[f(z_0)-f(z_0^*)]du$$
$$+2\int_0^1(1-u)\left[f(z_0)\frac{df(z_0^*)}{du}-f(z_0^*)\frac{df(z_0)^*}{du}\right]du$$
$$+[f(1+i\eta)-f(1-i\eta)]\left[\int_0^1[f(z_0)+f(z_0^*)]du\right.$$
$$\left.\left.-f(i\eta)-f(-i\eta)\right]\right\}. \quad (35)$$

Since η is a small quantity, (34) and (35) may be expanded in powers of η with only the first few powers being retained. This gives

$$\tan\beta = -\delta\{f(1)-\tfrac{1}{2}\eta^2[f''(1)-f''(0)]$$
$$+(\eta^4/24)[f''''(1)-f''''(0)]\}$$
$$-\eta\delta^2\Big\{f''(1)\int_0^1 f(u)du+2\int_0^1 f'^2(u)du$$
$$+\tfrac{1}{6}\eta^2\Big[8f'(0)f''(0)-11f'(1)f''(1)$$
$$+3f'(0)f''(1)+3f''(1)f''(0)$$
$$-\int_0^1[f''''(1)f(u)-8f''^2(u)]du\Big]\Big\}. \quad (36)$$

Similarly

$$y_k = \eta-\delta\left\{\int_0^1 f(u)du+\tfrac{1}{2}\eta^2[f''(0)-f'(1)+f'(0)]\right\}$$
$$-\delta^2\eta\Big\{f(1)f'(1)-(3/2)f^2(1)$$
$$+\int_0^1[2(1-u)f'^2(u)+f'(1)f(u)]du$$
$$-\tfrac{1}{6}\eta^2\Big[4f'(0)f''(0)+3f(1)f''(0)+5f''(1)f(1)$$
$$-3f'(1)f''(0)+6f'^2(1)-3f'(1)f'(0)-3f'^2(0)$$
$$+\int_0^1[f'''(1)f(u)-8(1-u)f''^2(u)]du\Big]\Big\}. \quad (37)$$

Combining (2), (36), and (37) we obtain the expressions for the co-ordinates of the point of intersection of the bounding rays

$$x_S-1 = p/q;\quad p = 1/\delta^2-\Big\{f(1)f'(1)-3f^2(1)/2$$
$$+\int_0^1[2(1-u)f'^2(u)+f'(1)f(u)]du\Big\}$$
$$q = \int_0^1[f''(1)f(u)+2f'^2(u)]du$$
$$+\tfrac{1}{6}\sigma^2\Big\{8f'(0)f''(0)-11f'(1)f''(1)$$
$$+3f''(1)[f'(0)+f''(0)]$$
$$-\int_0^1[f''''(1)f(u)-8f''^2(u)]du\Big\} \quad (38)$$

where all terms of the order of σ^2 and higher in the numerator and all terms of order of σ^4 and higher in the denominator have been neglected as small corrections.

The larger the value of x_S, the smaller the spot distortion. Since σ^2 is a small quantity, x_S can be made large by choosing $f(u)$ so that the first term in the denominator vanishes, i.e.,

$$\int_0^1 [f''(1)f(u) + 2f'^2(u)]du = 0. \quad (39)$$

It may be shown (e.g., by direct substitution) that the solution of this equation can be put in the form

$$f(u) = F(u) + Au \quad (40)$$

where $F(u)$ is any analytic function of u subject only to the restriction stated below, while A is a root of the quadratic equation

$$2A^2 + A[4F(1) + \tfrac{1}{2}F''(1)] + \int_0^1 [F''(1)F(u) + 2F'^2(u)]du = 0. \quad (41)$$

Since $f(u)$ is assumed to be a real function, $F(u)$ must be chosen so that the roots of (41) are real.[7]

It is possible (in principle at least) to select $F(u)$ so that the factor multiplying σ^2 in the denominator of (38) should vanish making the term in σ^4 previously ignored as a small correction, the leading term. This choice of $F(u)$ would make x_S infinite for all practical purposes. However, as far as tube design is concerned, complete elimination of spot distortion is of little value if it is accompanied by a marked reduction in the magnitude of the deflection. The deflection of the central ray ($\eta = 0$), which can serve as a measure of the deflecting power of the system, is

$$\tan \beta^\circ = -\delta f(1). \quad (42)$$

The optimum solution of the design problem would minimize spot distortion without markedly reducing ing the deflection obtained in a standard tube.

In the following section some special solutions are discussed which accomplish this purpose satisfactorily.

Design Data

In selecting specific plate designs conforming to conditions (40) and (41) for minimizing spot distortion, the main problem is to obtain a large deflection while preventing the electron beam from hitting the plates.

In the examples presented below we make the following assumptions, consistent with standard tube practice as exemplified, for example, by the 7JP4;

1. We let $\delta = 0.025$, i.e., for an anode potential of $U = 6{,}000$ volts the deflecting potentials on the two plates would be ± 300 volts.

2. With the length of the deflecting space set at unity, the half width of the beam is chosen as $\sigma = 0.02$. In the 7JP4 the length of the deflecting space is 3 cm so that our numerical results for length should be multiplied by 3 to reduce them to the same scale. Consider now the special function

$$f(z) = C_1 z - C_3 z^3. \quad (43)$$

The constants in it are determined from the following simple considerations: The smallest spacing between the plates which will allow the deflected beam to clear the plates is $g = 0.055$; furthermore, condition (39) gives $C_3/C_1 = 0.406$. The corresponding values for β° and x_S are

$$\beta^\circ = 15°7' \qquad x_S = 2.8 \times 10^4.$$

Thus this particular design gives a good value for the deflection while completely eliminating spot distortion.

Another simple design which leads to good results is based on the function

$$f(z) = C_1 z - C_4 z^4 \quad (44)$$

where the conditions for minimum spot distortion and maximum deflection consistent with keeping the electron beam from impinging on the plates require that

$$C_4/C_1 = 0.240 \qquad g = 0.076.$$

The resulting values for β° and x_S are

$$\beta^\circ = 14° \qquad x_S = 1.6 \times 10^4$$

The actual curves corresponding to functions (43) and (44) are shown in Fig. 3.

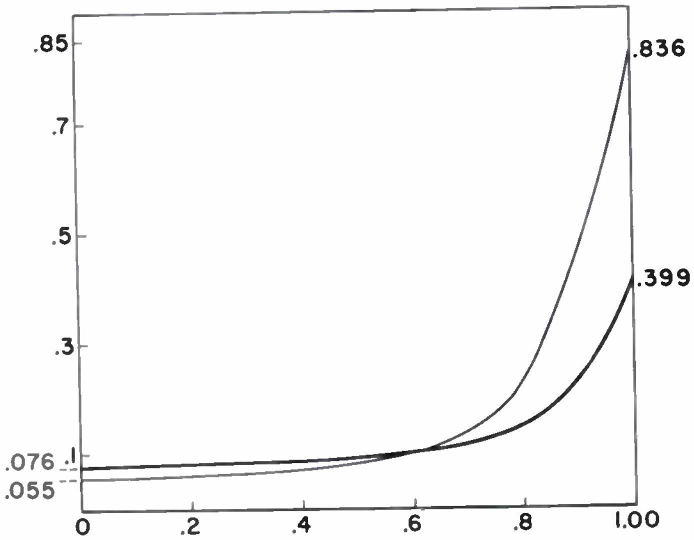

Fig. 3—Plate designs which minimize spot distortion. The heavy curve corresponds to a two-term cubic design function and the lighter curve to a two-term quartic.

It is expected that the results obtained can be improved by introducing more terms in the expressions for the design functions, such as $f(z) = C_1 z + C_2 z^2 - C_4 z^4$, or other simple polynomials.

Acknowledgment

In conclusion, the author wishes to thank I. J. Epstein for checking the formulas and performing the numerical calculations in this paper.

[7] For a more complete treatment of the integral equation problem, see J. E. Rosenthal, "The solution of a certain general type of integral equation," *Proc. Nat. Acad. Sci.*, vol. 36, p. 267; April, 1950.

Progress in Development of Test Oscillators for Crystal Units*

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Summary—**Early crystal unit test oscillators as conceived some 20 years ago were principally duplicates of the actual equipment in which the crystal units were to be utilized, a practice which resulted in a large variety of test circuits and procedures for testing. It is now recognized that a knowledge of the equivalent electrical elements making up the crystal unit is essential to the circuit engineer, and that the older conception of frequency and activity, the latter being an attempt to express the quality of a crystal unit in terms of a particular oscillator circuit, do not define adequately its characteristics. The equivalent electrical circuit of the crystal unit contains essentially a resistance, an inductance, and 2 capacitances, which together with frequency define the performance of the unit. Crystal units are available in the frequency range from about 1,000 cycles to over 100 Mc. Their resistance range may vary from less than 10 ohms to over 150,000 ohms, the inductance from a few millihenries to nearly 100,000 henries and the capacitances from about 0.001 $\mu\mu$f to 50 $\mu\mu$f. Modern test oscillators, with frequency and capacitance measuring apparatus as auxiliary equipment, will measure these quantities with accuracies sufficient to meet present needs. The transmission measuring circuit also is described and is proposed as the standard reference circuit for comparison with the test oscillators.**

I. Introduction

SINCE ITS ADVENT about twenty years ago, the quartz crystal unit has become a familiar component part of electrical circuits. It has made available to the engineer an electrical network, physically small and possessing properties which have been found to be quite unique in the realm of frequency control. The crystal clock as the ultimate in precise timekeeping has only recently been challenged. The extensive use of crystals as a principal means of controlling the frequency of oscillator circuits and in filter networks is well known.

Hundreds of varieties of crystal units have been manufactured, and the demand has been so great that little time has been spared to simplify and standardize their testing procedure. Adequate circuits for testing filter crystals have existed for some time, as the nature of the use of a crystal in filter circuits necessitates knowledge of the fundamental elements of the crystal network. As for oscillator crystals, however, the engineer has had little interest in these properties. As long as the crystal held the frequency to within his desired tolerance and could be utilized in the particular oscillator circuit at hand, the design engineer was satisfied. As a result, the oscillator crystals were calibrated in a large variety of test circuits, some of which were either equivalent circuits or were samples of the actual equipment in which eventually they were to be used. Perhaps the all-time high for the latter was the calibration of some 10 million crystal units manufactured during the war for use in two transmitters, one being a tank radio transmitter and the other of similar design for field artillery communication. These crystals were tested ten at a time in either one or the other of these transmitters, and in the process several transmitters of each type were worn out.

It was during the past war that first attempts were made to standardize testing procedure for crystal units. It is the purpose of this paper to review some of the features of the earlier test sets and to describe some of the more recent developments of test oscillator design.

II. The Crystal Unit

Before going into the details of crystal testing procedure, it may be well to examine briefly some of the characteristics of a crystal. It is of no particular intent to describe the large variety of crystal units which have been manufactured and are in production today. Fig. 1 shows a series of crystal units designed by the Bell Telephone Laboratories and being manufactured, in part, by the Western Electric Company. They cover the frequency range from about 1,000 cycles as illustrated by the 20 series to 100 Mc as illustrated by the 28 series. The 20 series covers the frequency range 1,000 to 330,000 cycles, the 21 and 26 series 16,000 cycles to about 3 Mc, and the 22 and 27 series 200 kc to 15 Mc. The 25 series are similar to the 20 series but adapted for use in filter networks.

All of these units may be represented electrically by a simple network of Fig. 2, which shows the familiar equivalent network of a crystal unit consisting of a parallel circuit of resistance R, inductance L, and capacitance C in series, shunted with a condenser C_0. The reactance and resistance of a typical quartz plate as a function of frequency are shown graphically below the schematic of the crystal. With a crystal plate of the dimensions shown in the figure, and with the two major

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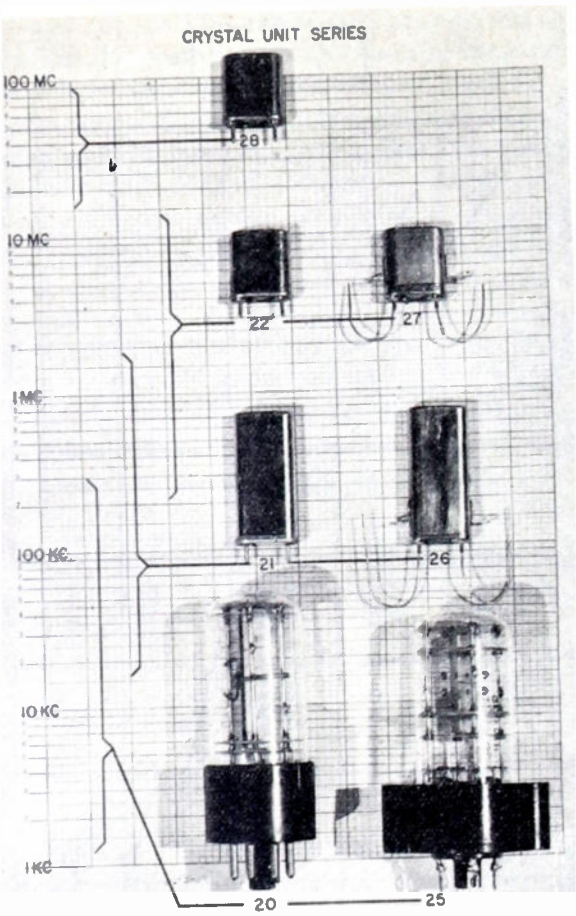


Fig. 1—A series of crystal units in the frequency range 1,000 cycles to 100 Mc. Note: The indicated frequency range of each series is subject to revision and may not cover completely that shown above.

surfaces having electrodes suitably connected to a frequency scanning circuit, responses equivalent to abrupt reactance and resistance changes will occur. No attempt has been made to show all the possible responses but only those useful to the filter and oscillator design engineer. The chart is not drawn to scale, since the regions of abrupt reactance and resistance changes would be included in approximately the thickness of the line.

The first mode of vibration the *CT* mode, is a face shear mode, the frequency being a function of the length and width of the crystal plate. The terms *AT* and *CT* are used rather loosely here; more correctly these modes define certain angles of cut which obtain low temperature coefficients of frequency. Thus a crystal unit oriented for the *CT* mode would not be exactly oriented for the *AT* mode, although the two angles are within a few degrees of each other. Between the *CT* and *AT* modes are overtones of the *CT* mode, which are not in general use at the present time. The overtone modes of the *AT* cut however have been found to be valuable in the manufacture of crystal units operating above 15 Mc.

The principal purposes of this figure are first to show the large number of modes of vibration of a single piece of quartz and secondly to point out that for each mode,

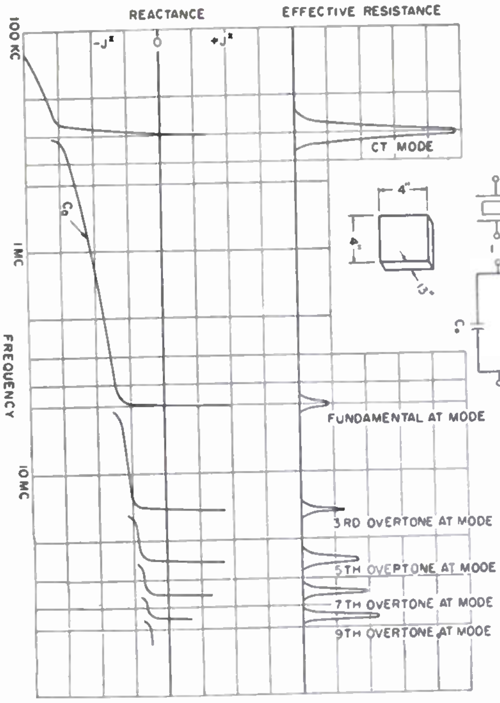


Fig. 2—The equivalent electrical network and characteristics of a crystal unit.

the values of L, C, and R differ. In the example shown, only C_0, the static capacitance of the plate operating as a condenser, remains relatively the same, although it may differ due to the use of different electrode sizes for different modes of vibration.

The engineer usually is interested in one particular mode of vibration and may easily design his circuits to reject the others. His next problem is to decide at what point on the reactance curve he wishes to operate the crystal. Two conditions of operation have become almost universal, one at the point where the reactance is zero, and the other at some point where the reactance of the crystal is positive.

Under the first condition, the crystal is said to be operated under the condition of series resonance and the frequency of operation is closely equal to

$$\frac{1}{2\pi\sqrt{LC}}.$$

The other condition of operation is called the positive reactance condition; the reactance of the crystal is inductive and equal in magnitude to that of a condenser C_T which may be connected in series with or in parallel with it.

The oscillator engineer is interested in the resistance and the frequency at which the crystal is at series resonance, if the crystal is to operate near series resonance, or the effective resistance and frequency of the crystal when operated either in series with or shunted by a condenser. The filter engineer may be interested in the inductance and the capacitances of the condensers C and C_0, in addition to the resistance and series-resonance frequency.

It is therefore important that the crystal test set be capable of furnishing either directly or indirectly these data, and therefore the magnitudes of the crystal parameters are of great importance to the design of crystal test equipment.

Fig. 3 is a chart showing the frequency and resistance range of some typical crystal units. The range of resistance is such that it may vary from less than 1 ohm to over 100,000 ohms, and the frequency from about 1 Kc to 100 Mc. Fig. 4 shows the inductance of a similar series of crystal units. With inductances of these magnitudes and with the corresponding resistances shown in Fig. 3, the Q of the crystal units may vary from about 5,000 to several million. These values of resistance and inductance are real quantities and may be determined from measurement in bridge and other test circuits. Fig. 5 shows the range of the capacitances C and C_0. The latter may be measured with a capacitance bridge. C is usually determined in terms of a ratio with respect to C_0. This ratio may be found by two relatively simple frequency measurements and a capacitance measurement of C_0.

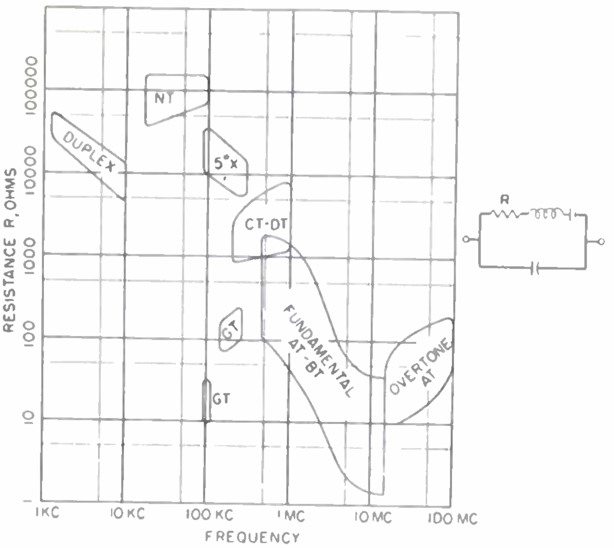


Fig. 3—The resistance R of typical crystal units.

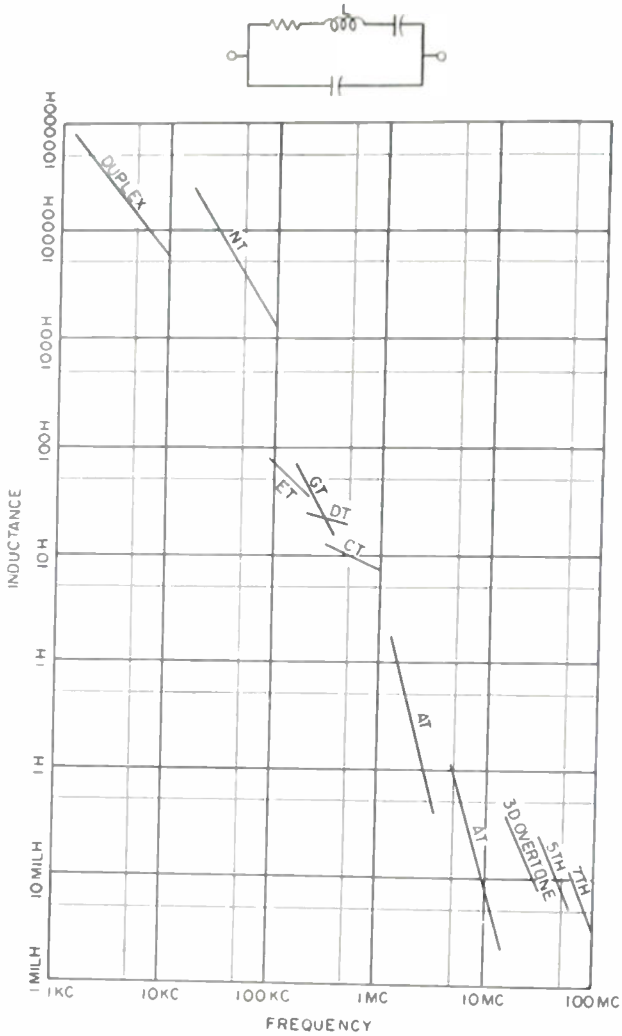


Fig. 4—The inductance L of typical crystal units.

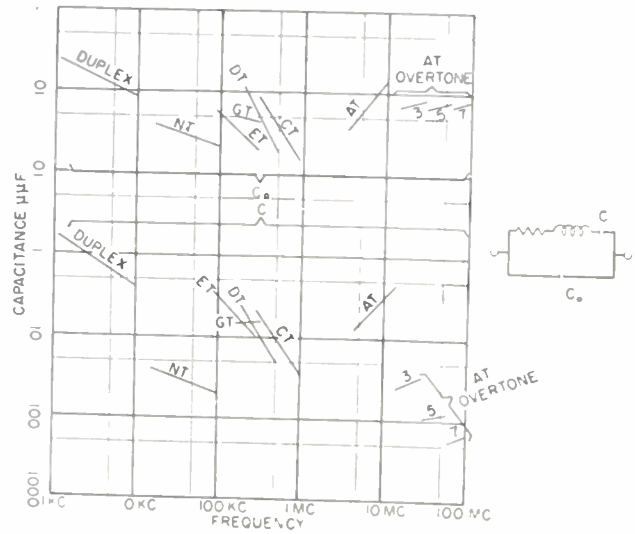


Fig. 5—The capacitances C and C_0 of typical crystal units.

But few of these data were available to the manufacturer of the early oscillator crystal units and the crystals were tested principally in duplicates of the circuits in which they were ultimately to be used.

III. An Early Crystal Test Set

It may be of interest to describe a crystal test set utilized to calibrate crystals for the first crystal-controlled Western Electric Broadcast Transmitters manufactured in 1928, and some of the early attempts to standardize testing procedure by that company and the Bell Telephone Laboratories, Inc.

Fig. 6 shows the schematic of the circuit used both for testing crystals and in the transmitter itself. It is the well-known Miller circuit reduced to its essentials. The oscillator tube was a 50-watt filamentary triode, which, while not worked to its full capacity, operated the crystal at considerably higher potentials than is the custom today. In the original design, X-cut crystals were utilized and the final frequency adjustment was made by changing the operating temperature of the oven. It was not until several years later that an auxiliary condenser was added in shunt with the crystal in order that its frequency could more easily be varied. The tolerance for broadcast stations at that time was ±500 cycles, but since this was before the advent of the low temperature-coefficient crystal, and frequency changes up to 100 cycles per megacycle per degree had to be contended with, the calibration and adjustment of the crystals was somewhat difficult. There was an operating level requirement in terms of oscillator grid current, which was defined by the minimum drive on the following radio-frequency amplifier. Physically, the test set was the complete oscillator section of the transmitter.

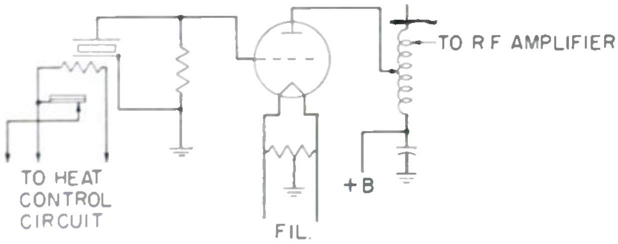


Fig. 6—An early crystal test circuit.

For the next 12 years, crystal specifications were written around test circuits which were either simple circuit boards containing the essential oscillator components or complete oscillator sections of transmitters or receivers. Some of the test sets also contained elaborate ovens and took up considerable bench space. It was not until 1940 that one of the first attempts was made to standardize crystal testing procedure.

A. D-152153 Crystal Unit Test Set

The circuit of the early test set is shown in Fig. 7 and is again the Miller circuit. It contained several coils, a plate-tuning condenser and an adjustable condenser in shunt with the crystal unit. The useful frequency range

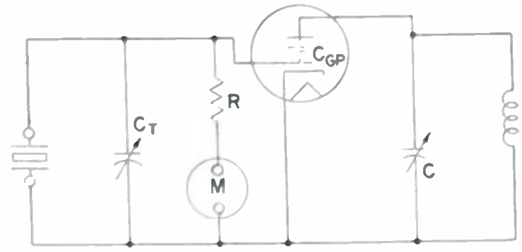


Fig. 7—The D-152153 crystal unit test set circuit.

was from about 500 kc to 15 Mc. The operation of this oscillator was based upon the adjustment of the condenser C_T, so that the crystal unit could be operated at the same frequency as in certain transmitters and receivers being manufactured at that time. In this respect, the oscillator was found to be quite satisfactory, and if the capacitance looking into the grid of the tube of the test set was made equal to the capacitance looking into the grid of the tube in which the crystal was to be used, good frequency correlation was obtained. The effect of the tube input capacitance variation with the tuning of the *LC* circuit was minimized by detuning this circuit on the positive reactance side.

The next step in the use of the test set was to correlate the test-oscillator grid current against the requirements of the crystal units when operated in various transmitters and receivers. From these data, taken with hundreds of crystals, a series of curves of frequency and grid current was obtained which were a compromise between good engineering practice and the ability of the crystal to operate satisfactorily in its destined equipment. This grid current was later defined as activity.

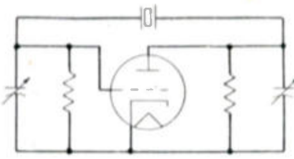
It was found almost immediately, upon starting a study of operation level, that some means had to be provided to assure the user of the test set that its characteristics had not changed during the collection of the data and that worn-out tubes and parts could be replaced without changing the oscillator's original operating conditions. This was accomplished with a series of coils which could be inserted in place of the crystal. They allowed adjustments to be made and tubes to be replaced without changing the over-all operation of the test from that originally conceived.

This oscillator, which was the forerunner of an Army test set known familiarly as the TSM-1, has performed well and is still being utilized in many crystal shops. It is of interest to note that the quality levels established by shop practice some six or seven years ago by more or less empirical means, when translated into equivalent effective resistance, are practically the same levels which are in use today.

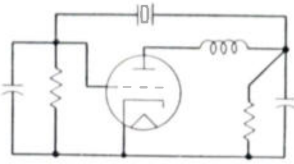
In light of the present methods of specifying crystal characteristics in terms of effective resistance and frequency, the oscillator leaves much to be desired. In spite of the alignment coils, the oscillators have to be made mechanically identical to compare closely with each other, and tubes have to be selected with considerable care. The newer test sets do not have these objections and also permit the more fundamental measurements of series-resonance frequency and resistance and inductance.

B. Adaptions of the Pierce Circuit

The simple Pierce oscillator circuit of Fig. 8(a) was adapted for test oscillators during the war in several test sets. It had the same disadvantages of the TSM-1, namely, difficulty in maintenance of arbitrary performance levels. Fig. 8(b) is an adaption of the Pierce circuit for *CT* and *DT* crystal units in the range of 200 to 1,000 kc. This test set, the TS-221/TM, was a war project designed primarily for five varieties of crystal units, two of which were in large production for use in tank and field artillery transmitters. The test set was finished about the time the war ended and did not see



(a)



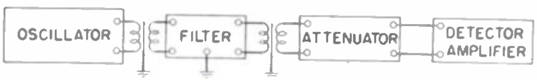
(b)

Fig. 8—Variations of the Pierce crystal oscillator circuit.

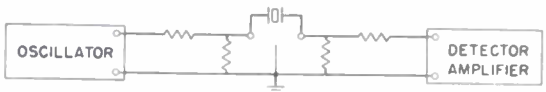
service in the shop. It was also furnished with coils for alignment purposes. These test sets do not measure the electrical characteristics of the crystals and it is expected that they will eventually be replaced by the newer equipments. However, a modified version of the Pierce circuit has been adapted to a series of wire-mounted crystals. This oscillator will be described later.

C. *Transmission Circuits*

Electrical filters were in existence many years previous to 1928 and circuits were available for the measurement of their characteristics. Fig. 9(a) shows a typical setup. The highest oscillator frequency used in early filter measurements was about 100 kc, and the majority of the filters and early crystal filters operated below this frequency. It was natural that with the advent of crystal filters around 1929 that the transmission circuits would be utilized. Early crystal characteristics were obtained from a simple version of transmission circuit shown in Fig. 9(b). The crystal network was a combination of 4



(a)



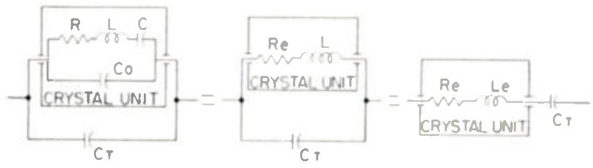
(b)

Fig. 9—(a) Filter transmission test circuit.
(b) Crystal transmission test circuit.

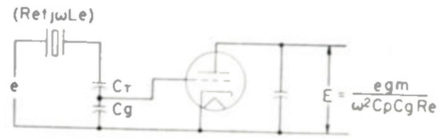
resistors, two shunting either side of the crystal to ground and the other two isolating the crystal from the rest of the circuit. With relatively simple precautions as to the size of these resistors, nearly all the characteristics of the crystal could be obtained from this circuit. The capabilities of the simple resistance network of the transmission-measuring circuit were not realized in test circuits in the early days of oscillator crystal manufacture but have been made a part of some of the more recent test oscillators. Oscillators were available in 1937 for the measurement of filter crystals, but they furnished relative frequency and resistance values obtained during temperature tests of the crystal units. The principal filter crystal characteristics were obtained from transmission-line measurement.

D. *Performance-Index Meter*

In 1944 one of the first instruments designed to measure the fundamental properties of crystal units was developed. From this instrument is obtained a figure of merit defined as the performance index, or the PI of the crystal unit. This quantity, expressed in ohms, is the impedance of the crystal unit when shunted with a condenser C_T at the frequency where the reactance of C_T is equal to positive reactance of the crystal. If the effective resistance of the crystal unit under these conditions is R_e, the equation of PI becomes that shown in Fig. 10(a). The quantity PI is analogous with the shunt impedance or $Q\omega L$ of a coil and condenser, and may be used as such



(a)



(b)

Fig. 10—The performance index (PI) meter circuit.

(a) $$\mathrm{PI} = Q_e\omega L_e = \frac{\omega^2 L_e^2}{R_e} = \frac{1}{\omega^2 C_T^2 R_e} = \frac{Q}{r} = \frac{C_o}{2\pi f_0 (C_o + C_T)^2}$$

$$r = \frac{C_0}{C}.$$

(b) $$\mathrm{PI} = \frac{E}{e}\,\frac{C_pC_g}{C_T^2 g_m}.$$

in the design of oscillator circuits. The PI meter circuit in a simplified form is shown in Fig. 10(b). It is quite similar to the circuit of the Q meter, which measures Q in terms of a voltage ratio. The PI meter measures the quantity PI in the same manner, as is shown by the simplified formula, and the value of PI may be obtained directly from a calibrated dial. Details of this meter are available in technical journals and will not be described further here.

The PI meter is relatively expensive to manufacture as compared to other crystal test apparatus and its operation is somewhat involved. It is more suitable for laboratory use than for shop use, but is, however, an excellent standard of reference for crystals operating under positive reactance conditions and is extremely useful in checking the performance of many test oscillators and other crystal circuits.

IV. Modern Crystal Test Oscillators

A. *Crystal-Impedance Meter*

Fig. 11(a) shows the circuit for a new series of test sets which have recently been made available to the industry, and are versions of the crystal impedance (CI) meter developed by the Frequency Control Branch, Squier Signal Laboratory, at Fort Monmouth, N. J. Two meters, one having the frequency range 76 to 1,110 kc, and the other having the frequency range 0.82 to 15 Mc are in current production by the Lavoie Laboratories, Morganville, N. J. They are essentially oscillators with a crystal network containing a built-in series capacitance C_T and decade resistances. The series-resonance resistance and frequency and the effective resistance and frequency of crystal units operating under positive resistances may be measured. Fig. 11(b) is an expression for the loop gain of the circuit, and for oscillation $E/e=1$. In operation the crystal is switched in the circuit, the two LC circuits being tuned near maximum grid current. The crystal is then switched out, the decade resistance section is switched in and is adjusted until an equivalent grid current has been obtained. This resistance will be the series-resonance resistance R of the crystal if there is no condenser in series with the crystal, or the effective resistance R_e of the crystal if it has the capacitance CT in series with it. The LC circuits are then readjusted to obtain the crystal frequency. The resistance is then replaced by the crystal and the correct crystal frequency may be measured.

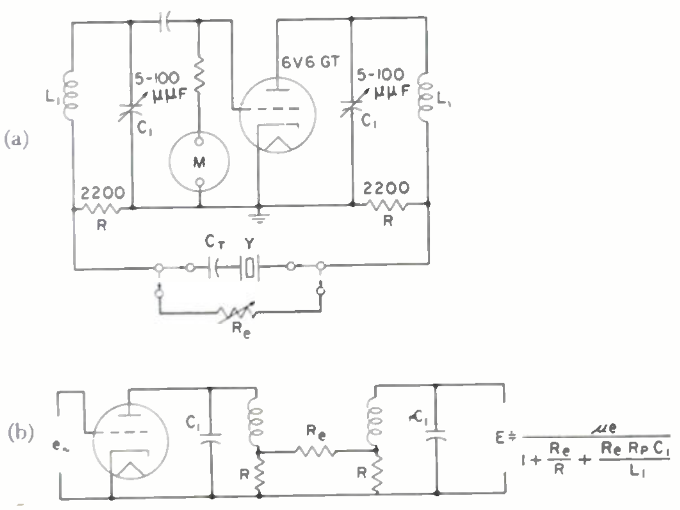


Fig. 11—The crystal impedance (CI) meter circuit.

The sensitivity of the oscillator to resistance changes varies with frequency. At the high-frequency end of the oscillator the right-hand portion of the denominator is predominant, making the meter sensitive to low values of crystal resistance. At the low-frequency end of the oscillator the left side of the denominator predominates, making the meter more sensitive to changes in higher values of effective resistance. The resistance of the crystal units in general vary in a similar manner and therefore a satisfactory measurement of resistance may be obtained for most purposes.

The two CI meters represent a great advance in crystal test set design. Their accuracy as compared to bridge and transmission circuit measurements is about 5 cycles per Mc for frequency and 10 per cent for resistance and inductance, provided precautions are taken with regard to the current through the crystal.

In common with perhaps almost all vibrating devices, the operation of a crystal unit is affected in varying degrees by the amplitude of motion. This is a function of crystal current, and affects both the resistance and frequency of the unit and is independent of changes due to internal heating. Some crystal units using AT- and BT-cut quartz plates are not greatly affected with currents up to several milliamperes. This is not true, however, of some of the wire-mounted, low-frequency crystal units where frequency variations with a current of 1 ma may be excessive when compared with lower values. In some cases the resistance of the unit may double over a relatively small change in current. Development is in progress to reduce these variations and it is hoped that eventually the wire-mounted crystal units will be as free from this undesirable characteristic as some of the higher-frequency crystal units. In the meantime, it will be necessary to define rather exactly the crystal current if consistent resistance and frequency measurements are to be obtained.

To bridge the gap between the present and the time when more stable crystal units will be available, a special oscillator has been developed which is a satisfactory test circuit and at the same time a good "use" circuit. Crystal current conditions may therefore be the same in use as in calibration.

B. *The Positive Reactance Test Oscillator*

The oscillator of Fig. 12 is being utilized to test a series of wire-mounted crystal units in the range 10 to 1,000 kc. All of the units operate under a nominal positive reactance condition equivalent to 20 $\mu\mu$f in series

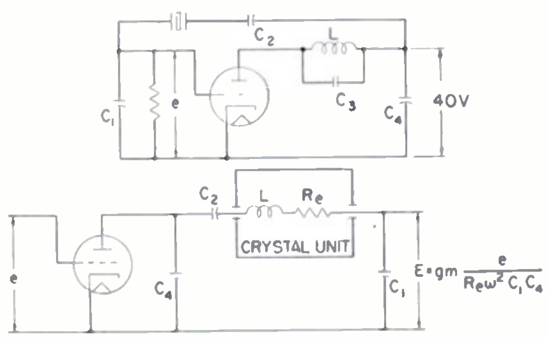


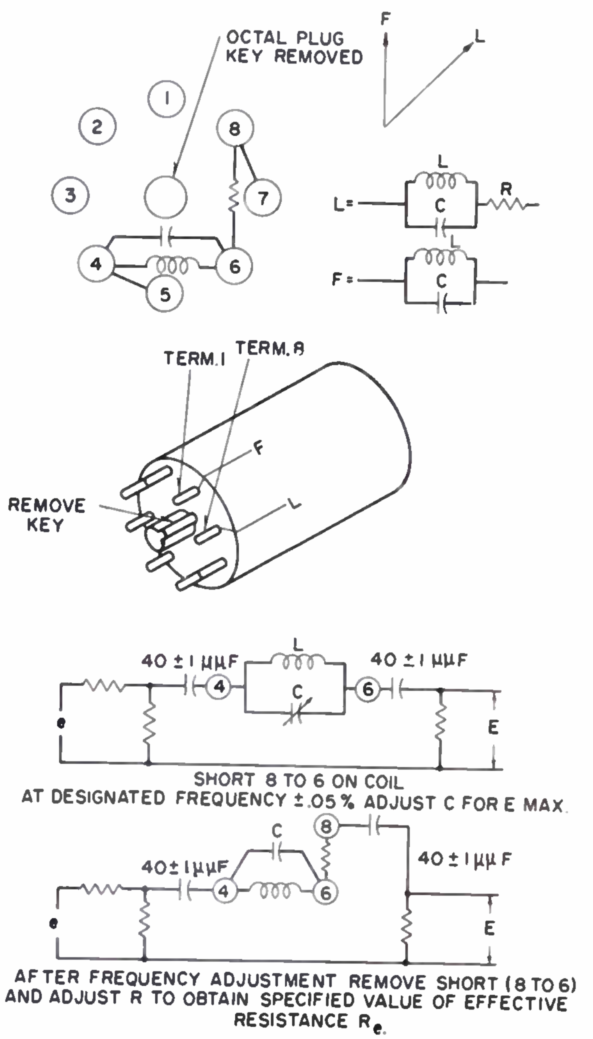
Fig. 12—A positive reactance test oscillator circuit.

$$C_T = \frac{C_2 C_1}{C_1 + C_2} \quad C_4 \gg C_T$$

$$E = \frac{gm e C_T^2 PI}{C_4 C_1}.$$

with the crystal. As a "use" circuit, the adjustment of the condenser C_2 allows sufficient frequency variation to take up manufacturing variations of about ± 0.004 per cent and if the crystals are manufactured within that tolerance, they may be set on exact frequency. A relatively constant output voltage of approximately 4 volts is available across the condenser C_4. The oscillator is made up in 12 fixed frequency bands with no tuning within each band except for the LC_3 circuits, which are adjusted to a designated frequency. Their purpose is to prevent the crystal units from operating at undesired overtone frequencies. Each band covers a range of frequency equivalent to the ratio 1.6 to 1. As a standard oscillator the circuit is relatively simple to set up. The condenser C_1 may be adjusted with sufficient accuracy with a capacitance bridge. The condenser C_4 may be any 10-per cent capacitor. The LC_3 circuits are adjusted to a specified frequency by shorting the crystal socket and condenser C and allowing the circuit to oscillate as an electric oscillator. Fig. 12 shows the expression for the loop gain of the oscillator, and $E/e=1$ for oscillation. Since the loop gain is proportional to PI, the other circuit elements being constant, the oscillator grid current is closely proportional to PI. The adjustment of condenser C_2 is obtained by inserting in the crystal socket an alingment coil. This coil has previously been calibrated and adjusted to have a PI equal to that of a minimum-quality crystal at its designated frequency.

Fig. 13 shows the construction and the method of testing a series of alignment coils. By means of the LC circuit, a positive reactance equivalent to that of the operating capacitance C_T may be obtained. This coil is the equivalent of a crystal unit in so far as its reactance and resistance are equivalent and it therefore fulfills its purpose in the oscillator. It is adjusted in the transmission circuit and used in the oscillator circuit under two conditions. It is first adjusted for maximum transmission in the transmission circuit at a designated frequency by varying the capacitance C. The resistance R is shorted for this measurement. The coil is used in the oscillator with the resistor R shorted, and the oscillator capacitance C_2 is adjusted to obtain the designated coil frequency. This adjustment therefore correlates the condenser C_2. The coil is also adjusted in the transmission circuit to have an over-all effective resistance R_e



Coil Reference Number	Coil Freq. (kc)	For Oscillator Freq. Band	Calibration Current in mμa	Crystal Quality PI (ohms)	L* (mh)	R_e (ohms)	C Approx. (μf)	R Approx. (ohms)
1	13	10–16	50	1,870,000	500	200,000	300	200,000
2	20	16–25	50	800,000	191	200,000	300	200,000
3	32	25–40	50	310,000	67	200,000	350	200,000
4	50	40–62.5	50	130,000	24	200,000	400	200,000
5	80	62.5–100	50	56,000	9	180,000	400	180,000
6	80	62.5–100	20	280,000	93	36,000	20	20,000
7	130	100–160	20	170,000	28	22,000	30	16,000
8	200	160–250	20	115,000	8	13,000	60	10,000
9	320	250–400	20	70,000	3	8,900	80	4,000
10	200	160–250	25	250,000	10	6,500	40	3,000
11	320	250–400	25	96,000	4	6,500	40	3,000
12	520	400–640	20	41,000	1	5,700	80	3,000
13	830	640–1,025	15	18,500	0.25	5,000	120	3,000

* Minimum $Q=20$.

Fig. 13—Standard coils for the positive reactance test oscillator.

Note: 1. Reactance coils are for the alignment of oscillator for frequency and crystal quality. This is accomplished by utilizing the oscillator's crystal socket as a switch which either shorts R for frequency adjustment (position F) or allows R to be inserted in series with the coil L for adjustment of the level of the oscillator (position L) to a given minimum crystal quality.

2. The values of R and C given in the table here are approximate, the exact values being determined by measurement in circuits shown.

3. Care should be taken to maintain minimum capacitance between terminals of the coil and ground during the frequency and effective resistance adjustment of the coils.

hich has been defined by the PI requirement of the rystal. This is accomplished by the addition of the re-istor R in the circuit. The coil is also used under these onditions in the oscillator obtaining a meter reading quivalent to a crystal having the desired minimum quality or minimum PI. Changing the connections for he two conditions is accomplished by two positions of he coil in a socket on the oscillator. The socket is of the ctal type and the key of the coil plug has been removed.

The two CI meters and the oscillator just described rovide the manufacture of crystal units satisfactory scillators for the majority of crystal units in the range 0 kc to 15 Mc. There is also available an oscillator uitable for testing the audio-frequency crystals in the ange of 1 to 10 kc. Its circuit is similar to one of the ircuits now to be described and under development to xtend the range of crystal testing to the higher fre-quencies and also to make available oscillators for test-ng crystal units requiring low current.

C. Oscillators Under Development

Fig. 14 shows the circuits for three test oscillators nder development at the Bell Telephone Laboratories

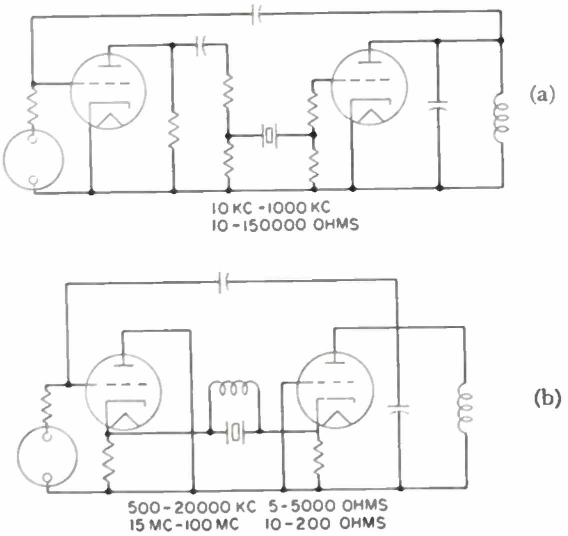


Fig. 14—Series-resonance oscillator circuits.

for filter and oscillator crystal units requiring low crystal current. Fig. 14(a) is a two-tube oscillator with the crystal in a resistance network. This circuit is being designed to cover the frequency range 10 kc to 1,000 kc. Built-in decade resistances are not a part of these oscillators as they are intended for shop use where only a maximum resistance level need be set up. The resistance networks are made in plug-in sections to allow the measurements of crystal resistances of varying values. Plug-in capacitors are also utilized for positive reactance measurements.

The second and third oscillators utilize the Butler cathode-follower circuit of Fig. 14(b). The crystal unit is again in a resistance network which is made plug-in to permit a wide range of crystal resistances to be measured. Other than the cathode connection of the crystal network, both circuits function in a similar manner.

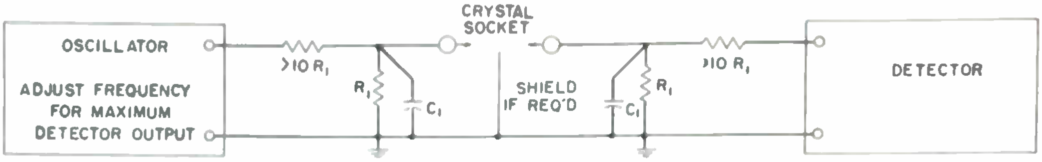
One oscillator has been developed to cover the frequency range 0.5 to 20 Mc, and a second is under development for overtone crystal units operating above 15 Mc. The coil which shunts the crystal is essential for operation above 50 Mc where the resistance of the crystal unit may be of the same magnitude or greater than the reactance of the crystal capacitance C_0.

The method of measurement of these oscillators is the same as that of the CI meters, although it is possible to adjust the LC circuits for maximum grid current to obtain a comparable accuracy of frequency measurement. Their advantage over the present CI meters lies in their ability to measure, with somewhat greater accuracy, low values of resistance throughout the crystal frequency range. This is accomplished by the use of plug-in crystal resistance networks which are not affected by other circuit parameters.

With the completion of the development of the high-frequency test oscillators, the manufacturer will have available test sets which will cover the entire range of operation of crystal units, and, with the procedures set up for the operation, an accuracy of measurement may be obtained which is sufficient so far as present requirements are concerned. However, it is always desirable to have available an independent system of measurement to provide a check of the operation of these oscillators. Such a system should be simple to set up and to operate, and should have an accuracy of measurement somewhat greater than that expected from the shop equipment.

V. The Transmission Test Circuit as a Reference Standard

The transmission test circuit may be set up to meet these requirements. Fig. 15 shows the details of this circuit and also the formula with which its accuracy of measurement may be predetermined. The circuit consists of an oscillator resistance network, detector, and, of course, appropriate frequency-measuring equipment. The oscillator should operate in the frequency range of the crystals and should provide constant output over the limited transmission range of the crystal. It may be a crystal oscillator whose frequency can be adjusted by means of a condenser in series with or shunting the crystal. The resistance networks may be $\frac{1}{2}$-watt carbon resistor, their values being derived from the formulas given below. The components of the circuit should be well shielded. The detector may be an amplifier with a rectifying output meter or a simple rectifier circuit containing a resistance, a crystal diode rectifier, and a dc microammeter. For high detection sensitivity, a radio receiver with some type of output level indicator may be used. The diode rectifier with its associated resistance and meter has been found to be quite satisfactory for frequency measurements above 10 Mc. The detector does not measure exact voltage output, but is used only



TRANSMISSION MEASURING CIRCUIT

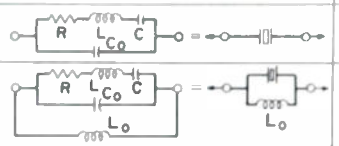
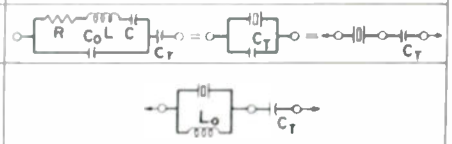
		CONDITION OF OPERATION	
		SERIES RESONANCE	POSITIVE REACTANCE
I	CRYSTAL SCHEMATIC $R/X_0 \ll 1$ $X_0 = -\frac{1}{2\pi f C_0}$		
II	CRYSTAL SCHEMATIC $R/X_0 > 1$ · $f > 50$ MC		
III	FREQUENCY, $f \rightarrow$	$f_{SR} = \frac{1}{2\pi\sqrt{LC}}$ $r = \frac{C_0}{C}$	$f_{PR} = f_{SR}\sqrt{1+\frac{C_0}{r(C_0+C_T)}} \doteq f_{SR}\left[1+\frac{C_0}{2r(C_0+C_T)}\right]$
IV	FREQUENCY ERROR IN CYCLES PER MEGACYCLE DUE TO C_0 $R/X_0 \ll 1$	$-\frac{2R_1R}{rX_0^2} \times 10^6$	$-\frac{2R_1R}{rX_0^2}\frac{C_T}{C_0+C_T} \times 10^6$
V	FREQUENCY ERROR IN CYCLES PER MEGACYCLE DUE TO C_0 $R/X_0 > 1$ $f > 50$ MC	$-\frac{4R_1R}{rX_0^2}\beta \times 10^6$ $\beta \times 100$ = PER CENT L_0C_0 DETUNED FROM f_{SR}	$-\frac{4R_1R\beta}{rX_0^2}\frac{C_T}{C_0+C_T} \times 10^6$
VI	FREQUENCY ERROR IN CYCLES PER MEGACYCLE DUE TO C_1	$\frac{R_1^2}{rX_1X_0} \times 10^6$, $X_1 = -\frac{1}{2\pi f C_1}$	$\frac{R_1^2}{rX_1X_0}\left(\frac{C_T}{C_0+C_T}\right)^2 \times 10^6$
VII	FREQUENCY ERROR IN CYCLES PER MEGACYCLE, DUE TO DETECTOR SENSITIVITY s	$.707\sqrt{S}\frac{(2R_1+R)}{rX_0} \times 10^6$	$.707\sqrt{S}\frac{(2R_1+R_e)}{rX_0}\left(\frac{C_T}{C_0+C_T}\right)^2$
VIII	FREQUENCY ERROR IN CYCLES PER MEGACYCLE DUE TO ERROR IN C_T EQUAL TO ΔC_T MMF	———	$-\frac{1}{2}\frac{\Delta C_T}{C_T}\frac{C_0 C_T}{(C_0+C_T)^2 r} \times 10^6$
IX	RESISTANCE	R =	$R_e \doteq R\left(\frac{C_0+C_T}{C_T}\right)^2$
X	ERROR IN RESISTANCE PER CENT	$\left[\frac{8R_1R}{X_0^2}+\frac{4R_1^2}{X_1X_0}\right] \times 100$	$\left[\frac{-8R_1R}{X_0^2}+\frac{4R_1^2}{X_0X_1}\frac{C_T}{C_0+C_T}\right] \times 100$

Fig. 15—Transmission-measuring circuit and details.

to indicate the maximum voltage or current output when the frequency of the oscillator is varied. The accuracy of measurement, however, is dependent upon the sensitivity of the detector, which may be designated by the letter s and is equal to the smallest detectable change in current divided by the current.

Line I contains the usual unit designations and connections for either series resonance of positive reactance measurements. A simple adapter may be utilized for connecting the crystal unit to the condensers and to the transmission circuit. The ratio of the crystal resistance R to the reactance X_0 of the condenser C_0 is of greatest importance in measurements both in the transmission circuit and in the test oscillator. If the resistance of the crystal unit is high and the reactance X_0 is low the resulting reactance of the crystal unit may never be positive. Under these conditions measurements of resistance in either an oscillator or transmission circuit may be greatly in error. Fig. 16 is a plot of the frequency and resistance of an 85-Mc crystal unit as a function of the number of turns of the shunting coil. Note that if the coil is adjusted to within ±1 per cent, equivalent to 0.1 turn, the error in resistance would be less than 1 per cent and frequency less than about 1 cycle per Mc. These values compare with the resistance of 75 ohms and a frequency difference of −0.002 per cent when the measurements are made without the shunting coil. Line

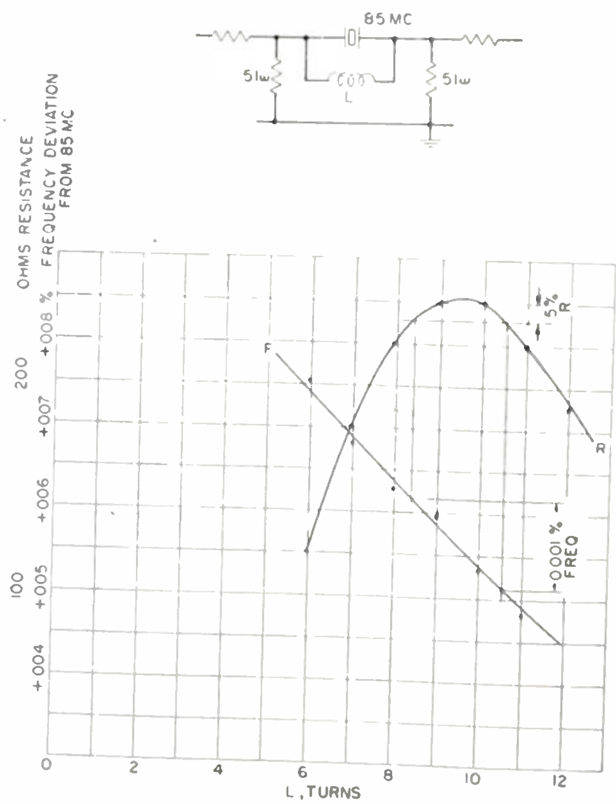


Fig. 16—The frequency and resistance of an 85-Mc crystal unit in relation to the size of the shunting coil.

II of Fig. 15 shows the connections for the crystal and the shunting coil.

Line III contains formulas for the series-resonance frequency and the positive reactance frequency in terms of the series-resonance frequency. r is equal to the ratio of C_O to C. Line IV shows formulas for the error in frequency measurement as a function of the resistance and reactance X_O of the crystal unit. Note that if, in the case of positive reactance measurements, C_T becomes infinite, the condition of series resonance applies. Line V shows the error of measurement when the crystal unit is shunted by a coil. For example, if a coil is tuned to within 1 per cent of the series resonance frequency, β becomes 0.01 and the error of measurement as compared to not utilizing the coil is reduced by 1/50. Line VI shows the error in frequency measurements due to shunt capacitances C_1. Line VII is the error in measurement as a function of the sensitivity of the detector.

With these formulas, the transmission circuit may be designed for any specified accuracy of measurement. As an example of the use of these formulas let us assume that a series of *AT* crystal units is to be manufactured having the following characteristics:

Maximum resistance = 20 ohms
Ratio of capacitance $r = 250$
Range of $C_O = 5.0$ to 10 $\mu\mu$f, with 10 $\mu\mu$f the capacitance of the highest frequency
Frequency range = 5 to 10 Mc
Total holder and socket capacitance to ground C_1 = maximum 5 $\mu\mu$f
Accuracy of frequency measurement = 1 cycle per Mc
Detector sensitivity = 0.1 per cent.

One set of network resistances may be used for the entire series by using values for the least accurate measurement throughout. The limiting crystal is the one which has the greatest resistance and the smallest values of X_O. This is the 10-Mc crystal. From the formulas in Line IV, R_1 is calculated to be approximately 16 ohms. Since the ratio of R to X_O is small, no shunting coil will be needed for these crystals. From the value of R_1 just found, the error due to the shunting capacitance of C_1 will be 0.1 cycle per Mc which may be neglected. However, if in Line VII 0.001 is substituted for s, it will be found that it is necessary to further reduce the shunting resistance R_1 to 6 ohms to obtain an error less than 1 cycle per Mc. This resistance is therefore the one to be used in the network.

In positive reactance measurements the error of measurement of the condenser C_T is of importance. A formula for the determination of this error is given in Line VIII. In the case of the examples just cited, if these crystals were to be operated at positive reactance conditions equivalent to 32 $\mu\mu$f, the accuracy of the latter would have to be within ± 0.1 $\mu\mu$f. This same accuracy would also apply to the adjustment of the condenser in a test oscillator.

The formulas for resistance for positive reactance and series resonance are given in Line IX. The error in resistance measurement is given in Line X. In the example just cited with the resistance R_1 equal to 6 ohms, the error in the resistance measurement would be -0.04 per cent.

The transmission circuit has been utilized over the complete frequency range of available crystal units. As an example of the capability of this circuit, Fig. 17 shows

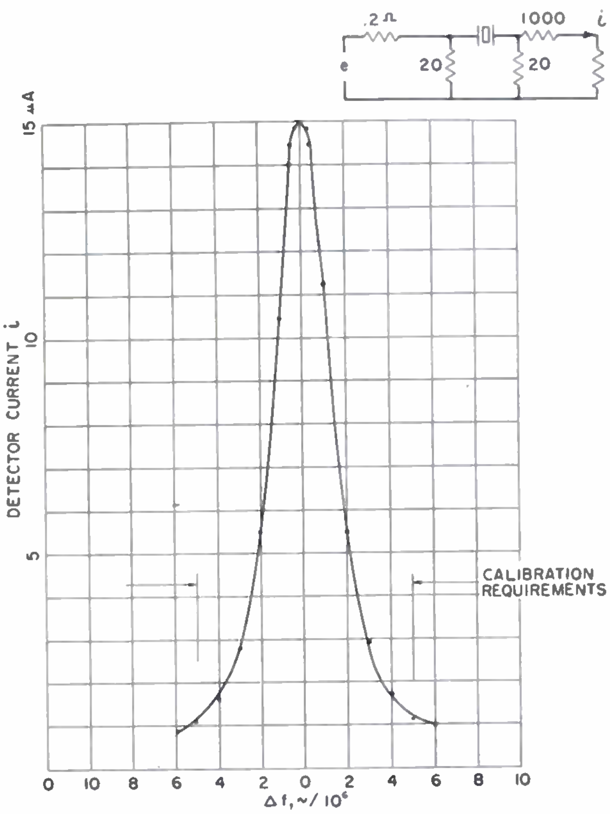


Fig. 17—Transmission characteristics of a high-quality 100-kc crystal unit.

a transmission curve made with a high quality 100-kc frequency standard crystal unit having a Q of over one million. The test-set requirement for these crystals is ± 5 cycles per Mc. The curve shows that the measurement may be made with a relatively insensitive detector to meet this requirement.

Conclusions

In conclusion, it may be of interest to show typical data sheets (see Fig. 18) which may be furnished by the manufacturer of a crystal unit to engineers interested in their use. All of these data may be obtained with suitable capacitance bridges, reactance bridges, frequency measuring equipment, and the test sets which have been described. The data sheets give the engineer complete information necessary for the utilization of these crystal units.

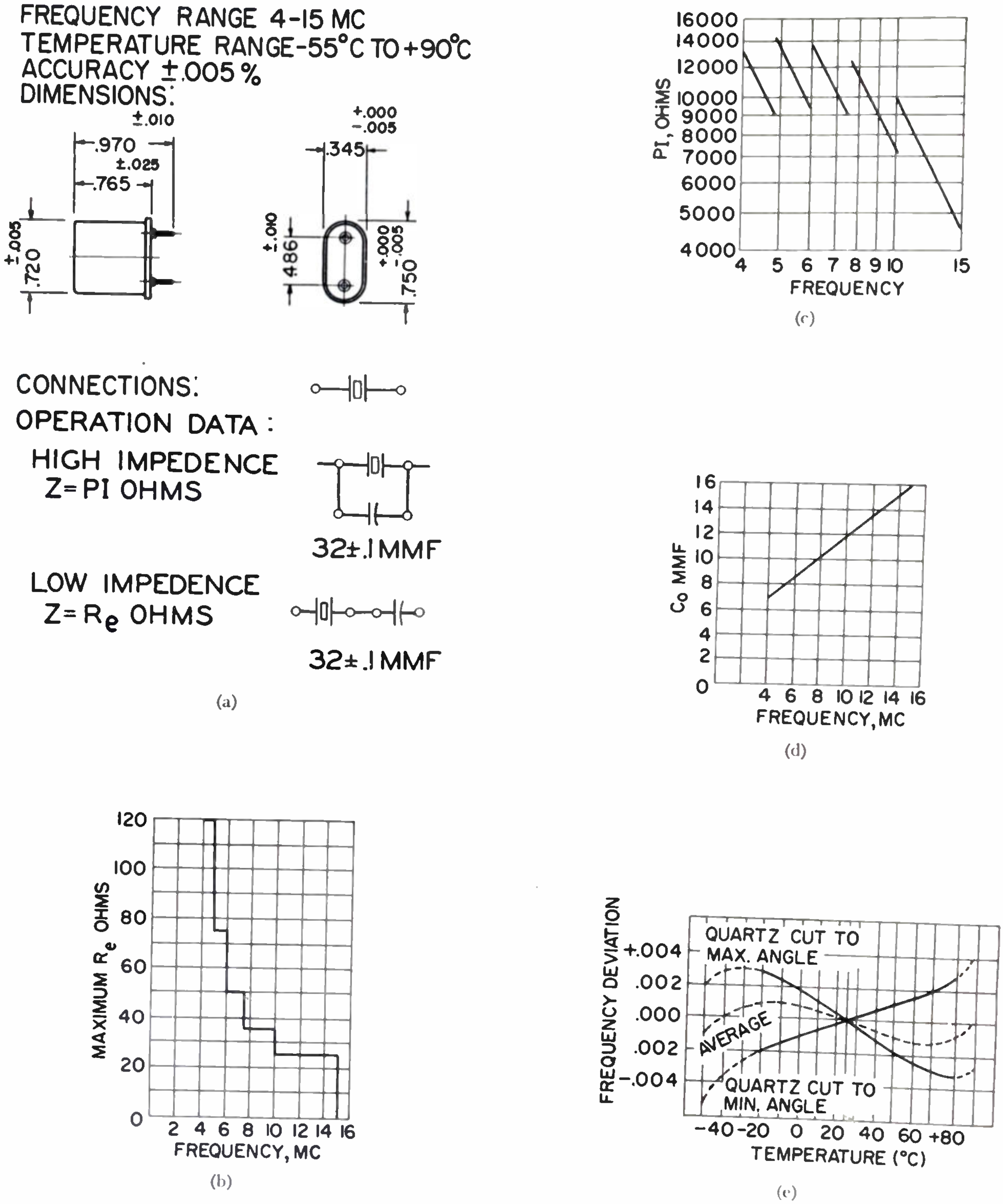


Fig. 18—Crystal unit data sheets. (a) XX-II crystal unit. (b) Data 1. (c) Data 2. (d) Data 3. (e) Data 4

Note: Minimum R_e: see Data 1.
Maximum PI: see Data 2.
Maximum crystal current $=0.020\ A$.
Capacitance of terminals to can $=0.8\pm0.2\ \mu\mu f$
Frequency F_T of crystal unit at operational capacitance $=C_T\ \mu\mu f$.

$$F_T = F_{32}\left[1 + \frac{C_o(32 - C_T)}{2r(C_o + 32)(C_o + C_T)}\right].$$

F_{32} = Frequency at 32 $\mu\mu f$.
$r = 250$.
C_o: see Data 3.
Typical frequency-temperature characteristics: see Data 4.

Standards on Circuits: Definitions of Terms in Network Topology, 1950*

COMMITTEE PERSONNEL

Standards Committee, 1950–1951

Circuits Committee

Accessible Terminal. A network node that is available for external connections.

Arm. *See:* Branch.

Branch (Arm). A portion of a network consisting of one or more two-terminal elements in series.

Branch Point. *See:* Node.

Bridged-T Network. A T network with a fourth branch connected across the two series arms of the T, between an input terminal and an output terminal.

Cascade. *See:* Tandem.

Circuit. A network providing one or more closed paths.

Connected. A network is connected if there exists at least one path, composed of branches of the network, between every pair of nodes of the network.

Cut-Set. A set of branches of a network such that the cutting of all the branches of the set increases the number of separate parts of the network, but the cutting of all the branches except one does not.

Degrees of Freedom on a Mesh Basis. *See:* Nullity.

Degrees of Freedom on a Node Basis. *See:* Rank.

Delta Network. A set of three branches connected in series to form a mesh.

Dual Networks. *See:* Structurally Dual Networks.

Element. Any electrical device (such as inductor, resistor, capacitor, generator, line, electron tube) with terminals at which it may be directly connected to other electrical devices.

Four-Pole. *See:* Two-Terminal Pair Network.

H Network. A network composed of five branches, two connected in series between an input terminal and an output terminal, two connected in series between another input terminal and output terminal, and the fifth connected from the junction point of the first two branches to the junction point of the second two branches.

Junction Point. *See:* Node.

L Network. A network composed of two branches in series, the free ends being connected to one pair of

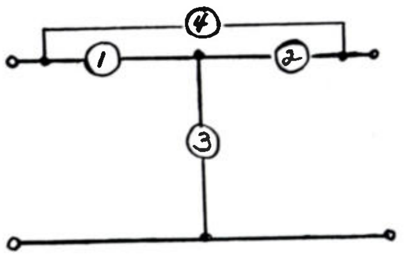


Fig. 1—Bridged-T network branches 1, 2, and 3 comprise the T network and branch 4 is the fourth branch.

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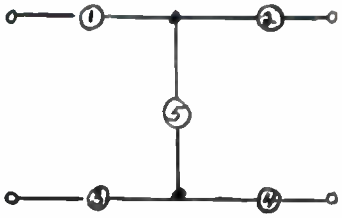

Fig. 2—H network. Branches 1 and 2 are the first two branches between an input and an output terminal; branches 3 and 4 are the second two branches; and branch 5 is the branch between the junction points.

terminals, and the junction point and one free end being connected to another pair of terminals.

Ladder Network. A network composed of a sequence of H, L, T, or pi-networks connected in tandem.

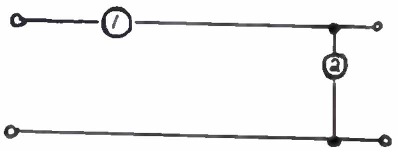

Fig. 3—L network. The free ends are the left-hand terminal pair, and the junction point and one free end are the right-hand terminal pair.

Lattice Network. A network composed of four branches connected in series to form a mesh, two nonadjacent junction points serving as input terminals, while the remaining two junction points serve as output terminals.

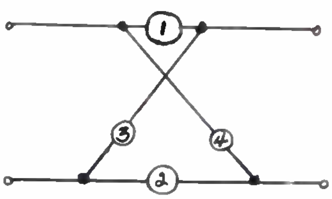

Fig. 4—Lattice network. In the mesh 1, 2, 3, 4, the junction points between branches 4 and 1 and between branches 3 and 2 are the input terminals, and the junction points between branches 1 and 3 and between branches 2 and 4 are the output terminals.

Loop. *See note under* Mesh.

Mesh. A set of branches forming a closed path in a network, provided that if any one branch is omitted from the set, the remaining branches of the set do not form a closed path. (Note—The term *Loop* is sometimes used in the sense of *Mesh.*)

Network. A combination of elements.

Node (Junction Point) (Branch Point) (Vertex). A terminal of any branch of a network or a terminal common to two or more branches of a network.

Nonplanar Network. A network which cannot be drawn on a plane without crossing of branches.

***N*-Terminal Network.** A network with N accessible terminals.

***N*-Terminal Pair Network.** A network with $2N$ accessible terminals grouped in pairs. In such a network one terminal of each pair may coincide with a network node.

Nullity (Degrees of Freedom on Mesh Basis). The number of independent meshes that can be selected in a network. The nullity N is equal to the number of branches B minus the number of nodes V plus the number of separate parts P. $N = B - V + P$.

Parallel Elements. (a) Two-terminal elements are connected in parallel when they are connected between the same pair of nodes. (b) Two-terminal elements are connected in parallel when any cut-set including one must include the others.

Parallel Two-Terminal Pair Networks. Two-terminal pair networks are connected in parallel at the input or at the output terminals when their respective input or output terminals are in parallel.

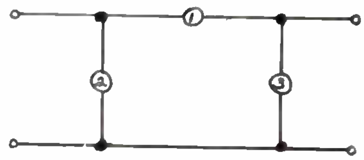

Fig. 5—Pi network. The junction point between branches 1 and 2 forms an input terminal, that between branches 1 and 3 forms an output terminal, and that between branches 2 and 3 forms a common input and output terminal.

Pi Network. A network composed of three branches connected in series with each other to form a mesh, the three junction points forming an input terminal, an output terminal, and a common input and output terminal, respectively.

Planar Network. A network which can be drawn on a plane without crossing of branches.

Quadripole. *See:* Two-Terminal Pair Network.

Rank (Degrees of Freedom of a Node Basis). The number of independent cut-sets that can be selected in a network. The rank R is equal to the number of nodes V minus the number of separate parts P. $R = V - P$.

Separate Parts of a Network. The parts which are not connected.

Series Elements. (a) Two-terminal elements are connected in series when they form a path between two nodes of a network such that only elements of this path, and no other elements, terminate at intermediate nodes along the path. (b) Two-terminal elements are connected in series when any mesh including one must include the others.

Series Two-Terminal Pair Networks. Two-terminal pair networks are connected in series at the input or at the output terminals when their respective input or output terminals are in series.

Star Network. A set of three or more branches with one terminal of each connected at a common node.

Structurally Dual Networks. A pair of networks such that their branches can be marked in one-to-one correspondence so that any mesh of one corresponds to a cut-set of the other. Each network of such a pair is said to be the dual of the other.

Structurally Symmetrical Network. A network which can be arranged so that a cut through the network produces two parts that are mirror images of each other.

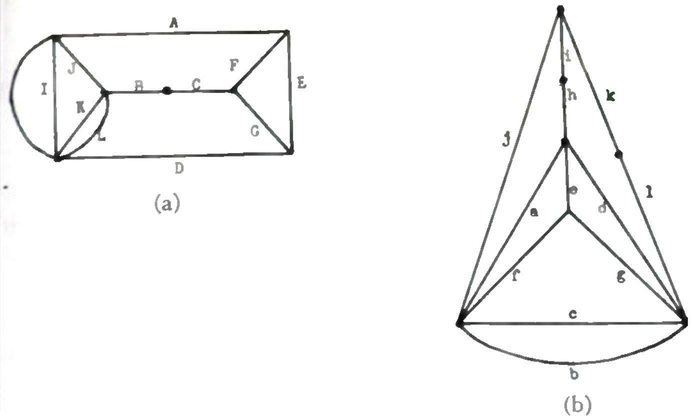

Fig. 6—Structurally dual networks. For example, the mesh *EFG* in (a) corresponds to the cut-set *efg* in (b), the mesh *bc* in (b) to the cut-set *BC* in (a), and the mesh *JAEGCB* in (a) to the cut-set *jaegcb* in (b).

Symmetrical Network. *See:* Structurally Symmetrical Network.

Tandem (Cascade). Two-terminal pair networks are in tandem when the output terminals of one network are directly connected to the input terminals of the other network.

Terminal. A point at which any element may be directly connected to one or more other elements.

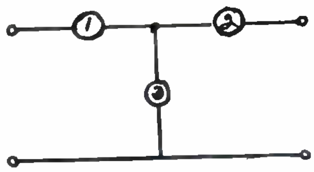

Fig. 7—T network. One end of each of the branches 1, 2, and 3 is connected to a common point. The other ends of branches 1 and 2 form, respectively, an input and an output terminal, and the other end of branch 3 forms a common input and output terminal.

Terminal Pair. An associated pair of accessible terminals, such as input pair, output pair, and the like.

T Network. A network composed of three branches with one end of each branch connected to a common junction point, and with the three remaining ends connected to an input terminal, an output terminal, and a common input and output terminal, respectively.

Tree. A set of connected branches including no meshes.

Two-Terminal Pair Network (Quadripole) (Four-Pole). A network with four accessible terminals grouped in pairs. In such a network one terminal of each pair may coincide with a network node.

Vertex. *See:* Node.

Y Network. A star network of three branches.

Standards on Television: Methods of Measurement of Electronically Regulated Power Supplies, 1950*

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1. INTRODUCTION

Power to operate the electrical circuits of a television transmitter, its studio circuits and field pickup equipments, is usually supplied from the alternating-current mains of a public utility. Some of it is used through transformers to supply filament and heater power for the numerous vacuum tubes necessary to the operation of the system and some of it is converted to direct current for supplying the electrode potentials of these tubes.

Regulation or maintenance of the constancy of voltage and currents applied to these intricate circuits has become of great importance due to their complexity as well as their performance requirements.

1.1 General Description

Automatic line voltage regulators are of several varieties; e.g., saturated magnetic field transformers, electronically controlled circuits and regulators supplying both alternating- and direct-current outputs; constant speed alternating- and direct-current generators; servo adjustable units; batteries floating across a line rectifier; and so forth.

Direct-current supply for most vacuum-tube circuits is conveniently obtained through tube rectifiers and filters which may be supplied from the ac source through step-down or step-up transformers. Regulation of the output voltage and maintenance of a low source impedance are the functions of vacuum-tube circuits associated with these rectifiers.

An electronically regulated power supply may be thought of as a direct-coupled feedback power amplifier which operates by amplifying an internal dc voltage obtained usually from a battery or gas regulator tube. The output is the power delivered to the load circuit and is obtained from the power output stage of the amplifier circuit. To maintain constant the output voltage for various load currents, and for variations in the amplifier circuit introduced principally by the power supply source or rectifier, negative feedback is introduced between output and input. Sufficient gain and bandwidth are usually supplied to reduce the output impedance to a very low value over the low and middle video-frequency band. The low output impedance provides a source of voltage independent (within limits) of the load demands. Rectifier ripple, introduced in the power output amplifier section, is usually reduced to negligible values.

The important characteristics of electronically regulated power supplies currently used at television installations are described, and methods of measuring these characteristics are detailed.

1.2 Definitions

1.2.1 Output Voltage Stabilization

The change in output voltage, at a specified constant load current, resulting from a change of input voltage between two specified values.

1.2.2 Output Voltage Regulation

The change in output voltage, at a specified constant input voltage, resulting from a change of load current between two specified values.

1.2.3 Output Ripple Voltage of a Regulated Power Supply

For the purpose of this Standard, that portion of the output voltage harmonically related in frequency to the input voltage and arising solely from the input voltage. In television, ripple voltage is usually expressed explicitly in peak-to-peak volts to avoid ambiguity. (Note—IRE (in published Standards **38 IRE 17. S1** and **48 IRE 2., 11., 15. S1**) defines "per-cent ripple" as the ratio of effective (rms) value of the ripple voltage to the average value of the total voltage expressed in per cent.)

1.2.4 Regulation Pull-Out (Regulation Drop-Out)

The load currents at which the power supply fails to regulate when the load current is gradually increased or decreased.

1.2.5 Output Impedance

Of a device, the impedance presented by the device to the load.

1.2.6 Unregulated Voltage in an Electronically Regulated Power Supply

The voltage at the output of the rectifier filter.

1.2.7 Regulated Power Supply Efficiency

The ratio of the regulated output power to the input power.

1.2.8 Miscellaneous

1.2.8.1 Unit Warm-Up Time. For purposes of this measurement, the interval between the time of application of input power to the unit and the time at which the regulated power supply is supplying regulated power at rated output voltage.

1.2.8.2 Output Capacitor Discharge Time. For purposes of this measurement, the interval between the time at which the input power is disconnected and the time when the output voltage of the unloaded regulated power supply has decreased to a specified safe value.

2. MEASUREMENT

2.1 Basic Methods of Measurement

2.1.1 Output Voltage Stabilization

The changes in the output dc voltage of electronically regulated power supplies due to fluctuations in line supply voltage are quite small in comparison to the total output voltage. Normal dc voltmeters cannot indicate a change of tens of millivolts in the presence of hundreds of volts. It therefore becomes necessary to employ a reference source of potential to balance the normal output of the regulated supply under test, and to use a millivoltmeter to read the change in output voltage that corresponds to a specified change in input voltage.

Instantaneous variations of the input voltage or load current above or below the limiting values may temporarily overload or paralyze the regulator amplifier in the power supply and produce surges in the output voltage. Such abnormal conditions are beyond the scope of this Standard.

2.1.2 Output Voltage Regulation

The output voltage regulation tests use the same apparatus and connections as the previously described one, except that the load current is varied from the specified minimum to the specified maximum. The difference between the voltage of the reference source and that of the supply under test for various load currents can then be obtained.

2.1.3 Output Ripple Voltage

The output ripple voltage can be measured most conveniently in terms of peak-to-peak amplitude by means of a sensitive oscilloscope, which will also display its complex wave shape. A selective amplifier, such as a current analyzer, is required to measure the value of each harmonic, while the total could be measured with a suitable peak-to-peak voltmeter.

2.1.4 Regulation Pull-Out

The regulation pull-out point, the value of load current at which the supply no longer regulates, is dependent on the output as well as the input voltage. At this point, the output voltage may not show any significant drop, but generally the output ripple components increase suddenly. An arbitrary value of ripple increase can be fixed as defining this point. If the ripple voltage has been measured for regulated operation, a value of ten times this ripple voltage would conveniently define the pull-out point. Measurement of this factor for various output voltages and for various input voltages will define the area of operation of a regulated power supply.

2.1.5 Output Impedance

Alternating components of the load current of the regulated power supply produce voltages across the output terminals of the device which are functions of its impedance. The output impedance is of importance when, for example, the power supply unit supplies low-frequency or cascaded-amplifier stages, or is supplying power to several circuits operating at different signal levels, or is a portion of a feedback circuit.

The output impedance of an electronically regulated power supply may vary between tens of ohms and thousandths of ohms. For the higher values, bridge measurements[1] may suffice. For low values of impedance, the following method is offered and may give more accurate results.[2]

An accurately calibrated resistor, of a value approximately equal to the power supply impedance, is connected in series with the output of the unit under test. The power supply furnishes current to a load circuit (for example, see Fig. 3) whose current demand can be controlled in frequency and also preferably in amplitude. The voltage across the resistor at the test frequency is compared with that across the power supply output terminals by means of an oscilloscope or other sensitive ac voltmeter. At frequencies which are multiples of input power frequency this difference must be large compared with the ripple voltage, otherwise it will be obscured by the ripple. The impedance of the power supply can then be calculated from the values found for the range of frequencies and amplitudes used in this test.

2.1.6 Unregulated Voltage

Unregulated voltage is measured by means of a dc voltmeter and an oscilloscope connected to the rectifier filter output.

2.1.7 Regulated Power Supply Efficiency

The efficiency of the power supply in terms of dc output power divided by the total input power, including its own tube heater power, is derivable from the measurement of output power and input power.

2.1.8 Miscellaneous

2.1.8.1 A measurement of the warm-up time of a regulated power supply will include the heating time of the various tubes, with or without time delay switches, and the charging time of the several capacitors. This time generally can be measured in seconds with the aid of a stop watch.

2.1.8.2 Output capacitor discharge time of the regulated power supply is measured without any load but with the voltmeter across its output terminals. This time generally can be measured in seconds with the aid of a stop watch.

[1] F. V. Hunt and R. W. Hickman, "On electronic voltage stabilizers," *Rev. Sci. Instr.*, vol. 10, p. 6; January, 1939.

[2] J. H. Hersey, "Dynamic impedance of regulated power supplies," *Bell Lab. Rec.*, vol. 27, p. 216; June, 1949.

2.1.8.3 The design of power transformers used in regulated power supplies is often governed not by losses but by physical volume and ventilation. A design which may be conservatively rated for use in open air may be unsatisfactory in a confined space. The safe operating temperature of a transformer—usually recommended as 55° C above surrounding air—may be exceeded due to its placement adjacent to and on the same mounting as power tubes used in the regulated power supply. Ventilation is likewise an important factor in the temperature rise of the components of a regulated power supply.

In order to measure the increase in temperature of a regulated power supply under load, it is necessary to operate it under conditions which will produce normal full load heating for a sufficient length of time for the temperature of its components to become constant. (Approximately 2 or 3 hours' operation under load generally suffices.) The temperature rise may be obtained either by thermometer measurements or by measuring the increase in resistance of a winding of the power transformer. (Transformer winding temperatures are always to be ascertained by the resistance method.)[3]

The resistance for the calculation of temperature rise may be obtained from measurements made by any suitable method at the beginning and at the end of a rated power run. The rise in temperature can be calculated from the following formula:

$$R_t = R_T\left[1 + \alpha_T(t - T)\right]$$

$$(t - T) = \frac{R_t - R_T}{\alpha_T R_T}$$

in which

T = room temperature in degrees C

t = the calculated average temperature of the heated transformer winding

R_t = the measured hot resistance

R_T = the resistance measured at the beginning of the test, i.e., at room temperature, and α_T is the temperature coefficient of the copper conductors, taken as $1/234.5+T$ or $\alpha_T = 0.00393$ at 20° C.

A "hottest spot" correction of 10° C is to be added to this calculated average temperature rise.

2.2 Practical Measuring Devices

The circuit shown in Fig. 1 is for the measurement of output voltage stabilization and of regulation.

Fig. 2 is a reproduction of the appendix to footnote reference 1 in which a measurement technique for output impedance is detailed. The simple ac bridge which is suggested serves admirably for measuring internal impedance in the range of ohms.

For lower impedances in the orders of hundredths and thousandths of an ohm, the circuit shown in Fig. 3 has been found more useful.[2]

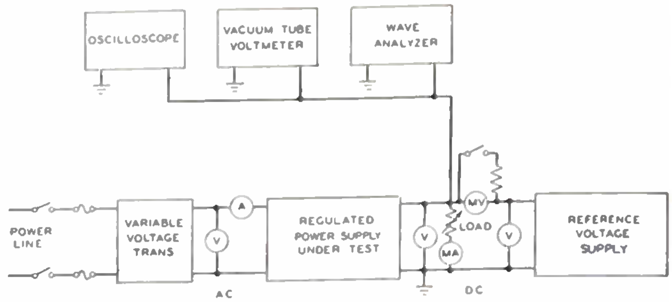

Fig. 1—Schematic of circuit employed for determining regulated power-supply characteristics.

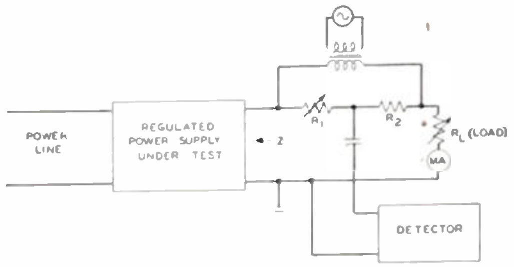

Fig. 2—Bridge circuit for measurement of output impedance of a regulated power supply. At balance $Z = R_1(R_L/R_2)$.

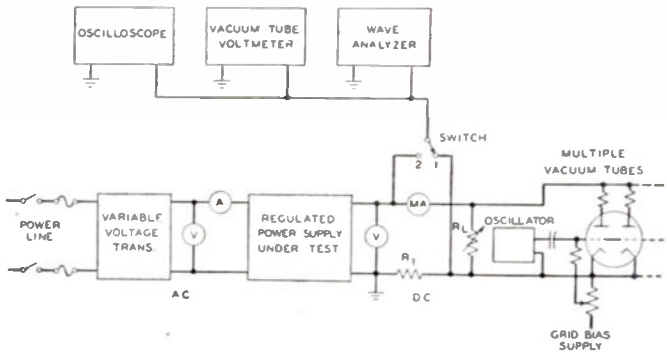

Fig. 3—Testing circuit for measuring output impedance of a regulated power supply.

2.3 Requirements of Measuring Equipment

The reference voltage supply shown in Fig. 1 supplies sufficient current to operate the millivoltmeter. A low internal-resistance regulated power supply operating through a line voltage regulator is to be preferred over a reference battery for this use, since the internal resistance of the series of cells may be of considerable value. A second regulated power supply similar to the one being tested and known to regulate small currents could be used for this reference voltage supply since the current demand from it for this test is very low.

For ripple voltage, regulation pull-out and output impedance measurements, a sensitive calibrated oscilloscope is necessary. Voltage drops produced by ripple currents and ac load currents flowing through the low output impedances of regulated power supplies intended

[3] *American Standard for Transformers, Regulators, and Reactors*, ASA C57.1-1942 and C57.2-1943.

r television use are in the millivolt range. The indi-ıting oscilloscope requires ac amplification, internal · external, to produce sufficiently large deflections for :curate measurements.

.For the output impedance measurement, a vacuum-ıbe load circuit is connected to the power supply as ıown in Fig. 3. Applied to the control grids of the ıbes is the output of an oscillator which may be djusted in frequency and in amplitude. The oscillator ıould cover the range from the lowest frequency ca-able of being measured by the above oscilloscope to the ·equency where the output capacitor of the regulated ower supply determines the impedance range. This pper limit of frequency may be in the hundreds of kilocycles per second. The load current through the impedance of the power supply must be great enough to produce a measurable deflection on the oscilloscope. For an illustrative example, a current of 100 milliamperes rms through an impedance of 0.01 ohm will produce only 2.8 millivolts peak-to-peak. Shown as R_1 in Fig. 3 is a calibrated resistor whose value should be of the same order as that of the impedance being measured. Direct-current ammeter shunts are convenient to use for this purpose and are calibrated in resistance to an accuracy of 1 per cent and better. Their impedance at the top frequency to be measured can usually be assumed as equal to the dc resistance.

3. PROCEDURE

The following is a test procedure using, as an example, typical regulated power supply furnishing 500 ma at 00 volts. **The voltages normally employed in regulated ·ower supplies are sufficiently high to endanger human ıfe. Every reasonable precaution should be taken by esting personnel to minimize the shock danger while .djustments and tests which demand normal operation ·f the power supply are being made.**

3.1 Output Voltage Stabilization

:.1.1 Connect the supply under test according to the dia-gram of Fig. 1, using a variable voltage transformer to ;et the input at exactly rated input voltage.

}.1.2 Adjust the reference voltage supply to equal the dc voltage output under open circuit before the meter switch is opened.

3.1.3 Open meter switch and carefully readjust reference voltage supply to obtain a midscale reading on millivoltmeter.

3.1.4 Vary the input voltage over an appropriate range centered on the rated input voltage, recording the output voltage as the sum of the millivoltmeter reading at each point and the reference voltmeter reading.

The same procedure is followed at dc load currents from zero to the maximum rated current.

The results can be plotted as shown in Fig. 4.

In making the above measurements, it is suggested that the necessary data be obtained as quickly as possible, since even a slight amount of drift is likely to affect the accuracy of measurement.

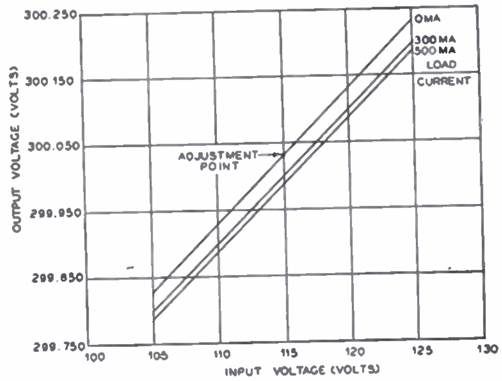

Fig. 4—Typical output voltage stabilization curves.

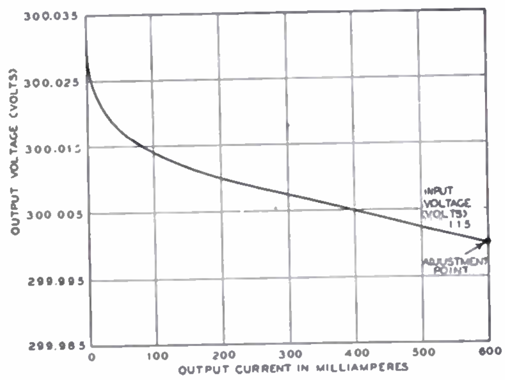

Fig. 5—Typical output voltage regulation curve.

3.2 Output Voltage Regulation

Using the same technique as before, adjust the regulated power supply to supply its rated terminal voltage and maximum current at normal input voltage. (Shown as the adjustment point in Fig. 5.)

The load current is reduced in steps sufficiently small so that a plot similar to that shown in Fig. 5 can be made.

3.3 Output Ripple Voltage

3.3.1 Connect the regulated power supply to supply current to a nonreactive load and connect the oscilloscope (provided with a suitable blocking capacitor) across the output terminals.

3.3.2 Vary the load current in steps throughout the normal range and measure the ripple voltage present across the output terminals in terms of peak-to-peak millivolts at each test point.

The results on the typical power supply are shown in Fig. 6.

2.1.8.3 The design of power transformers used in regulated power supplies is often governed not by losses but by physical volume and ventilation. A design which may be conservatively rated for use in open air may be unsatisfactory in a confined space. The safe operating temperature of a transformer—usually recommended as 55° C above surrounding air—may be exceeded due to its placement adjacent to and on the same mounting as power tubes used in the regulated power supply. Ventilation is likewise an important factor in the temperature rise of the components of a regulated power supply.

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The resistance for the calculation of temperature rise may be obtained from measurements made by any suitable method at the beginning and at the end of a rated power run. The rise in temperature can be calculated from the following formula:

$$R_t = R_T\,[1 + \alpha_T(t - T)]$$

$$(t - T) = \frac{R_t - R_T}{\alpha_T R_T}$$

in which

T = room temperature in degrees C

t = the calculated average temperature of the heated transformer winding

R_t = the measured hot resistance

R_T = the resistance measured at the beginning of the test, i.e., at room temperature, and α_T is the temperature coefficient of the copper conductors, taken as $1/234.5+T$ or $\alpha_T = 0.00393$ at 20° C.

A "hottest spot" correction of 10° C is to be added to this calculated average temperature rise.

2.2 Practical Measuring Devices

The circuit shown in Fig. 1 is for the measurement of output voltage stabilization and of regulation.

Fig. 2 is a reproduction of the appendix to footnote reference 1 in which a measurement technique for output impedance is detailed. The simple ac bridge which is suggested serves admirably for measuring internal impedance in the range of ohms.

For lower impedances in the orders of hundredths and thousandths of an ohm, the circuit shown in Fig. 3 has been found more useful.[2]

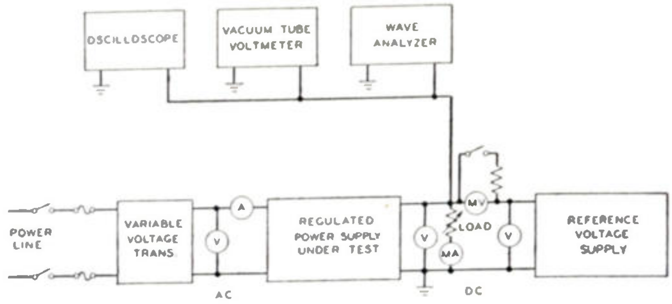

Fig. 1—Schematic of circuit employed for determining regulated power-supply characteristics.

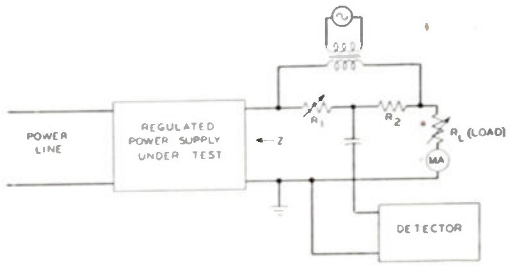

Fig. 2—Bridge circuit for measurement of output impedance of a regulated power supply. At balance $Z = R_1(R_L/R_2)$.

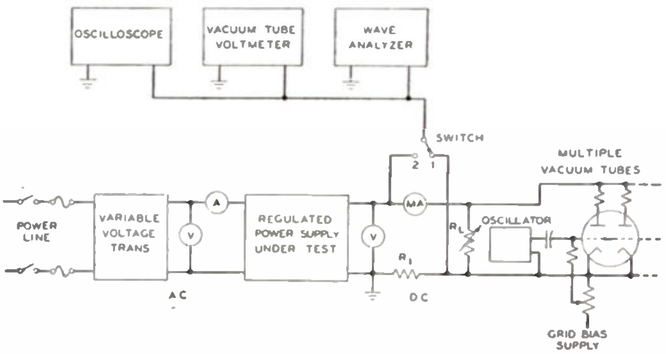

Fig. 3—Testing circuit for measuring output impedance of a regulated power supply.

2.3 Requirements of Measuring Equipment

The reference voltage supply shown in Fig. 1 supplies sufficient current to operate the millivoltmeter. A low internal-resistance regulated power supply operating through a line voltage regulator is to be preferred over a reference battery for this use, since the internal resistance of the series of cells may be of considerable value. A second regulated power supply similar to the one being tested and known to regulate small currents could be used for this reference voltage supply since the current demand from it for this test is very low.

For ripple voltage, regulation pull-out and output impedance measurements, a sensitive calibrated oscilloscope is necessary. Voltage drops produced by ripple currents and ac load currents flowing through the low output impedances of regulated power supplies intended

[3] *American Standard for Transformers, Regulators, and Reactors,* ASA C57.1-1942 and C57.2-1943.

or television use are in the millivolt range. The indicating oscilloscope requires ac amplification, internal or external, to produce sufficiently large deflections for accurate measurements.

For the output impedance measurement, a vacuum-tube load circuit is connected to the power supply as shown in Fig. 3. Applied to the control grids of the tubes is the output of an oscillator which may be adjusted in frequency and in amplitude. The oscillator should cover the range from the lowest frequency capable of being measured by the above oscilloscope to the frequency where the output capacitor of the regulated power supply determines the impedance range. This upper limit of frequency may be in the hundreds of kilocycles per second. The load current through the impedance of the power supply must be great enough to produce a measurable deflection on the oscilloscope. For an illustrative example, a current of 100 milliamperes rms through an impedance of 0.01 ohm will produce only 2.8 millivolts peak-to-peak. Shown as R_1 in Fig. 3 is a calibrated resistor whose value should be of the same order as that of the impedance being measured. Direct-current ammeter shunts are convenient to use for this purpose and are calibrated in resistance to an accuracy of 1 per cent and better. Their impedance at the top frequency to be measured can usually be assumed as equal to the dc resistance.

3. PROCEDURE

The following is a test procedure using, as an example, a typical regulated power supply furnishing 500 ma at 300 volts. **The voltages normally employed in regulated power supplies are sufficiently high to endanger human life. Every reasonable precaution should be taken by testing personnel to minimize the shock danger while adjustments and tests which demand normal operation of the power supply are being made.**

3.1 Output Voltage Stabilization

3.1.1 Connect the supply under test according to the diagram of Fig. 1, using a variable voltage transformer to set the input at exactly rated input voltage.

3.1.2 Adjust the reference voltage supply to equal the dc voltage output under open circuit before the meter switch is opened.

3.1.3 Open meter switch and carefully readjust reference voltage supply to obtain a midscale reading on millivoltmeter.

3.1.4 Vary the input voltage over an appropriate range centered on the rated input voltage, recording the output voltage as the sum of the millivoltmeter reading at each point and the reference voltmeter reading.

The same procedure is followed at dc load currents from zero to the maximum rated current.

The results can be plotted as shown in Fig. 4.

In making the above measurements, it is suggested that the necessary data be obtained as quickly as possible, since even a slight amount of drift is likely to affect the accuracy of measurement.

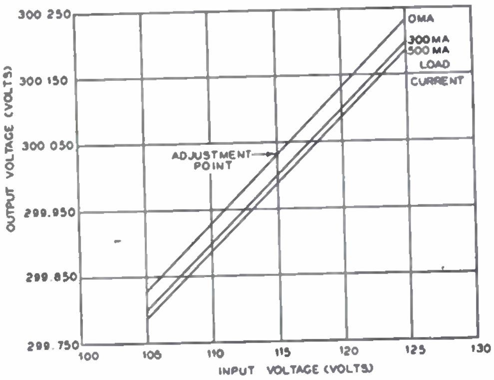

Fig. 4—Typical output voltage stabilization curves.

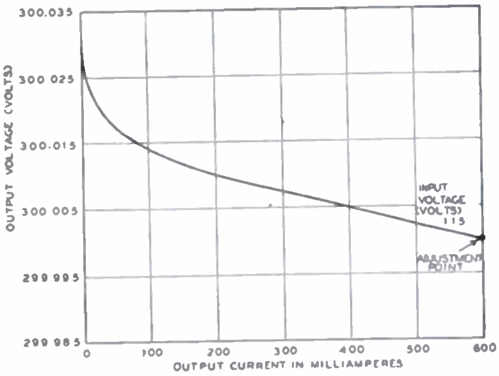

Fig. 5—Typical output voltage regulation curve.

3.2 Output Voltage Regulation

Using the same technique as before, adjust the regulated power supply to supply its rated terminal voltage and maximum current at normal input voltage. (Shown as the adjustment point in Fig. 5.)

The load current is reduced in steps sufficiently small so that a plot similar to that shown in Fig. 5 can be made.

3.3 Output Ripple Voltage

3.3.1 Connect the regulated power supply to supply current to a nonreactive load and connect the oscilloscope (provided with a suitable blocking capacitor) across the output terminals.

3.3.2 Vary the load current in steps throughout the normal range and measure the ripple voltage present across the output terminals in terms of peak-to-peak millivolts at each test point.

The results on the typical power supply are shown in Fig. 6.

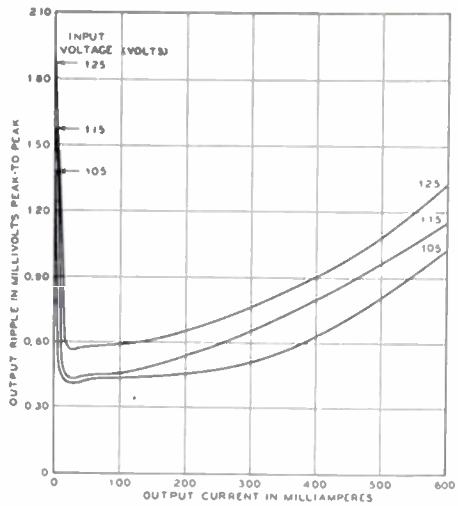

Fig. 6—Typical curves for output ripple voltage as a function of load current.

3.4 Regulation Pull-Out

The same connections as in paragraph 3.3 are used to measure this characteristic.

3.4.1 Set the load currents at several chosen values within the normal range of the unit, and increase the voltage regulating adjustment of the regulated power supply at each value to the point where the unit fails to regulate. This point is indicated on the oscilloscope by a sudden increase in ripple content.

3.4.2 Record the dc voltage and current at the point where the ripple content has increased to ten times the value measured in the regulating condition.

Since this test includes maximum load, care should be exercised so that none of the components of the unit is damaged.

The above procedure is carried through at input voltages of plus and minus 10 per cent from the nominal line voltage.

If the unit is suspected of not regulating under light loading, the above procedure can be carried through by decreasing the output voltage.

Plotted as shown in Fig. 7, the enclosed area defines the range of operation of the regulated power supply.

3.5 Magnitude of Output Impedance

3.5.1 The connections for the bridge measurements are shown in Fig. 2.

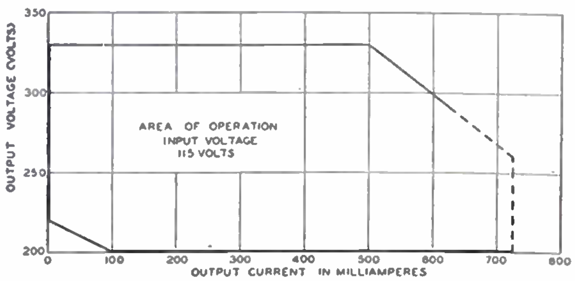

Fig. 7—Typical curve showing normal range of operation.

3.5.1.1 Balance the ac bridge circuit, shown in Fig. 2, by adjusting R_1 at each test frequency from the lowest frequency capable of measurement by the sensitive oscilloscope (about 20 cycles per second) to the frequency at which the capacitor across the output terminals determines the impedance range (from fifty to several hundred kc). Difficulty may be experienced in balancing this simple bridge, especially at the extreme low values of impedance which may be encountered. In this case, see Section 3.5.2.

3.5.1.2 At the balance point Z is $R_1 \times (R_L/R_2)$. By making R_2 equal to R_L the desired impedance can be read directly from the values of R_1 necessary to obtain a null in the detector circuit.

3.5.2 The procedure indicated in Fig. 3 for the measurement of output impedance is to measure the voltage drop across R_1 at the oscillator frequency by means of the sensitive oscilloscope or the wave analyzer. This value of voltage E_1 divided by the value of R_1 is the ac current I_1 supplied to the load circuit. Moving the switch from position 1 to position 2, the voltage E_2 at the same frequency is measured across the output terminals of the regulated power supply.

The output impedance of the unit at this frequency is

$$Z_f = \frac{E_2}{I_1} = \frac{E_2 R_1}{E_1}.$$

See Fig. 8 for a typical curve.

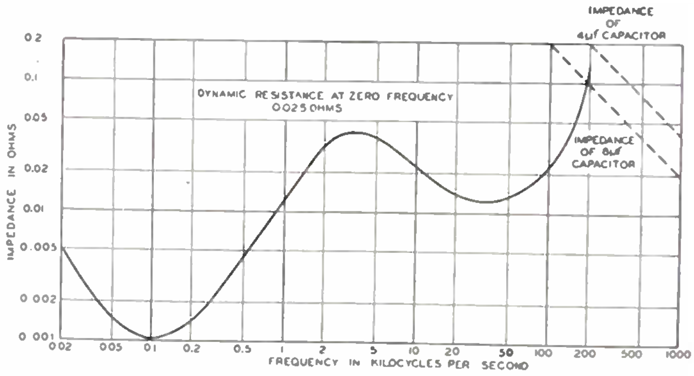

Fig. 8—Output impedance. (Input voltage, 115v, 60 cps; 300v, 570 ma dc output current; 56 ma peak-to-peak ac load current.)

3.6 Unregulated Voltage

To measure the unregulated voltage and the ripple voltage of the plate supply to the output tubes of the regulated power supply, it is necessary to connect the measuring instruments to the terminals of the rectifier filter. Caution should be exercised during this test since the potentials at this point may be several hundreds of volts higher than the rated terminal voltage of the power supply.

3.6.1 With the input line disconnected, connect a dc voltmeter and the oscilloscope across the filter output.

3.6.2 Adjust the input voltage and the output voltage to their rated values.

3.6.3 Increase in steps the current output of the regulated power supply and record the dc voltage and the ripple voltage as measured by the oscilloscope.

3.6.4 Plot the results as shown in Figs. 9 and 10.

From these data, the plate voltage across the output tube or tubes in multiple can be calculated and whether the output stage operates within its maximum ratings

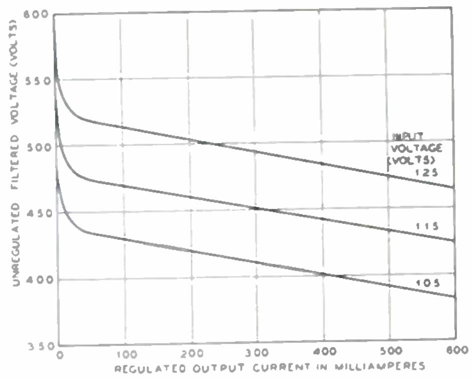

Fig. 9—Typical curves showing unregulated output voltage as a function of load current.

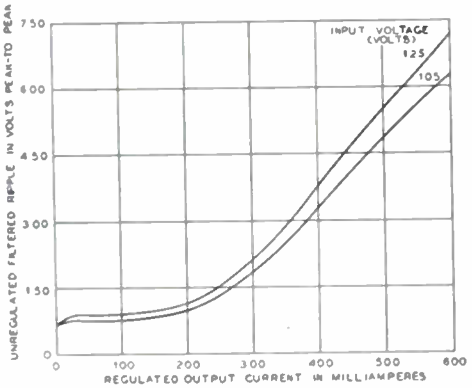

Fig. 10—Typical curves showing ripple voltage at unregulated output voltage terminals as a function of load current.

can be determined. The division of currents through several paralleled output tubes is not indicated by these results, but can only be determined by separate current measurements.

3.7 Regulated Power Supply Input Current and Efficiency

3.7.1 Measure the input current and input power for normal regulated output voltage and currents throughout the range by standard techniques.

3.7.2 Plot the results as shown in Fig. 11.

The efficiency is the ratio of the dc output power to the input power and will vary considerably over the range of output.

3.8 Miscellaneous

3.8.1 Warm-up Time

Measure the warm-up time with a stop watch at the

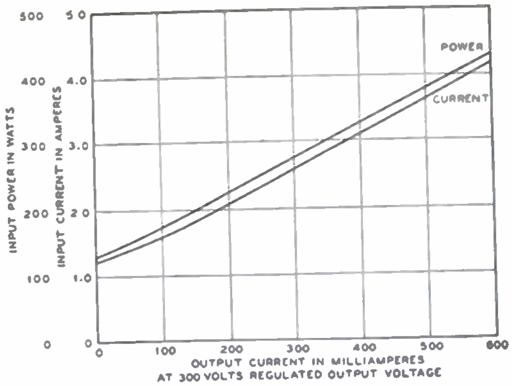

Fig. 11—Typical curves showing input power and current as functions of load current.

rated output voltage and maximum current of the regulated power supply.

3.8.1.1 Adjust the output voltage to normal, and the current to its rated maximum value.

3.8.1.2 Allow the power supply to cool to ambient temperature.

3.8.1.3 Switch on the input power and begin measuring time from that instant to the time when the unit is delivering regulated voltage at rated current.

3.8.2 Output Capacitor Discharge Time

3.8.2.1 Remove all the load circuits from the regulated power supply output terminals except the voltmeter. If an external voltmeter is used, its impedance should not be so low as to affect the measurement.

3.8.2.2 Adjust the output to normal voltage.

3.8.2.3 Switch off the input power and measure time from that instant to the time when the voltmeter reading has decreased to 50 volts.

3.8.3 Temperature Rise

3.8.3.1 Allow the power transformer to assume the ambient temperature. Measure the resistance of a winding of the power transformer, (the secondary windings are connected to the rectifier plate pins which are generally convenient) with a dc bridge or other convenient means.

3.8.3.2 Allow the regulated power supply to deliver rated output volt-amperes until its temperature becomes constant. Several hours will probably be required.

3.8.3.3 Turn off the power and again measure the resistance of the same winding as above.

3.8.3.4 Calculate the temperature rise as detailed in Section 2.1.8.3.

Example:

$R_T = 37.4$ ohms at 20° C

$R_t = 41.2$ ohms hot

$(t-T) = 3.8/0.00393 \times 37.4 = 3.8/0.147 = 25.8°$.

Adding the correction of 10° to 25.8° makes the sought-for value 35.8° C temperature rise.

A Multichannel PAM-FM Radio Telemetering System*

J. P. CHISHOLM†, E. F. BUCKLEY‡, MEMBER, IRE, AND G. W. FARNELL§

Summary—A multichannel time-division system of the PAM-FM type is described. It can provide as many as 64 channels with an information bandwidth per channel greater than 800 cps or fewer channels with a corresponding increase in the information bandwidth per channel. Modulation and multiplexing in the airborne equipment, and demultiplexing in the ground station, are performed by simple pentode gating circuits which derive their gating signals from crystal-resistor matrices. Transmission of dc signals over the radio-frequency link is avoided by a process of double modulation. The novel technique of reactance switching provides a simple FM transmitter with wide linear frequency deviation. The airborne portion of the system is miniaturized through the use of potted unit construction. In each channel, the over-all linearity is better than ±1 per cent, the cross talk is less than 1 per cent, and the signal-to-noise ratio is greater than 55 db.

Introduction

If the flight testing of pilotless aircraft is to be economically practical, many separate pieces of information must be obtained from each test flight. This information is obtained normally by radio telemetering. The electric outputs of various end instruments, such as pressure gauges and accelerometers, are combined for transmission over one radio link; at the receiving station, the combined information signal is separated into the individual information signals so that a continuous record of each measured quantity is obtained. In many applications, the airborne equipment must be extremely compact and light in weight.

The two basic methods of combining or multiplexing a number of electric signals into one composite signal are frequency division and time division. An example of the use of frequency division is in the relatively common frequency-modulation-frequency-modulation (FM-FM) system.[1] Experience has shown that in telemetering applications not more than about six information channels can be provided readily by this type of system. For this reason, frequency-division systems are not readily adaptable to the telemetering of a large number of information signals unless subcommutation in the time domain is employed. However, a direct method of time division such as the one discussed below permits the multiplexing of a very large number of high-capacity channels.

Several different methods of pulse modulation allow multiplexing to be performed in the time domain. The common methods include pulse-amplitude modulation (PAM), pulse-width modulation (PWM), pulse-position modulation (PPM), and pulse-code modulation (PCM). Moreover, the radio-frequency carrier may be modulated in amplitude or in frequency. Theoretical investigation has shown[2] that while PCM-AM and PCM-FM offer the best signal-to-noise ratios, PAM-FM is one of the better of the remaining systems. Since no known method of pulse-code modulation permits the design of an airborne multiplexing unit which is sufficiently simple and compact for use in small pilotless aircraft, attention has been centered on the development of a PAM-FM system.

The particular PAM-FM system described below is a direct time-division system which uses two eight-channel multiplexers to provide 16 information channels and 6,400 data samples per second per channel. The frequency response of each channel extends theoretically from zero to nearly 3,200 cps. The system may be altered to provide 32 channels or 64 channels with frequency response extending to 1,600 cps or 800 cps, respectively. The information-handling capacity of the system is determined primarily by the total number of data samples per second, which in the above case is $16 \times 6{,}400 = 102{,}400$. The data samples may be allotted equally to 2, 4, 8, 16, 32, . . . channels as may be desired, and the corresponding theoretical information bandwidth in cycles per second is half the number of data samples per second per channel.[3] The airborne portion of the 16-channel telemeter described herein uses approximately 1.7 tubes per channel, and the entire equipment, including power supplies, but exclusive of end instruments, is packaged in a volume less than 200 cubic inches. The major components of this PAM-FM telemetering system are discussed in the following order: airborne multiplexer, ground-station demultiplexer, synchronizing circuits, channel demodulators, and FM transmitter.

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[1] J. C. Coe, "Telemetering guided-missile performance," Proc. I.R.E., vol. 36, pp. 1404–1414; November, 1948.

[2] V. D. Landon, "Theoretical analysis of various systems of multiplex transmission," *RCA Rev.*, vol. 9, pp. 287–351, June, 1948; and pp. 433–482, September, 1948.

[3] W. R. Bennett, "Time division multiplex systems," *Bell. Sys. Tech. Jour.*, vol. 20, pp. 199–221; April, 1941.

Airborne Multiplexer

Pulse-amplitude modulation and time-division multiplexing are performed simultaneously in the airborne multiplexer unit.[4] Assuming that the wave form shown in Fig. 1(a) represents the signal voltage from one of the end instruments, the function of the telemetering system is to reproduce this wave form at the receiving terminal on the ground. Theoretically, only two samples per cycle of the highest-frequency component of this signal need be transmitted in order that the wave form can be reproduced identically at the ground station.[3] However, in practice, a minimum of about 2.1 samples per cycle is required. To perform the sampling, the signal voltage is applied to the control grid of a pentode gating tube which has the suppressor grid biased beyond plate-current cutoff. During intervals δ at times T, $2T$, $3T$ . . . (see Fig. 1(b)), a positive gating pulse of fixed amplitude is applied to the suppressor grid. The resulting plate wave form is a series of negative pulses

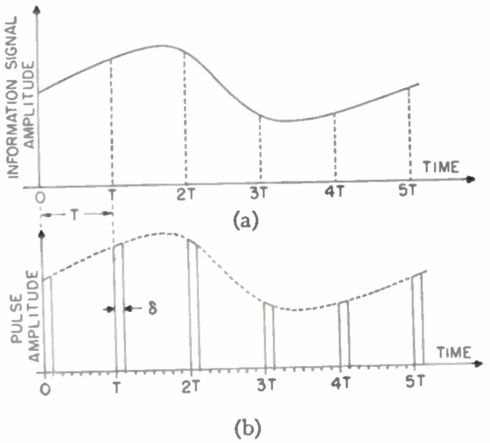

Fig. 1—Pulse-amplitude modulation in a single channel.

with amplitudes proportional to the corresponding instantaneous values of the signal voltage. Thus, a short sample of the signal voltage is taken once during each frame period T. These so-called channel pulses are shown inverted in Fig. 1(b). During the remainder of each frame period, other signals may be sampled successively for similar short intervals δ.

The circuit which performs the simultaneous modulation and multiplexing for an eight-channel system is shown in Fig. 2. One gating tube is used in each of the eight information channels, with the eight information signals applied to the eight control grids. A common load resistor R_L is used for all eight tubes, and positive gating pulses are applied in sequence to the suppressor grids. The suppressor grids of the pentodes used in this application have sharp cutoff characteristics. The voltage across the common load resistor becomes proportional to each of the information signals in turn as the corresponding tube is gated "on." The wave form of this composite signal is shown inverted in Fig. 3, wherein seven channels contain useful information and the remaining channel contains a synchronizing pulse of characteristically high amplitude for use in the receiving equipment. This illustration shows two complete frame periods with certain specific signal variations. This ideal wave form could be transmitted only by a system of infinite bandwidth; in the actual system, any one information pulse does not occupy all of the interval δ allotted to it.

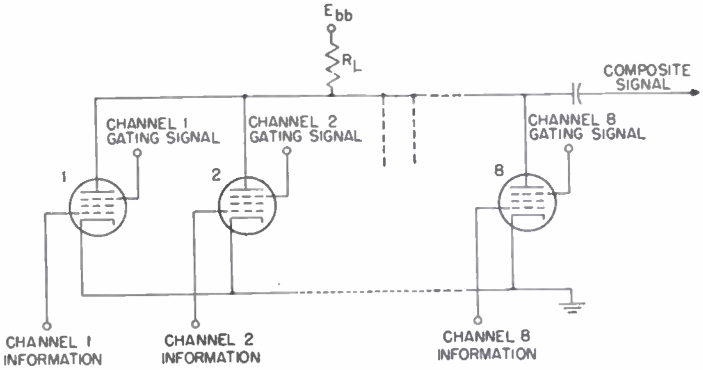

Fig. 2—Representative 8-channel multiplexer. Gating signals are obtained from crystal-resistor matrix. Pentodes are 6AS6 or CK5784.

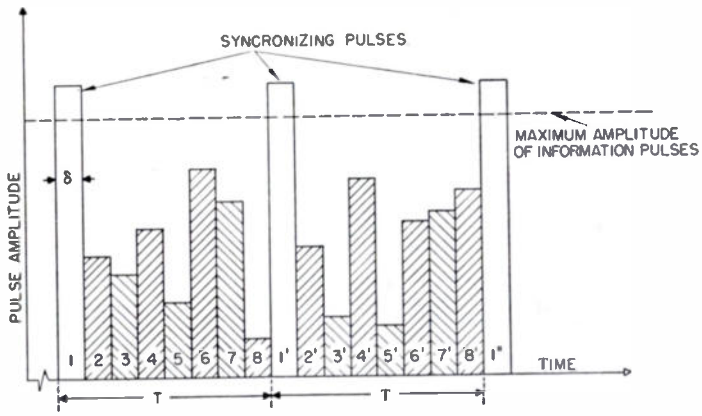

Fig. 3—Composite output of 8-channel multiplexer.

The block diagram of the airborne portion of a 16-channel system is shown in Fig. 4. The two 8-channel multiplexers are interconnected by means of an auxiliary electronic commutator to provide 16 channels. The composite output of each 8-channel multiplexer is selected during alternate periods T to modulate the FM transmitter. Since alternate samples of each information signal are omitted at the output of the commutator, the 16-channel system has a frame period of $2T$. During each of these frames, the 15 information signals and the one synchronizing signal modulate the transmitter in turn. Since there are 102,400 useful data samples per second, the duration of each channel interval δ is slightly less than 10 microseconds.

[4] L. L. Kilpatrick, "A high-capacity matrix-commutated telemetering system," Master's Thesis, Massachusetts Institute of Technology, 1948.

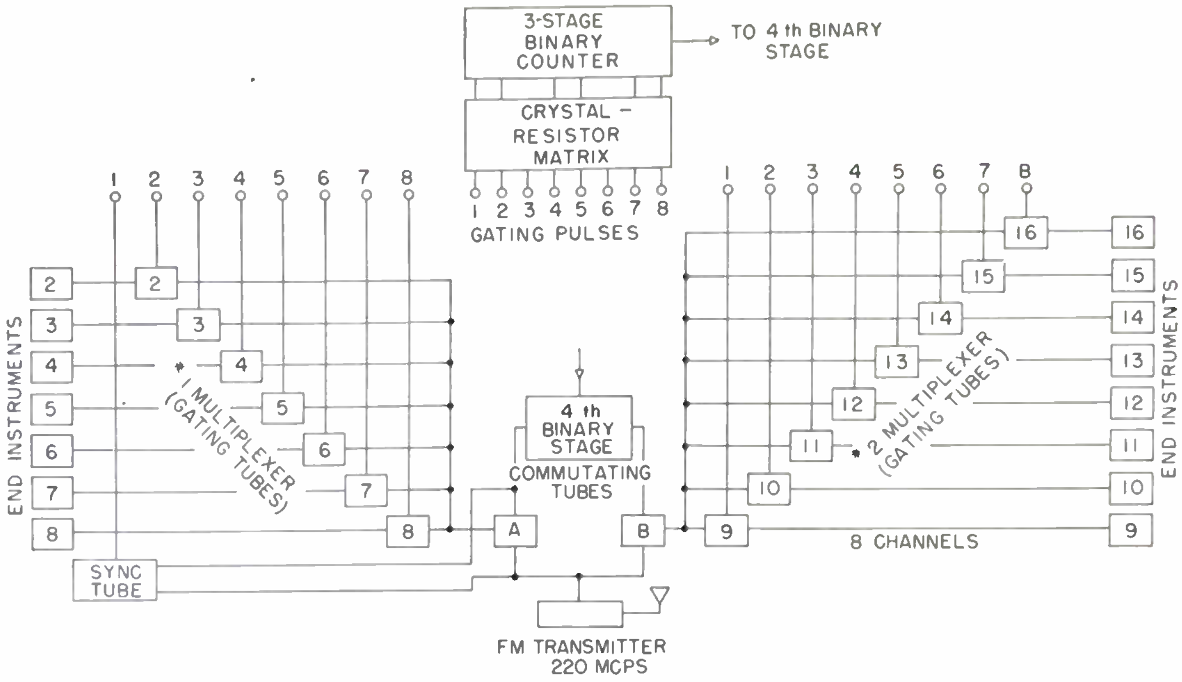

Fig. 4—Block diagram of 16-channel PAM-FM airborne unit.

In the airborne unit, the sequential gating or switching pulses are obtained from a crystal-resistor matrix which is operated by a three-stage binary counter. The binary counter and crystal-resistor matrix are illustrated in detail in Fig. 5(a). Each stage of the counter is a conventional flip-flop circuit, and isolating cathode followers connect each plate in the flip-flop circuits to one of the vertical lines of the matrix. The significant voltage wave forms encountered in the matrix are shown in Fig. 5(b). Assume first that the crystal diodes are not connected. During the first interval δ, section (a) of each double triode is conducting and the cathode of each conducting section is at a potential of V volts with respect to ground; hence, the open-circuited voltage of line 1 with respect to ground is V volts. Since the cathode voltage of the nonconducting section of each cathode-follower tube is essentially zero, it may be seen that the voltage of lines 2, 3, and 5 is $2V/3$ volts, that of lines 4, 6, and 7 is $V/3$ volts, while that of line 8 is zero. Line 1 is called the selected line because its voltage is greater than that of any other line, and the minimum differential of $V/3$ volts could be used as the gating signal for channel 1. However, if the crystal diodes are used, each line except line 1 is connected to the cathode of a nonconducting triode through the forward resistance of at least one crystal diode. Thus the selected line, line 1 in this case, is at V volts as before, whereas the remaining lines are all at essentially ground potential. The fact that only two output voltage levels are possible with the diodes in the circuit makes the adjustment of the bias voltage of the gating tubes much less critical.

At the end of the first interval δ a negative triggering pulse causes the first stage of the counter to assume its alternate stable state. Thus, section (b) of the first double triode becomes the conducting section. Examination of the circuit shows that line 2 is now the selected line and that all the other lines, including line 1, are at ground potential. As succeeding triggering pulses actuate the counter, lines 3 through 8 are selected in turn as shown by the wave forms of Fig. 5(b), and the sequence repeats continuously. These eight pulse trains generated by the matrix provide the switching pulses which are applied to the suppressor grids of the gating tubes in Fig. 2.

If an n-stage counter is used, the crystal-resistor matrix may be extended to provide 2^n sequential gating pulses. However, the number of elements required is $n \cdot 2^n$, where the parallel combination of a resistor and a crystal diode is considered as a single matrix element. Thus, while 24 elements are required for an 8-position matrix, a 16-position matrix requires 64 elements, and a 32-position unit uses 160 elements. The reason for using two 8-channel multiplexers in the airborne equipment in order to obtain 16 channels is to limit the physical size of the matrix.

For the sake of continuity, the discussion of the radio-frequency link will be postponed until the function of the receiving-station demultiplexer has been outlined.

Receiving Terminal

The first function of the receiving equipment at the ground station is to recover a composite signal which is identical with that modulating the transmitter. This signal is decomposed into amplitude-modulated pulse trains which are replicas of those associated with the sampling of the original information signals. Finally,

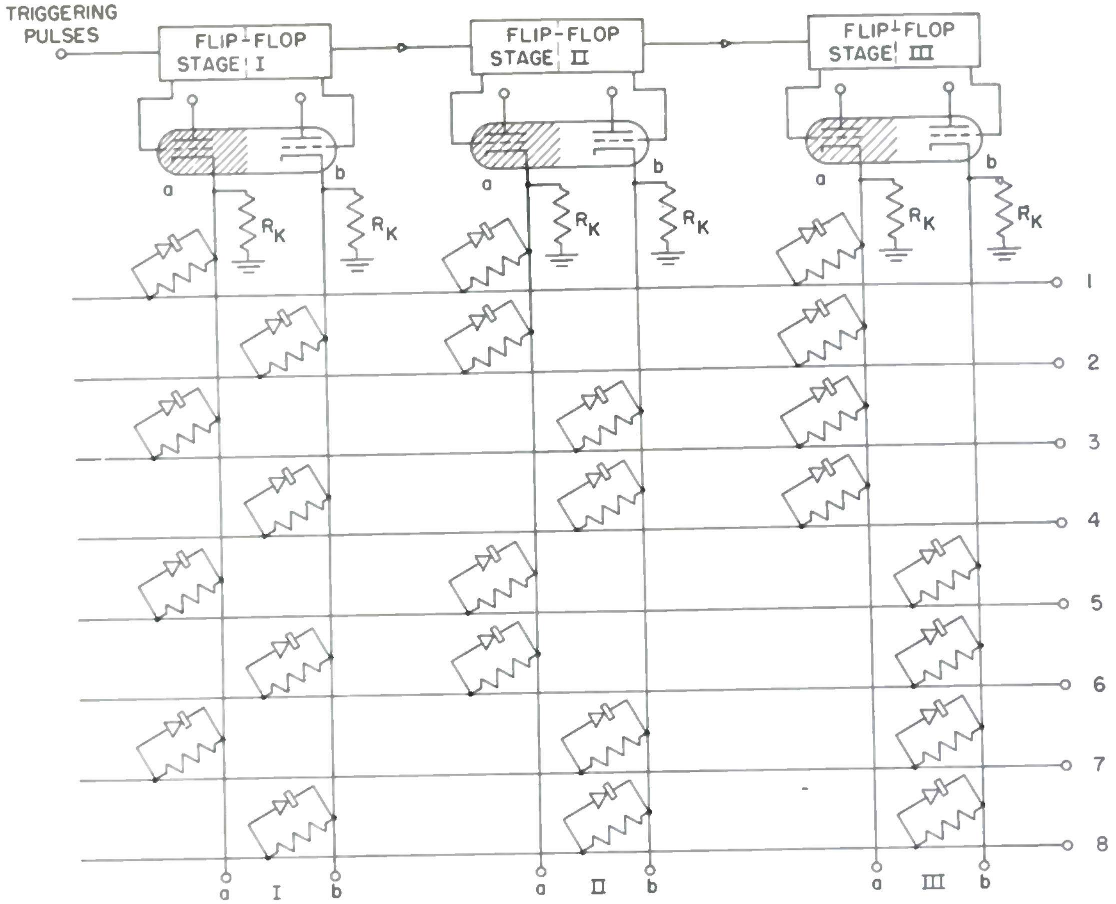

Fig. 5(a)—Matrix-type pulse generator. Circuit diagram. (See voltage wave forms in Fig. 5(b).)

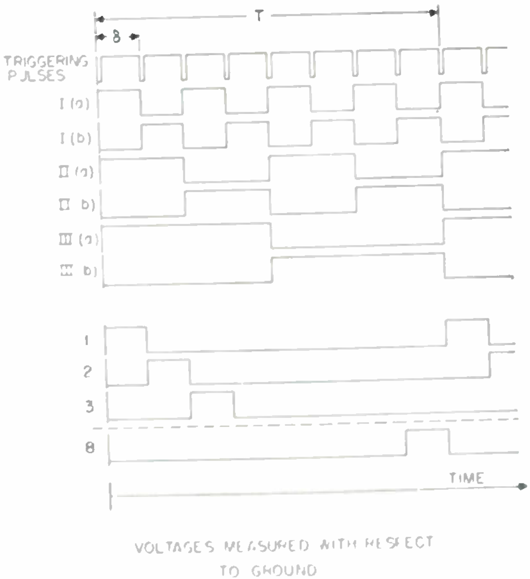

Fig. 5(b)—Matrix-type pulse generator. Voltage wave forms.

demodulator circuits reproduce the wave forms of the end-instrument variations.

The output of the discriminator of the FM receiver shown in the block diagram of Fig. 6 is a 16-channel composite signal similar to the 8-channel signal of Fig. 3. The discriminator output is connected to the control grid of each of the 16 gating tubes in the demultiplexer. The gating tubes are pentodes similar to those employed in the airborne multiplexer. However, in the ground station, each tube has a separate load resistor, whereas in the airborne unit all the gating tubes have a common load resistor. Sequential gating pulses which coincide in time with the corresponding pulses in the airborne unit are applied to the suppressor grids of the demultiplexer gating tubes. Thus, each gating tube produces an output only during the short interval that its suppressor grid is gated "on"; but its output is the required amplitude-modulated pulse train representing the sampled information in the corresponding channel.

A 16-position crystal-resistor matrix is used to generate the sequential gating pulses in the receiving terminal because size is not a critical factor. Obviously, the "on" interval of the demultiplexer gating tube in a given channel must coincide exactly with the "on" interval of the multiplexer gating tube in that channel. Not only must the gating pulses at the ground station occur in exact synchronism with the pulses in the airborne unit, but also they must be correctly indexed. This phase and frequency control is performed by the phase-locked pulse-repetition-frequency multiplier shown in Fig. 6. The circuit is controlled by the synchronizing pulses which are extracted easily from the composite signal because they are larger in amplitude

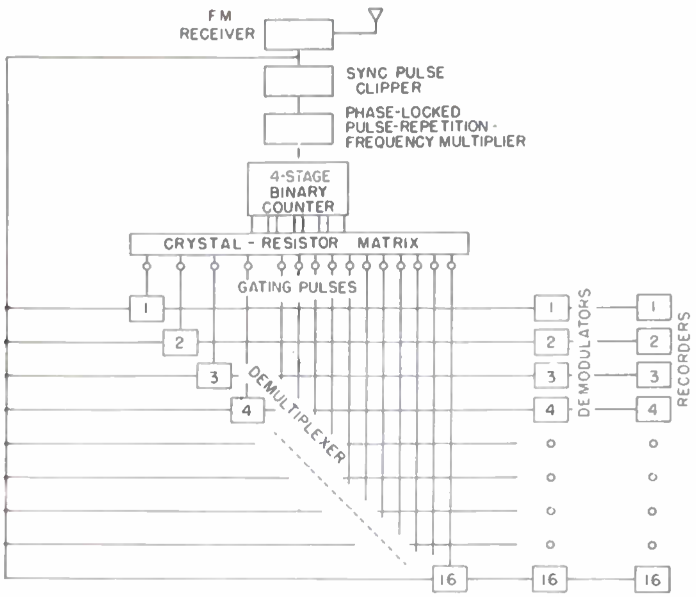

Fig. 6—Block diagram of 16-channel PAM-FM ground station.

than any of the information pulses. The frequency multiplier is essentially a gated multivibrator operating at 16 times the recurrence frequency of the synchronizing pulses. The differentiated output of this multivibrator is a train of triggering pulses which operates the four-stage binary counter. The multivibrator is stopped at the end of alternate frames (a frame is the time interval between successive synchronizing pulses) and is restarted each time by the succeeding synchronizing pulse. This pulse also resets the 16-position binary counter to position 1. Thus the demultiplexer is properly indexed at the beginning of every second frame. The interval between synchronizing pulses is compared in a time-discriminator with the interval occupied by the 16 cycles of the multivibrator which follow each indexing operation, and the output of the discriminator adjusts the grid-bias voltages of the multivibrator so that its frequency is just 16 times the repetition frequency of the synchronizing pulses. Since the frequency of the multivibrator automatically follows changes in the synchronizing frequency, the master oscillator which establishes the basic channel-switching rate in the airborne unit need not be exceedingly stable.

The lower wave form of Fig. 1 shows the samples of a typical information signal which are recovered at the output of one of the gating tubes in the ground station. If this train of pulses were passed through an ideal low-pass filter which cut off at exactly one half of the sampling frequency, information signals containing frequency components up to half of the sampling frequency could be handled by the system. Since practical filters do not have infinitely sharp cutoff characteristics, the frequency response of each channel is slightly less than one half of the sampling frequency. In applications where no low-frequency signals are encountered, the channel demodulators may be simple low-pass filters. However, in the direct pulse-amplitude-modulation process discussed earlier, all of the frequency components present in the data wave forms appear in the composite, or commutated, signal. This condition represents a decided disadvantage in telemetering applications because much of the information is in the form of slowly varying dc voltages. In order to transmit these low-frequency components, the system would have to contain dc amplifiers and the center frequency of the FM transmitter would have to be carefully stabilized; that is, the information would be carried in the absolute amplitude of the pulses, and a reference level would be required.

In order to eliminate the complications of dc transmission, the double-modulation technique illustrated in Fig. 7 is used. The information signal to be transmitted in each channel modulates a square-wave car-

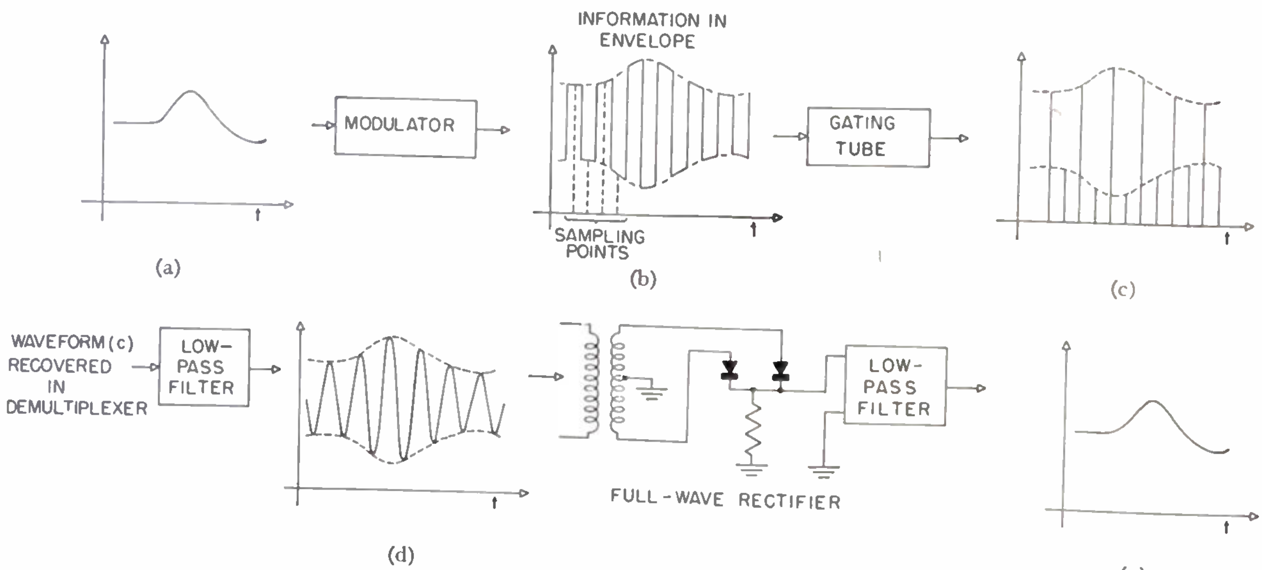

Fig. 7—Double modulation for elimination of dc reference level. (a) Information signal. (b) Modulated square-wave carrier. (c) Sampled carrier. (d) Modulated sine-wave carrier.

rier of frequency equal to exactly one half of the sampling frequency. For many types of end instrument, such as the strain-gauge bridge or the microtorque potentiometer, this modulation can be performed directly by using the square-wave carrier to excite the instrument. The end-instrument output, which is of the form shown in Fig. 7(b), is sampled in the normal manner by a gating tube. It should be noted that because the carrier wave and the gating pulses are synchronized, the modulated carrier is sampled alternately at the top and at the bottom as shown in Fig. 7(b) and (c). Thus, in the sampled wave form of one channel, the information is contained in the difference in amplitude of successive pulses, and the absolute amplitude of the pulses is unimportant. In this double-modulation system, dc information signals do not produce dc components in the composite wave form.

In the ground station, a wave form similar to that of Fig. 7(c) is obtained at the plate of one of the gating tubes of the demultiplexer. Demodulation is performed in two steps. The output of each gating tube is passed through a low-pass filter which cuts off just below the sampling frequency, and the amplitude-modulated sine wave of Fig. 7(d) appears at the filter output. A full-wave rectifier and a low-pass filter which cuts off just below the carrier frequency are used in the arrangement shown in Fig. 7 to recover the information signal. This double-modulation technique has been used successfully to reproduce wave forms which contain frequency components up to 70 per cent of the carrier frequency. The information signals from the demodulators are recorded on a multichannel oscillograph.

Reactance-Switch FM Transmitter

The little-known technique of reactance-switching[5–7] is employed in the FM transmitter in order to obtain wide linear frequency deviation and to permit modulation directly at the carrier frequency. A wide linear frequency deviation is desirable so that the information-handling capacity of the system shall not be limited by transmitter characteristics. Modulation directly at the carrier frequency is desirable to minimize power requirements and circuit complexity.

A reactance-switch modulator differs from a reactance-tube modulator in that the latter controls the frequency of an oscillator by varying the effective reactance connected continuously across the tuned circuit, while the reactance-switch modulator varies the time during which a fixed reactance is connected across the tuned circuit. Fig. 8(a) is a simplified schematic diagram of the reactance-switch modulator. An auxiliary inductor L_s, a crystal diode, a bias battery, and the source of modulating voltage are connected in series across the tuned circuit of the oscillator. Let v_m in Fig. 8(a) be the sum of the bias voltage and the modulating voltage. As long as the instantaneous value of the voltage across the tuned circuit is less than v_m, the diode is nonconducting and the angular frequency of the oscil-

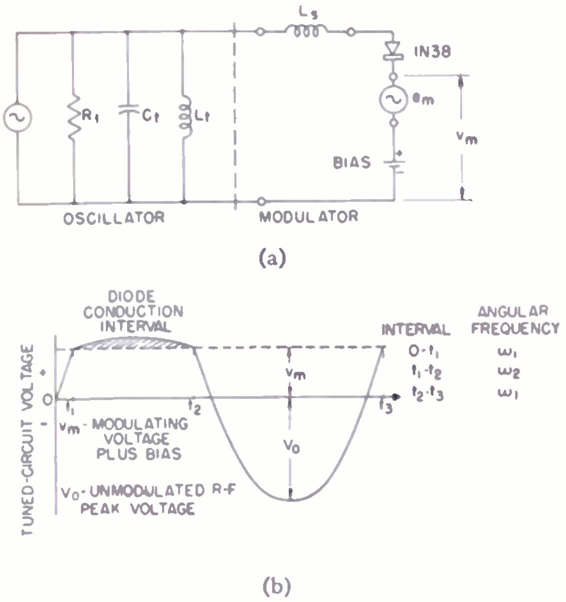

Fig. 8—Reactance-switch modulator. (a) Simplified schematic diagram. (b) Wave form of tuned-circuit voltage.

lator ω_1 is determined by C_t and L_t. During the time interval between t_1 and t_2, the tuned-circuit voltage exceeds v_m, the diode conducts, the inductor L_s is connected across L_t, and the instantaneous angular frequency is changed to a value ω_2. Thus the instantaneous frequency alternates between two values during each radio-frequency cycle. The resulting fundamental frequency of the tuned-circuit voltage depends upon the conduction interval of the crystal diode, and therefore the frequency deviation is a function of the modulating voltage e_m. This function can be made approximately linear over a frequency range of ± 1 per cent of the carrier frequency; moreover, the modulation is performed directly at the allocated carrier frequency of 220 Mc. Fig. 9 shows the linearity of frequency deviation and the uniformity of power output of a typical transmitter.

Performance

While the PAM-FM telemetering system is still in the developmental stage, the results of field tests have been encouraging. The information-handling capacity per unit volume is greater than that of other existing systems in the telemetering field. The versatility of this particular multiplexing technique should be noted. The 16-channel system described above transmits 102,400 data samples per second, thus providing an information bandwidth of about 3,000 cps per channel. However, an

[5] J. W. Kearney, "Reactance-switching: a method of producing wide-band frequency modulation," NDRC-Div. 15, Technical Memorandum 411-TM-90, (PB 14020), Radio Research Laboratory, Harvard University, 1944.

[6] D. W. Cottle, "Frequency modulation by reactance-switching," Master's Thesis, Massachusetts Institute of Technology, 1948.

[7] Staff of the Radio Research Laboratory, Harvard University, "Very High-Frequency Techniques" vol. I, pp. 422–425, McGraw-Hill Book Co., New York, N. Y.; 1947.

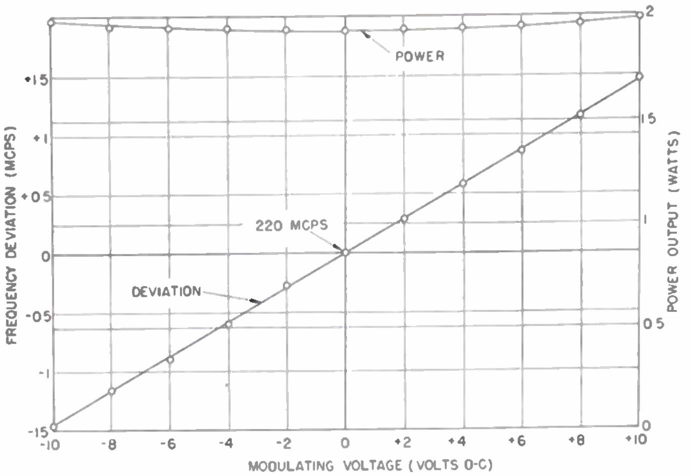

Fig. 9—Modulation characteristics of reactance-switch transmitter.

increase in the information bandwidth to, say, 6,000 cps entails only the doubling of the master-oscillator frequency and the appropriate increase in the cutoff frequency of the low-pass filters in the ground station. Moreover, as noted previously, a given number of data samples per second can be subdivided in many different ways. Because the matrix switch which provides the gating pulses is operated by a binary counter, the number of channels is most conveniently a power of 2. Since the number of gating pulses is 2^n, where n is the number of binary stages, the possible number of channels increases much more rapidly than the number of stages. Therefore, the number of tubes per channel decreases as the number of channels increases; for example, the 16-channel system requires 1.7 tubes per channel in the airborne unit, whereas a 64-channel system requires fewer than 1.4 tubes per channel.

In telemetering systems, the output signal of each channel should be a linear function of the corresponding input signal in order to facilitate data reduction. The over-all linearity of one channel of the 16-channel PAM-FM system including the radio-frequency link is shown in Fig. 10. The end instrument used in taking

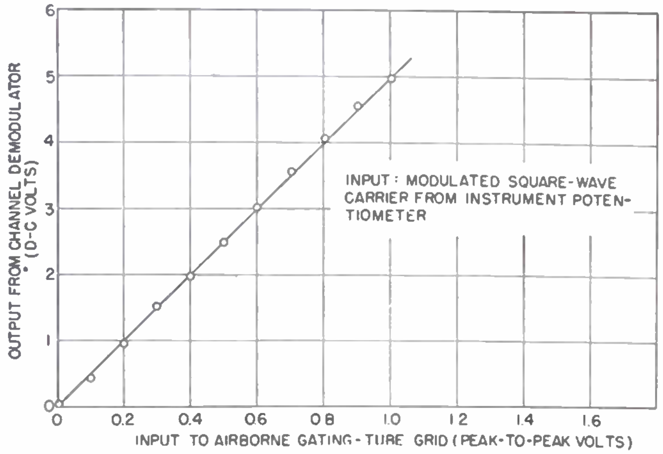

Fig. 10—Over-all linearity of one channel of 16-channel PAM-FM system.

these measurements was a potentiometer excited by the half-sampling-frequency square-wave carrier. The potentiometer output was applied directly to one of the airborne gating tubes and the system response was measured at the output of the corresponding demodulator. It can be seen that the departures from linearity are less than 1 per cent of full scale. Full-scale modulation is produced by a signal of 1-volt peak-to-peak applied to the grid of a gating tube.

A type-AN/APR-4 radio receiver was modified for use with this system. A radio-frequency signal of 5 μv is required to keep the demultiplexer synchronized, and as long as the system remains in synchronism, the signal-to-noise ratio measured at the output of any channel is about 55 db. An appreciable percentage of the noise is introduced by the switching and frequency-control circuits. Complete shielding would reduce this component of the noise somewhat, but it is extremely difficult to provide adequate shielding in the airborne equipment where volume and weight must be minimized.

Nonlinearities in the radio-frequency link do not introduce cross talk between channels in time-division systems. However, cross talk is introduced because finite radio-frequency bandwidth does not permit the transmission of ideal rectangular pulses. Rise and fall times are prolonged so that in the demultiplexer, energy from any one pulse appears in adjacent portions of the time intervals allotted to adjacent channels. To minimize this cross talk, the gating pulses used in the demultiplexer are appreciably shortened in such a manner that they occupy only the middle portion of each channel interval of the composite signal. The other major source of cross talk is pickup due to incomplete shielding. When the radio-frequency bandwidth is approximately 1 Mc and a full-scale signal modulates one channel of the 16-channel system, the cross talk in adjacent channels is down more than 54 db; when identical full-scale signals modulate all but one channel, the cross talk in that one channel is down more than 48 db.

It is common practice in telemetering to calibrate each channel immediately before flight and sometimes during flight. Thus, the accuracy of the PAM-FM system depends upon the accuracy of calibration, upon the errors due to noise and upon drifts in gain between calibrations. The errors due to noise can be made negligible and the drifts in gain are dependent largely on the stability of the power supplies used. If the supplies are regulated, the errors due to drift can be kept within 1 per cent of full scale. However, in telemetering applications, it is often necessary to use poorly regulated battery supplies because of size and weight limitations. The airborne multiplexer of the 16-channel PAM-FM system requires 6 amperes of filament current at 6 volts, and about 100 milliamperes of plate current at 135 volts. The total volume of the airborne power supply is less than 100 cubic inches.

The photograph in Fig. 11 illustrates the compact tructure of the airborne equipment; it does not show he associated end instruments or the power supplies. The potted unit in the lower foreground is the 8-position

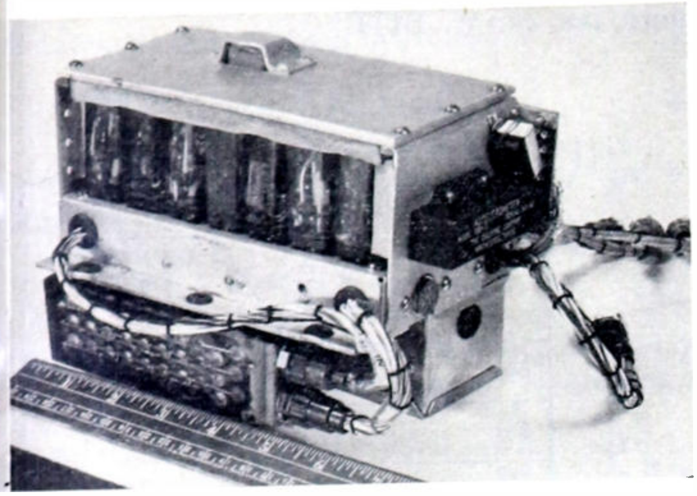

Fig. 11—Airborne equipment for 16 channels, exclusive of power supplies and end instruments.

crystal-resistor matrix, while the small box in the rear contains the reactance-switch transmitter. The main unit contains the master oscillator and the counting and gating circuits. Each vacuum tube with its associated circuit components is embedded in a potting compound and these subassemblies are mounted in miniature sockets. The potted construction enables the telemetering equipment to operate under very high accelerations. The complete unit shown in Fig. 11 operates satisfactorily under sinusoidal acceleration of amplitude 15 g and of frequency 50 cps. It will withstand constant acceleration in excess of 50 g along any axis for short periods. The weight of this unit is about 7 pounds and the volume is about 110 cubic inches.

The ground-station equipment of the 16-channel PAM-FM system appears in Fig. 12. This photograph does not include the power supplies or the recording apparatus. The relay rack contains the modified AN/APR-4 receiver, the synchronizing, demultiplexing, and demodulating circuits, and the dc amplifiers required to operate a multichannel recording oscillograph.

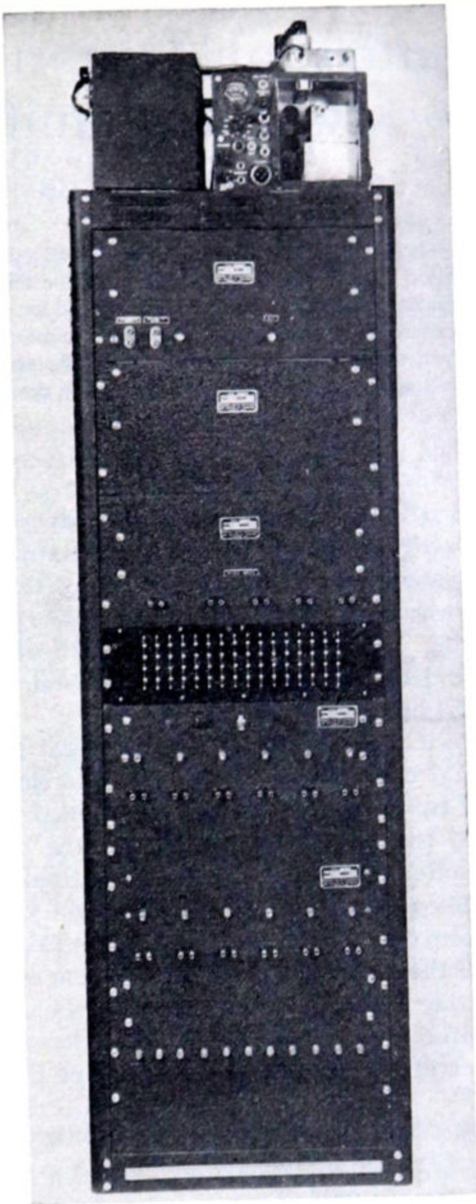

Fig. 12—Ground station for 16 channels, exclusive of power supplies and recording equipment.

Conclusions

The PAM-FM telemetering system described above has the following noteworthy features. It is extremely versatile in that relatively few circuit modifications are required to change either the number of channels or the maximum frequency response per channel. Pulse-amplitude modulation and time-division multiplexing are performed by a simple set of pentode gating tubes supplied with the appropriate gating pulses from a compact crystal-resistor matrix. A process of double modulation is employed to avoid the transmission of dc signals over the radio-frequency link. In the transmitter, wide-band frequency modulation directly at the carrier frequency is facilitated by the novel technique of reactance-switching. Potted unit construction provides a compact, rugged, and easily serviced airborne unit. The potential information-handling capacity of this PAM-FM telemeter is greater than that of other existing telemetering systems and the wide information bandwidth of each channel makes the circuitry applicable to multichannel speech-communication systems.

Acknowledgments

The authors are grateful for the encouragement and technical advice offered by W. H. Radford. They wish to acknowledge also the work of L. L. Kilpatrick in the development of the multiplexing circuits, and the work of D. W. Cottle in the development of the reactance-switch transmitter.

Quantization Distortion in Pulse-Count Modulation with Nonuniform Spacing of Levels*

P. F. PANTER†, SENIOR MEMBER, IRE, AND W. DITE†

Summary—**It is shown that the distortion introduced in a pulse-count-modulation system due to quantization can be minimized by nonuniform spacing of levels. Equations are derived for an arrangement of nonuniform level spacing that produces minimum distortion. It is also shown that minimum distortion is significantly less than distortion resulting from uniform quantization when the crest factor of the signal is greater than four.**

I. Introduction

It has been shown in a previous paper[1] that the process of quantization in a pulse-count-modulation system introduces distortion. In that paper, equations were derived for the special case of equal-level spacing. This paper studies the effect of nonuniform-level spacing with the view of obtaining minimum distortion.

It is shown that by taking the statistical properties of the signal into consideration, the distortion can be minimized by a proper level distribution, which is a function of the probability density of the signal. In practice, nonuniform quantization is realized by compression followed by uniform quantization. The most common form of compression is the so-called logarithmic one, where the levels are crowded near the origin and spaced farther apart near the peaks. It is shown that with logarithmic compression the distortion is largely independent of the statistical properties of the signal.

II. Distortion for Nonuniform Spacing

Consider a quantized signal $y(t)$ as shown in Fig. 1. Let the levels be symmetrically disposed about zero voltage level, but otherwise placed in an arbitrary manner in the interval $(-V, V)$. The problem of minimizing the distortion of the quantized signal by properly spacing the levels will be considered now. Denote the levels by $y_{-n}, y_{-n+1}, \cdots, y_0, y_1, \cdots y_{n-1}, y_n$, where $y_k = -y_{-k}$ and $y_0 = 0$. Further, assume that a signal value of y which satisfies

$$y_{k-1/2} < y < y_{k+1/2} \tag{1}$$

is transmitted as level y_k. The y's with fractional subscripts are the crossover values.

The measure of distortion that is adopted is the mean-square-distortion voltage defined as

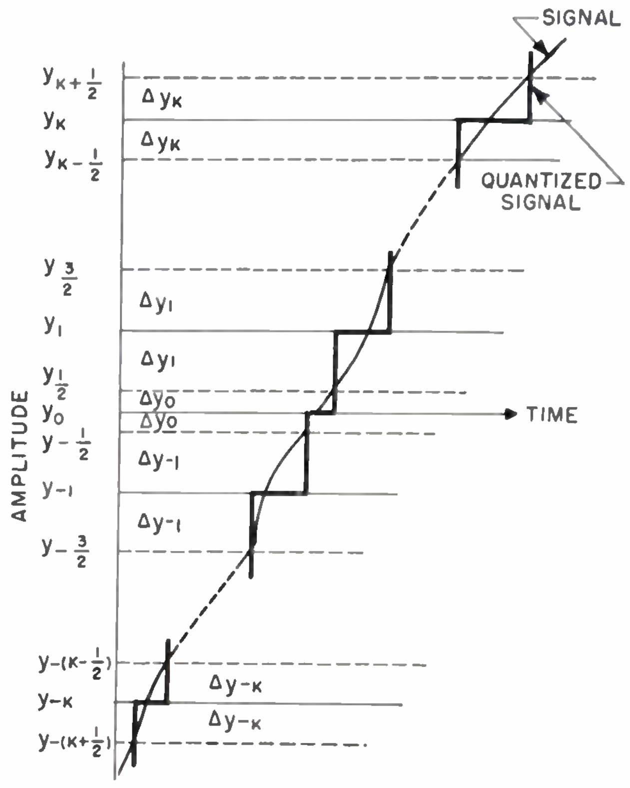

Fig. 1—Nonuniform quantization.

$$\sigma_k = \int_{y_{k-1/2}}^{y_{k+1/2}} (y - y_k)^2 P(y) dy, \tag{2}$$

where σ_k is the distortion in the kth level. The error of the transmitted signal is $(y - y_k)$ and $P(y)$ the probability density of the signal y.

First, derive a relation between the various y's in (2) so that σ_k in the kth level is a minimum.

Suppose that the levels are so close that $P(y)$ may be considered as nearly constant over the region of integration and equal to $P(y_{av})$ where

$$y_{av} = \frac{y_{k+1/2} + y_{k-1/2}}{2};$$

$$\sigma_k = \frac{P(y_{av})}{3}\left[(y_{k+1/2} - y_k)^3 + (y_k - y_{k-1/2})^3\right]. \tag{3}$$

Differentiating σ_k with respect to y_k gives

$$\frac{d\sigma_k}{dy_k} = P(y_{av})\left[-(y_{k+1/2} - y_k)^2 + (y_k - y_{k-1/2})^2\right] = 0$$

or

$$y_k = \frac{y_{k+1/2} + y_{k-1/2}}{2} = y_{av}. \tag{4}$$

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

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

Thus, the condition for making σ_k a minimum is that y_k lie half-way between $y_{k+1/2}$ and $y_k - {}_{1/2}$.

$$\left.\begin{aligned} y_{k+1/2} &= y_k + \Delta y_k, \\ y_{k-1/2} &= y_k - \Delta y_k. \end{aligned}\right\} \tag{5}$$

Substituting these values in (3), it follows

$$\sigma_k = \frac{2}{3} P(y_k)\Delta y_k^3. \tag{6}$$

The total mean-square-distortion voltage is found by summing over all levels, giving

$$\sigma_d = \frac{2}{3} \sum_{-n}^{n} P(y_k)\Delta y_k^3. \tag{7}$$

Now, it will be proved that the mean-square-distortion σ_d is a minimum when σ_k is constant, independent of the kth level.

By the definition of an integral, we may write

$$2\sum_{-n}^{n} P^{1/3}(y_k)\Delta y_k = \int_{-V}^{V} P^{1/3}(y)dy = 2K, \tag{8}$$

where K is a constant, since the integral is a function of only its limits. Let $\mu_k = P^{1/3}(y_k)\Delta y_k$; then (7) and (8) become, respectively,

$$\sigma_d = \frac{2}{3} \sum_{-n}^{n} \mu_k^3, \tag{9}$$

and

$$K = \sum_{-n}^{n} \mu_k. \tag{10}$$

The problem is now reduced to minimizing the sum of cubes subject to the condition that the sum of the variables is a constant.

From Lagrange's method of undetermined multipliers, it follows that (9) is a minimum when

$$\mu_{-n} = \mu_{-n+1} = \cdots = \mu_{n-1} = \mu_n = \frac{K}{2\mu + 1}. \tag{11}$$

From this, it follows

$$P^{1/3}(y_k)\Delta y_k = \frac{K}{2n+1} \tag{12}$$

or

$$\sigma_k = \frac{2}{3} \frac{K^3}{(2n+1)^3}.$$

The total minimum distortion power

$$\sigma_m = \frac{2}{3} \frac{K^3}{(2n+1)^2} = \frac{1}{12(2n+1)^2}\left(\int_{-V}^{V} P^{1/3}(y)dy\right)^3. \tag{13}$$

Since $P(y)$ is an even function and letting $N = 2n+1$ be the total number of steps,

$$\sigma_m = \frac{2}{3N^2}\left(\int_0^V P^{1/3}(y)dy\right)^3. \tag{14}$$

The ratio of the mean-square-distortion voltage to the mean-square-signal voltage is

$$D_m^2 = \frac{\sigma_m}{\sigma} = \frac{2}{3N^2} \frac{\left[\int_0^V P^{1/3}(y)dy\right]^3}{\int_0^V y^2 P(y)dy}, \tag{15}$$

Equation (15) gives the minimum distortion resulting with optimum level spacing.[2]

An approximate method of obtaining levels may be obtained by writing (where k is positive)

$$\begin{aligned} y_k &= \Delta y_0 + 2\Delta y_1 + \cdots + 2\Delta y_{k-1} + \Delta y_k \\ &= \frac{K}{N}\left[\frac{1}{P^{1/3}(y_0)} + \frac{2}{P^{1/3}(y_1)} + \cdots + \frac{2}{P^{1/3}(y_{k-1})} + \frac{1}{P(y_k)}\right] \\ &= \frac{K}{2V}\left[\frac{1}{P^{1/3}(y_0)} + \frac{2}{P^{1/3}(y_1)} + \cdots + \frac{2}{P^{1/3}(y_{k-1})} + \frac{1}{P(y_k)}\right]\frac{2V}{N}. \end{aligned} \tag{16}$$

The series may be approximated by an integral

$$y = A\int_0^z \frac{1}{P^{1/3}(z)}dz, \tag{17}$$

where we have changed the variable on the right to z to avoid confusion. A is a constant of proportionality so chosen that when $z = V$; $y = V$. Hence

$$y = V \frac{\int_0^z \frac{1}{P^{1/3}(z)}dz}{\int_0^V \frac{1}{P^{1/3}(z)}dz}. \tag{18}$$

By letting z vary from 0 to V, y will describe a curve as shown in Fig. 2. As z takes on the values of $z_0 = 0$, $z_1 = 2V/N \cdots z_k = 2kV/N \cdots z_n = 2nV/N$, we get the point $y_0 = 0$, $y_1 \cdots y_k \cdots y_n$.

While approximate, this derivation has the advantage that the levels are obtained quickly. The relation (12) may be used to make spot checks since

$$\begin{aligned} y_{k+1} &= y_k + \Delta y_k + \Delta y_{k+1} \\ &= y_k + \frac{K}{N}\left(\frac{1}{P^{1/3}(y_k)} + \frac{1}{P^{1/3}(y_{k+1})}\right). \end{aligned} \tag{19}$$

[2] This equation was first derived by P. R. Aigrain using a slightly different method.

From Fig. 2, it may be seen that the uniform spacing along the z axis is transformed into a nonuniform spacing along the y axis. The compressed output z is then subjected to uniform quantization. To recapitulate, nonuniform spacing of levels may be realized by passing the signal through a compressor with a given characteristic and applying uniform quantization to its output. Obviously, this also implies that a corresponding expansion is incorporated in the receiver.

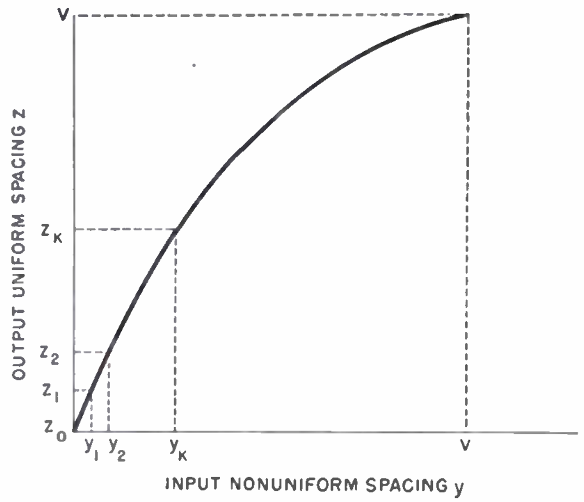

Fig. 2—Compression curve for nonuniform-level spacing.

$$y_k = \Delta y_0 + 2\Delta y_1 + \cdots + 2\Delta y_{k-1} + \Delta y_k$$

$$= \frac{K}{2V}\left[\frac{1}{P^{1/3}(y_0)} + \frac{2}{P^{1/3}(y_1)} + \cdots + \frac{2}{P^{1/3}(y_{k-1})} + \frac{1}{P(y_k)}\right]\frac{2V}{N},$$

$$\therefore \quad y \doteq A\int_0^z \frac{1}{P^{1/3}(z)}\,dz$$

$$= \frac{\int_0^z \frac{1}{P^{1/3}(z)}\,dz}{\int_0^V \frac{1}{P^{1/3}(z)}\,dz},$$

where

$z = V$ and $y = V$.

III. Logarithmic Compression

A compression curve that is relatively easy to obtain and has been used in practice is the so-called logarithmic compression curve. The positive half of the compression characteristic is given by

$$v_2 = k\log\left(1 + \frac{\mu v_1}{V}\right), \tag{20}$$

where

μ = compression parameter
v_1 = input voltage
v_2 = output voltage
V = maximum input voltage
k = an undetermined constant.

To find k, let $v_2 = V$ when $v_1 = V$, so that the maximum values of the input and compressed waves are equal. This gives

$$v_2 = \frac{V}{\log(1+\mu)}\log\left(1 + \frac{\mu v_1}{V}\right). \tag{21}$$

The parameter μ controls the degree of compression and may be chosen so that large changes in the input produce relatively small changes in the output. As shown in Fig. 3 for $\mu = 1{,}000$, a 60-db change in the input will cause only a 20-db change in the output. When μ is large, the levels are crowded about zero, $\mu = 0$ corresponds to uniform spacing.

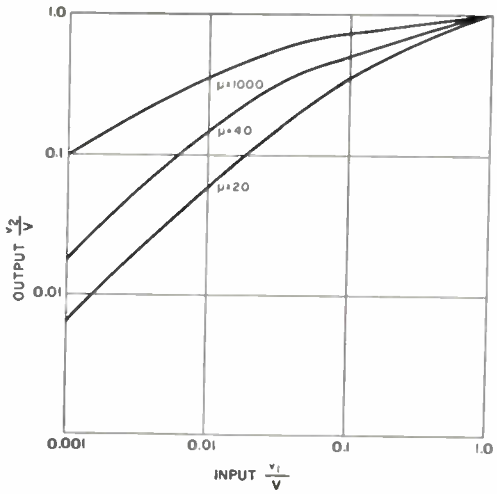

Fig. 3—Logarithmic compression.

$$v_2 = \frac{V}{\log(1+\mu)}\log\left(1 + \frac{\mu v_1}{V}\right).$$

Differentiating, (21) yields

$$dv_2 = \frac{\mu}{\log(1+\mu)}\,\frac{V}{V + \mu v_1}\,dv_1. \tag{22}$$

As the compressed signal is divided into N uniformly spaced levels, $\Delta v_2 = 2V/N$. This gives

$$\Delta v_1 = \frac{2\log(1+\mu)}{\mu N}(V + \mu v_1). \tag{23}$$

In the notation of the previous section, $\Delta y_k \doteq \Delta v_1/2$ so that

$$\Delta y_k = \frac{\log(1+\mu)}{\mu N}(V + \mu y_k) = \alpha(V + \mu y_k), \tag{24}$$

where

$$\alpha = \frac{\log(1+\mu)}{\mu N}. \tag{25}$$

From (23), it follows that the ratio of the largest to the smallest level is given approximately by

$$\frac{\Delta y_n}{\Delta y_0} \doteq 1 + \mu. \tag{26}$$

It is interesting to note that

$$\frac{(dv_2/dv_1)_{v_1=0}}{(dv_2/dv_1)_{v_1=v}} \text{ is also equal to } 1+\mu.$$

This ratio, when expressed in decibels, is often referred to as the compression of the system.

The distortion power is given approximately by (7) when the number of levels is large

$$\sigma_d = \frac{2}{3}\sum_{-n}^{n} P(y_k)\Delta y_k^3$$

$$= \frac{2\alpha^2}{3}\sum_{-n}^{n} (V+\mu y_k)^2 P(y_k)\Delta y_k$$

$$= \frac{2\alpha^2}{3}\sum_{-n}^{n} [V^2P(y_k) + 2V\mu y_k P(y_k) + \mu^2 y_k^2 P(y_k)]\Delta y_k. \quad (27)$$

Since $P(y_k)$ is an even function and $y_k = -y_k$, the second term in the summation vanishes. Thus,

$$\sigma_d = \frac{2\alpha^2}{3}\sum_{-n}^{n} [V^2P(y_k) + \mu^2 y_k^2 P(y_k)]\Delta y_k.$$

Converting this into an integral with $2\Delta y_k = dy$ and using (41) and (46) of the Appendix yields

$$\sigma_d = \frac{\alpha^2 V^2}{3}\int_{-V}^{V} P(y)dy + \frac{\alpha^2\mu^2}{3}\int_{-V}^{V} y^2 P(y)dy$$

$$= \frac{\alpha^2 V^2}{3} + \frac{\alpha^2\mu^2\sigma}{3}, \quad (28)$$

where σ is the mean-square-voltage and V is the peak value of the signal. Hence, the distortion is given by

$$D = \left(\frac{\sigma_d}{\sigma}\right)^{1/2} = \frac{\log(1+\mu)}{3^{1/2}\mu N}(C^2+\mu^2)^{1/2}, \quad (29)$$

when C is the ratio of peak to root-mean-square value of signal. For a given C, the distortion is a function of μ only and will be a minimum for optimum μ as shown in Fig. 4.

When μ is large compared to C, this becomes

$$D = \frac{\log(1+\mu)}{3^{1/2}N}. \quad (30)$$

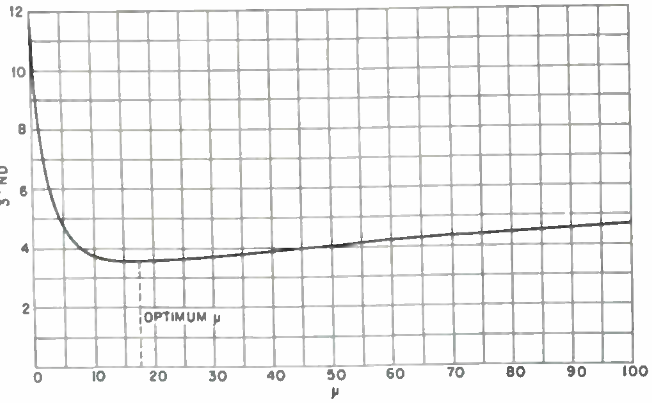

Fig. 4—Distortion plotted against the parameter μ.

$$C = 12 \quad \text{and} \quad 3^{1/2}ND = \log(1+\mu)\left(1+\frac{C^2}{\mu^2}\right)^{1/2}$$

Hence, when μ is large, the distortion is largely independent of the signal.

The distortion for uniform quantization may be obtained by letting μ become zero in (29).

$$D = C/3^{1/2}N. \quad (31)$$

For a sine wave,

$$C = 2^{1/2}$$

and

$$D = \frac{2^{1/2}}{3^{1/2}N} = \frac{2}{6^{1/2}N}.$$

When a signal is uniformly quantized, the distortion depends on the ratio of the peak to the root-mean-square value. This is illustrated in curve 1 of Fig. 5; curve 2 gives the minimum distortion for optimum logarithmic compression when the compression parameter μ is selected by curve 3.

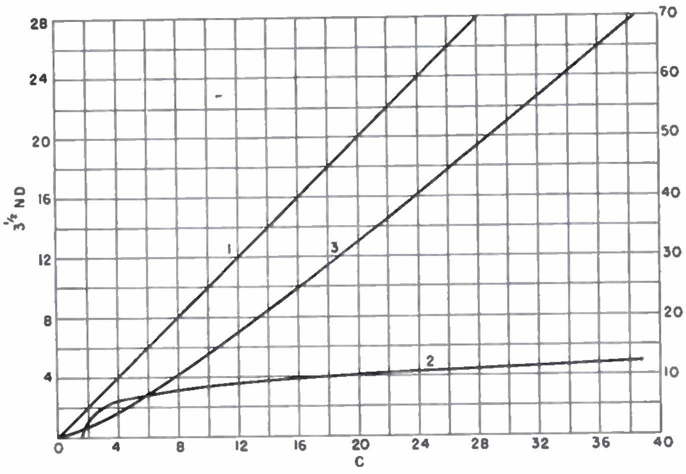

Fig. 5—Distortion characteristics. Curve 1 is for uniform quantization $3^{1/2}ND = C$. Curve 2 is for optimum logarithmic compression.

$$3^{1/2}ND = \log(1+\mu)\left(1+\frac{C^2}{\mu^2}\right)^{1/2}.$$

Curve 3 is for the case where

$$C = \left[\frac{\mu^2}{\frac{1+\mu}{\mu}\log(1+\mu) - 1}\right]^{1/2}.$$

IV. Application to Specific Signals

To illustrate the various points involved, let the preceding results be applied to the class of signals specified by the probability density given by

$$P(y) = \frac{1}{(3+2\lambda)^{1/2}\sigma^{1/2}\beta(1/2,\lambda+1)}\left(1-\frac{y^2}{(3+2\lambda)\sigma}\right)^{\lambda}, \quad (32)$$

which is discussed in the Appendix. Here λ is a parameter that determines the ratio of the peak value to the root-mean-square value. For $-1<\lambda<0$, the signal is relatively flat, while for $0<\lambda<\infty$, the signal has sharp peaks. When $\lambda=0$, all values are equally probable. If we let

$$A = \frac{1}{\sigma^{1/2}(3+2\lambda)^{1/2}\beta(1/2,\lambda+1)},$$

$$a = \sigma^{1/2}(\beta+2\lambda)^{1/2},$$

we find that the minimum distortion is given by

$$D_m^2(\lambda) = \frac{2A}{3N^2\sigma}\left[\int_0^a\left(1-\frac{y^2}{a^2}\right)^{\lambda/3}dy\right]^3 = \frac{Aa^3}{12N^2\sigma}\left[B\left(1/2, 1+\frac{\lambda}{3}\right)\right]^3 = \frac{2\lambda+3}{12N^2}\,\frac{B^3(1/2,1+\lambda/3)}{B(1/2,1+\lambda)}. \quad (33)$$

In this case

$$C = (2\lambda+3)^{1/2}. \quad (34)$$

The distortion for uniform-level spacing is from (31)

$$D = \frac{1}{3^{1/2}N}(3+2\lambda)^{1/2}. \quad (35)$$

The quantity ND is plotted versus C in Fig. 6. It is seen that for small values of C, there is little advantage in using the minimum spacing. It is not until the crest factor is above 6 that the difference between the two distortion figures is 6 db. The distortion for optimum spacing attains a final value of about $1.65/N$. Thus, the chief advantage of the optimum spacing occurs when the crest factor of the signal is high.

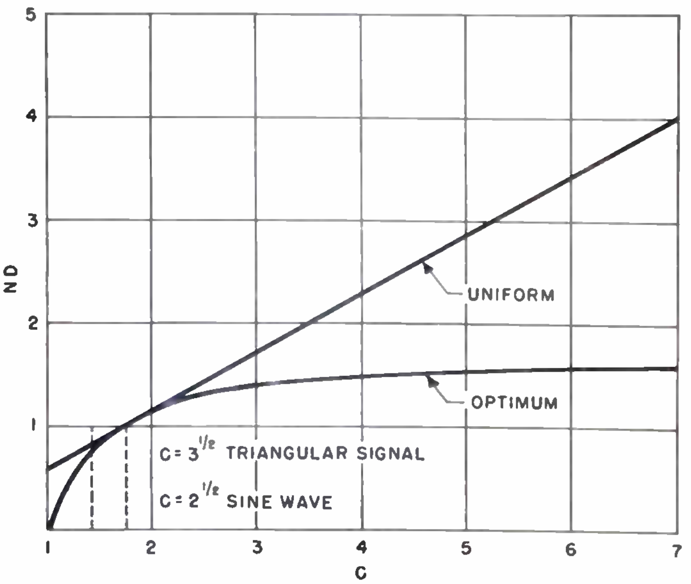

Fig. 6—Distortion plotted against crest factor for the Pearson distribution.

V. Appendix, Statistical Properties of the Signal

Pearson Distribution

The probability density given by

$$P(y) = \frac{1}{(3+2\lambda)^{1/2}\sigma^{1/2}B(1/2,\lambda+1)}\left(1-\frac{y^2}{(3+2\lambda)\sigma}\right)^{\lambda} \quad (36)$$

is known as the Pearson distribution. In (36), σ is power and λ is a parameter satisfying $-1<\lambda<\infty$. Further, the variable y is restricted to the interval

$$-[(3+2\lambda)\sigma]^{1/2} \leqq y \leqq +[(3+2\lambda)\sigma]^{1/2}.$$

Finally, $B(m, n)$ is the beta function.

For special values of λ, $P(y)$ reduces to simple forms. We have for $\lambda=0$

$$P(y) = \frac{1}{2(3\sigma)^{1/2}}, \qquad -(3\sigma)^{1/2} < y < (3\sigma)^{1/2},$$

which is a rectangular distribution with a peak-to-peak amplitude of $2\,(3\sigma)^{1/2}$. The case $\lambda=-1/2$ reduces to the distribution function of a sine wave whose maximum value is $A=(2\sigma)^{1/2}$. If $\lambda=-1$, we find that

$$P(y) = \infty$$

when $y=\pm\sigma^{1/2}$,

$$P(y) = 0$$

elsewhere.

The distribution consists of impulses at $y=\pm\sigma^{1/2}$ of strength 1/2; the signal is a square wave. When $\lambda\to\infty$, the probability density reduces to the normal distribution.

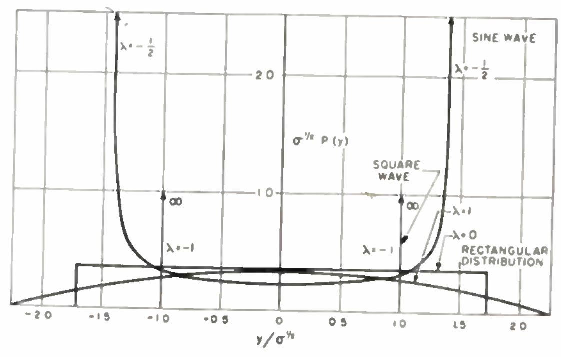

Fig. 7—Pearson distribution curves of $\sigma^{1/2}P(y)$ plotted against $y/\sigma^{1/2}$.

The probability density (36) is plotted in Fig. 7 for several values of λ with constant power. Actually $\sigma^{1/2}P(y)$ is plotted against $y/\sigma^{1/2}$ to get a set of curves applicable to all values of σ. From these curves, it is seen that $\lambda=0$ is the dividing line between two classes of signal. If λ is negative, the signals tend to concentrate near the maximum value. For λ positive, values near zero are emphasized.

Radiation from Wide-Angle Conical Antennas Fed by a Coaxial Line*

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Summary—**An approximate expression for the radiation from spherically capped conical antennas is derived by the Fourier-Lamé eigen-function method. Radiation patterns have been calculated for antennas with flare angle of $\pi/6$ and various lengths.**

I. Introduction

IN A PREVIOUS PAPER[1] the input impedances of spherically capped conical antennas fed by a coaxial line were calculated for large flare angles. By an extension of this earlier work an expression for the electromagnetic field of an antenna of arbitrary length a and flare angle θ_0 (see Fig. 1) is obtained. It is supposed that the characteristic impedance of the coaxial line

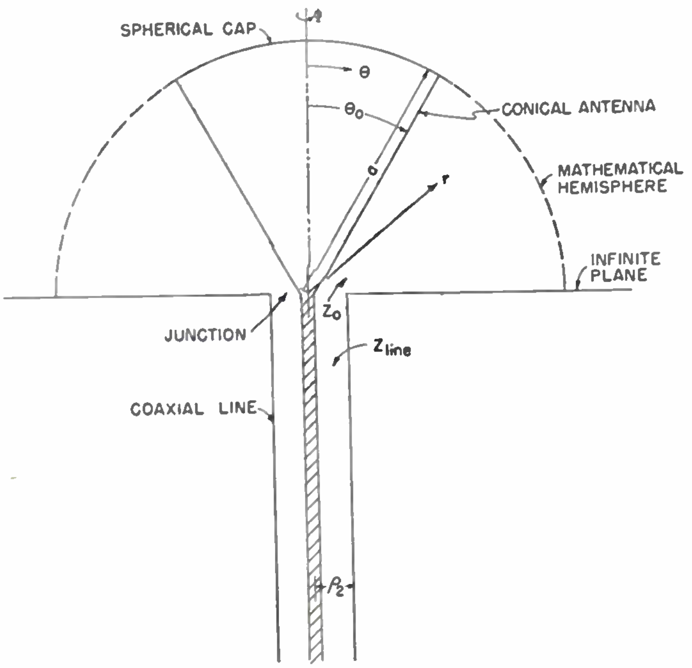

Fig. 1—Spherically capped conical antenna fed by coaxial feed line. Characteristic impedance of line equals characteristic impedance of antenna and flare angle is large, i.e., $\theta_0 \geqq \pi/6$.

$$Z_{\text{line}} = 60 \ln \frac{\rho_2}{\rho_1},$$

where ρ_2 and ρ_1 are respectively the radii of its outer and inner conductors, is equal to the characteristic impedance of the antenna

$$Z_0 = 60 \ln \cot \frac{\theta_0}{2}.$$

From symmetry considerations it is known that the only nonvanishing components of the electromagnetic field in the antenna region $\rho_2 \leqq r \leqq a$, $\theta_0 \leqq \theta \leqq \pi/2$, and the radiation region $r \geqq a$, $0 \leqq \theta \leqq \pi/2$ are H_ϕ, E_θ, E_r. Assuming that only the *TEM* mode exists in the antenna region, and expanding the field in the radiation region in an infinite series of eigen-functions, the radiation field is determined by forcing the annihilation of E_θ over the spherical cap and the continuity of E_θ over the mathematical surface $r=a$, $\theta_0 \leqq \theta \leqq \pi/2$.

II. Formulation of the Solution

In the antenna region for large flare angles, i.e., $\theta_0 \geqq \pi/6$, only the *TEM* mode need be considered. The θ component of the electric field for this *TEM* mode is given by the following expression:[2]

$$E_\theta(r, \theta) = \sqrt{\frac{\mu}{\epsilon}} \frac{\alpha e^{-ikr} - \beta e^{ikr}}{r \sin \theta} \tag{1}$$

where α and β are constants, $k = 2\pi \div$ free-space wavelength.

In the radiation region, it is easily shown that[2]

$$E_\theta(r, \theta) = i\sqrt{\frac{\mu}{\epsilon}} \sum_{n=1}^{\infty} A_n \left[h_{n-1}^{(2)}(kr) - \frac{n}{kr} h_n^{(2)}(kr) \right] P_n^1(\cos \theta), \tag{2}$$

where A_n's are constants, $h_n^{(2)}(x)$ are the spherical Hankel functions of the second kind, $P_n^1(\cos \theta) = -d/d\theta P_n (\cos \theta)$ are the associated Legendre polynomials, and the summation is over odd integers.

Since $E_\theta(a, \theta)$ must vanish over the spherical cap, from (2) it follows that

$$0 = \sum_{n=1}^{\infty} A_n \left(h_{n-1}^{(2)}(ka) - \frac{n}{ka} h_n^{(2)}(ka) \right) P_n^1(\cos \theta) \tag{3}$$

for $0 \leqq \theta \leqq \theta_0$. To match E_θ across the mouth of the antenna, i.e., across the mathematical surface, $r=a$, $\theta_0 \leqq \theta \leqq \pi/2$, it is necessary to satisfy the relation

$$\frac{\alpha e^{-ika} - \beta e^{ika}}{a \sin \theta} = i \sum_{n=1}^{\infty} A_n \left(h_{n-1}^{(2)}(ka) - \frac{n}{ka} h_n^{(2)}(ka) \right) P_n^1(\cos \theta) \tag{4}$$

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[1] C. H. Papas and R. King, "Input impedance of wide-angle conical antennas fed by a coaxial line," Proc. I.R.E., vol. 37, pp. 1269–1271; November, 1949.

[2] C. H. Papas and R. King, "Input Impedance of Wide-Angle Conical Antennas," Cruft Laboratory Technical Report No. 52, December 1, 1948.

obtained by equating (1) and (2) for $r=a$. Multiplying both sides of (3) and (4) by $P_n^1(\cos\theta)\sin\theta$ and integrating with respect to θ over their respective intervals of validity, it follows from the orthogonality property of the associated Legendre polynomials that

$$\left(\frac{\alpha e^{-ika}-\beta e^{ika}}{a}\right)P_n(\cos\theta_0) = iA_n\left[h_{n-1}^{(2)}(ka)-\frac{n}{ka}h_n^{(2)}(ka)\right]\frac{n(n+1)}{2n+1}. \qquad (5)$$

Solving (5) for the A_n's and then substituting into (2), the θ component of the electric field in the radiation region is obtained

$$E_\theta(r,\theta) = K\sum_{n=1}^{\infty}P_n(\cos\theta_0)P_n^1(\cos\theta)\frac{2n+1}{n^2+n}\,\frac{h_{n-1}^{(2)}(kr)-(n/kr)h_n^{(2)}(kr)}{h_{n-1}^{(2)}(ka)-(n/ka)h_n^{(2)}(ka)} \qquad (6)$$

where

$$K = \sqrt{\frac{\mu}{\epsilon}}\left(\frac{\alpha e^{-ika}-\beta e^{ika}}{a}\right).$$

At distances from the antenna large compared to a wavelength, i.e., $kr\to\infty$, the expression $[h_{n-1}^{(2)}(kr)-(n/kr)h_n^{(2)}(kr)]$ approaches $(i)^n e^{-ikr}/kr$. Hence for $kr\to\infty$

$$E_\theta(r,\theta) = K\frac{e^{-ikr}}{r}\sum_{n=1}^{\infty}P_n(\cos\theta_0)P_n^1(\cos\theta)\frac{2n+1}{n^2+n}\,\frac{i^n}{h_{n-1}^{(2)}(ka)-(n/ka)h_n^{(2)}(ka)}. \qquad (7)$$

The expression (7) yields the θ component of the electric field along the circle r, θ when $kr\gg1$. Dividing (7) by $(E_\theta(r,\pi/2)$ yields the normalized radiation distribution function

$$R(\theta) = \frac{E_\theta(r,\theta)}{E_\theta\left(r,\frac{\pi}{2}\right)} = \frac{\sum_{n=1}^{\infty}P_n(\cos\theta_0)P_n^1(\cos\theta)\frac{2n+1}{n^2+n}i^n[h_{n-1}^{(2)}(ka)-(n/ka)h_n^{(2)}(ka)]^{-1}}{\sum_{n=1}^{\infty}P_n(\cos\theta_0)P_n^1(0)\frac{2n+1}{n^2+n}i^n[h_{n-1}^{(2)}(ka)-(n/ka)h_n^{(2)}(ka)]^{-1}}.$$

$R(\theta)$ is a complex function of θ whose modulus $M(\theta)$ gives the distribution of the absolute value of the normalized electric field over a hemisphere of very large radius. Explicitly,

$$M(\theta) = \frac{\left|\sum_{n=1}^{\infty}P_n(\cos\theta_0)P_n^1(\cos\theta)\frac{2n+1}{n^2+n}i^n[h_{n-1}^{(2)}(ka)-(n/ka)h_n^{(2)}(ka)]^{-1}\right|}{\left|\sum_{n=1}^{\infty}P_n(\cos\theta_0)P_n^1(0)\frac{2n+1}{n^2+n}i^n[h_{n-1}^{(2)}(ka)-(n/ka)h_n^{(2)}(ka)]^{-1}\right|}. \qquad (8)$$

III. Radiation Patterns

For given values of the flare angle θ_0, the antenna length a, and the operating angular frequency $\omega(=kc$ where c is the velocity of light in vacuum), the radiation pattern of the antenna can be computed from (8). When $ka\ll1$, only the first terms in the series need be considered and consequently

$$M(\theta) = \sin\theta, \qquad (9)$$

which is to be expected for the radiation pattern of an infinitesimal vertical dipole on a perfectly conducting horizontal plane. As the parameter ka increases, the departure from (9) becomes more pronounced as is seen from the polar plots of Fig. 2 for the case $\theta_0=\pi/6$.

Acknowledgment

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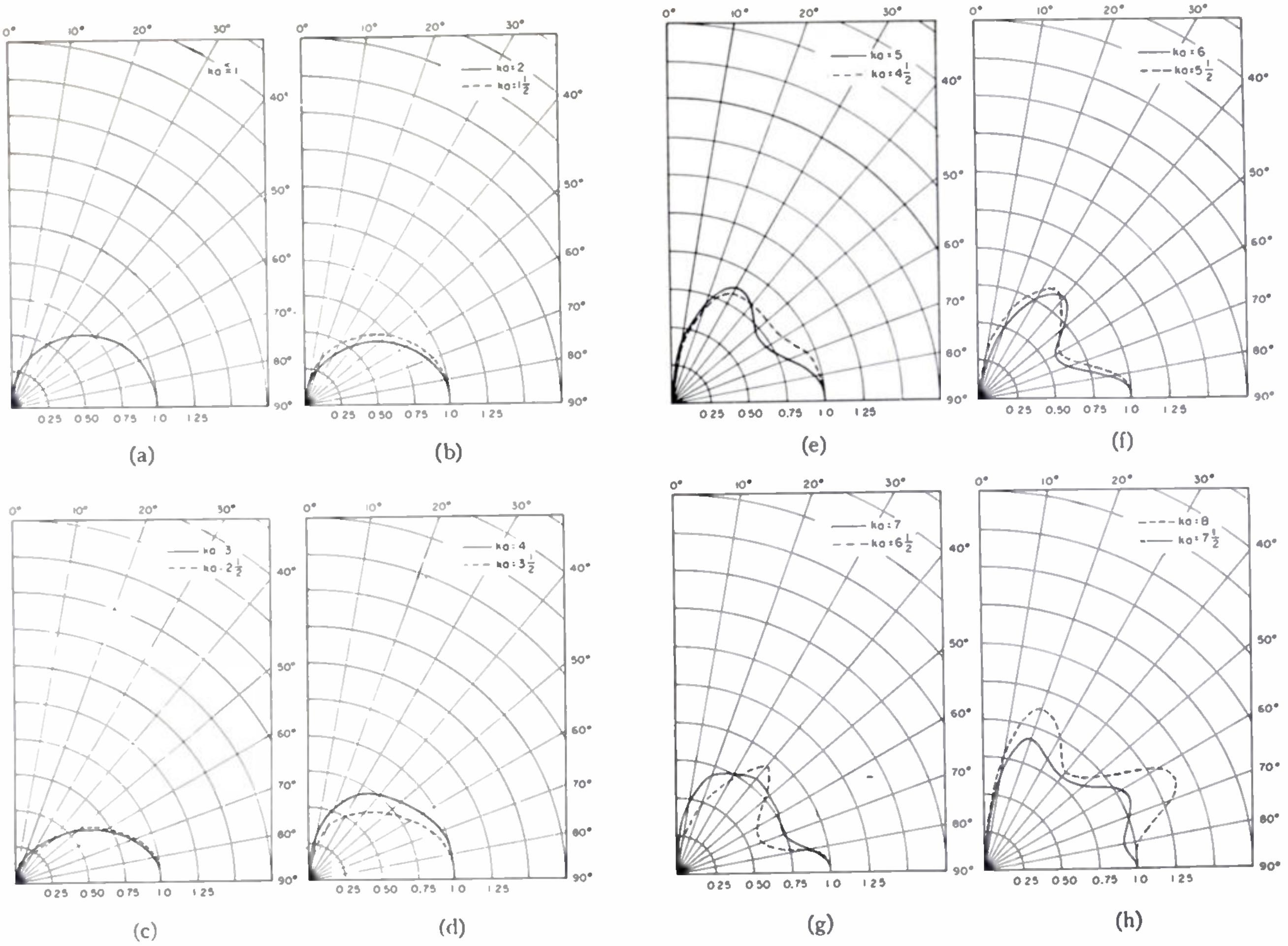

Fig. 2—Plots of the absolute values of the far-zone electric field as a function of the zenithal angle θ for various values of ka and flare angle $\theta_0 \cdot ka = 2\pi x$ length of antenna in free-space wavelengths.

Angular Jitter in Conventional Conical-Scanning, Automatic-Tracking Radar Systems*

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Summary—**Four sources of angular jitter in a conventional conical-scanning, automatic-tracking radar are discussed and each is expressed in terms of range and the several system parameters. The total-angular-jitter-versus-range curve is shown to have a characteristic shape exhibiting a certain range interval of optimum tracking. The importance of beamwidth in determining the magnitude of angular jitter is also stressed.**

Introduction

IN THE CONVENTIONAL fire control system (Fig. 1), present target position data obtained from the fire control director are used in the computer, to calculate target rates and accelerations. These target rates and accelerations are in turn used to predict the future target position, which is required in the solution of the fire control problem.

In present-day practice an automatic-tracking radar is used to control the director, and this tracking method yields data which include random deviations from the true target position. These perturbations are called "jitter." Jitter is a matter of consequence because the computer, in computing rates and accelerations from the radar data, greatly amplifies these random deviations with the result that dispersions in the gun (or missile) orders are many times the values of the original perturbations.

With the development of modern high-altitude, high-speed, highly maneuverable targets, the computer's task has been made more difficult on two accounts: (a) The computer smoothing time has been *shortened* because the targets are faster and more maneuverable, and (b) the time required for projectile flight has been *lengthened* because longer range missiles are used. Since the amplification of the tracking perturbations is pro-

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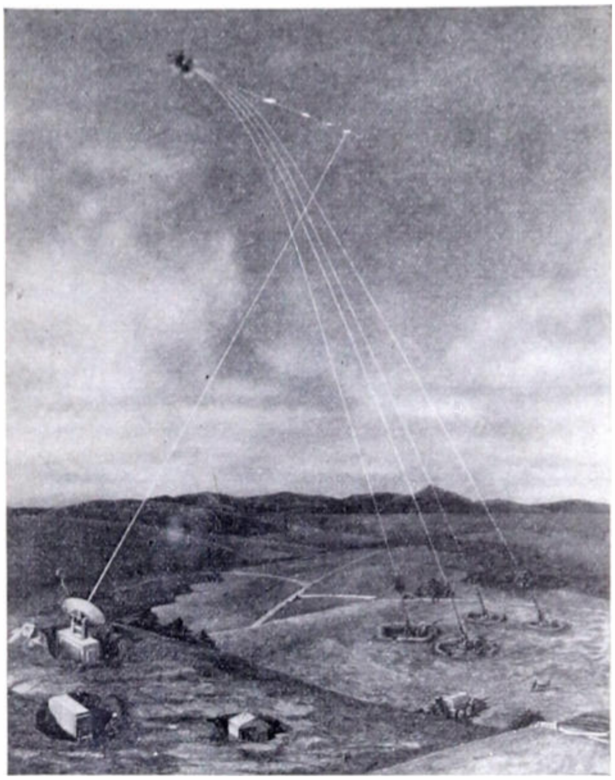

Fig. 1—Artist's conception of typical radar-computer-director-gun-moving target system in operation.

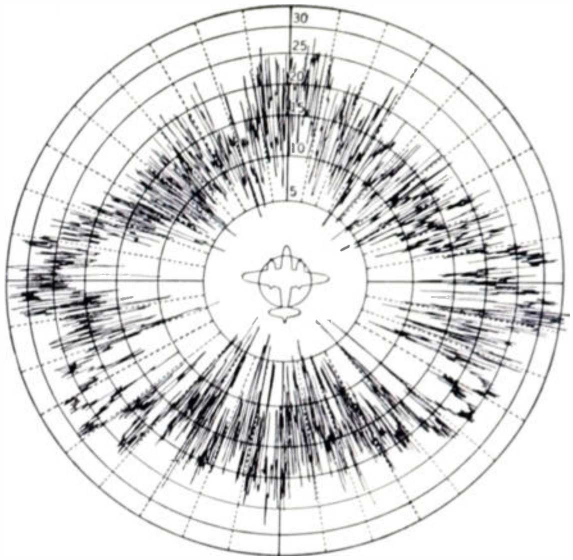

Fig. 2—Return power from a B-26 aircraft at 10-cm wavelength as a function of azimuth angle. The radar data were taken while the aircraft was rotated on a turntable under conditions excluding other return of radar energy. Return power is plotted in decibels. (From "Radar System Engineering," edited by L. N. Ridenour, McGraw-Hill Book Company, Inc., New York, N. Y., 1947.)

portional to the projectile flight time and inversely proportional to the computer smoothing time, it is necessary more than ever before that tracking jitter be reduced to a minimum.

The purpose of the present paper is to list the several sources of tracking jitter present in a conventional conical-scanning, automatic-tracking radar system and to show how they are dependent upon range and the system parameters. It is assumed that automatic tracking in range is accomplished with negligible jitter, and therefore the discussion is confined to the subject of angular jitter. Finally, because of the lack of quantitative information on the phenomena involved, much of the discussion will be qualitative in nature.

Sources of Angular Jitter

The sources of angular jitter in a typical conical-scanning, automatic-tracking radar system may be listed as follows: (a) random fluctuation in the received signal strength, (b) random fluctuation in the angle of arrival of the received signal, (c) receiver noise, and (d) residual servo jitter.

Random Fluctuation in Received Signal Strength

Fluctuation in the received signal strength is due mainly to the time-varying character of the target scattering cross section as viewed from the radar and is caused by the variations in target aspect as a whole and relative motions of the various parts of the target (Fig. 2). The presence of this phenomenon is substantiated by the echo-power-versus-azimuth-angle plots for typical aircraft reported in the literature.[1] Transmitter modulation and misfire and random changes in atmospheric insertion loss also are factors contributing to this fluctuation, but their effect is probably negligible and is outside the scope of the present paper.

The angular jitter θ_A resulting from this fluctuation in received signal strength is a function of the target's geometry, speed, and course. For a given target, velocity and course, the rms angular jitter θ_A is inversely proportional to the slope m' of the per cent-modulation-versus-error-angle curve. (This assumes that m' is constant for the small error angles of interest.) For small error angles, m' is proportional to the slope S of the normalized antenna *voltage* pattern at the spin axis (when the angular dimension used is beamwidth) and inversely proportional to the beamwidth. Therefore, θ_A is proportional to the beamwidth and inversely proportional to S (Fig. 3).

It may also be expected that θ_A will increase as the servo gain-bandwidth product increases. The servo gain-bandwidth product ξ is defined as

$$\xi = \int_0^\infty \frac{\theta_o}{\theta_i}\, d\omega$$

where

θ_0 = output angle as a function of frequency
θ_i = input angle as a function of frequency
ω = frequency in radians per second.

Just how θ_A varies with the servo gain-bandwidth product depends upon the energy-versus-frequency distribution of the received signal-strength fluctuation. How-

[1] L. N. Ridenour, "Radar System Engineering," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 76–79; 1947.

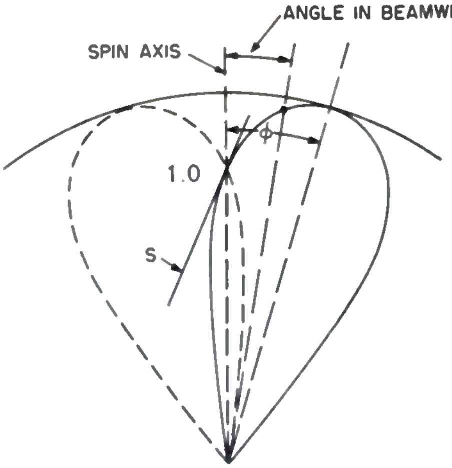

Fig. 3—Antenna voltage pattern. Slope S is tangent to the lobe at the spin axis.

ever, if it is assumed that this energy is fairly uniformly distributed over the servo bandwidth and a continuous function of frequency, θ_A is proportional to the square root of the servo gain-bandwidth product.

On the other hand, as long as the automatic gain control is operating in such a manner as to maintain a constant servo-loop gain, θ_A should be independent of range and other system parameters such as transmitter power, receiver noise figure, antenna efficiency, and so forth; in short, those parameters which define the echo signal-to-noise ratio.

Thus,

$$\theta_A = K_A \frac{\xi^{1/2} W_B}{S} \quad \text{mils} \quad (1)$$

where

K_A = proportionality constant
ξ = servo gain-bandwidth product
S = slope of normalized antenna *voltage* pattern at the spin axis
W_B = system beamwidth in mils.

Random Fluctuation in the Angle of Arrival of the Received Signal

Random fluctuation in the angle of arrival of the received signal is mainly due to the fact that the *effective center of radiation* of the radar echo from a target as geometrically complex as an aircraft wanders in a random fashion over the limits of the target as its aspect relative to the radar changes. (Random aberration in the atmosphere might also be a contributing factor.) This motion of the effective center of radiation is termed *glint* (Fig. 4) and is a target characteristic closely related to the observed fluctuation in received signal strength due to change in target attitude.

For a given target it is reasonable to assume that the effective center of the target moves a given number of yards for a given change in target aspect. Therefore, the angular jitter due to this phenomenon θ_B is inversely proportional to range. The rms value of θ_B also is proportional to the square root of the servo gain-bandwidth product if it is assumed that the energy-versus-frequency distribution in the angle-of-arrival fluctuation is fairly uniformly distributed over the servo bandwidth and a continuous function of frequency. θ_B, however, is independent of the echo signal-to-noise ratio and m' as long as the automatic gain control maintains the servo loop gain constant.

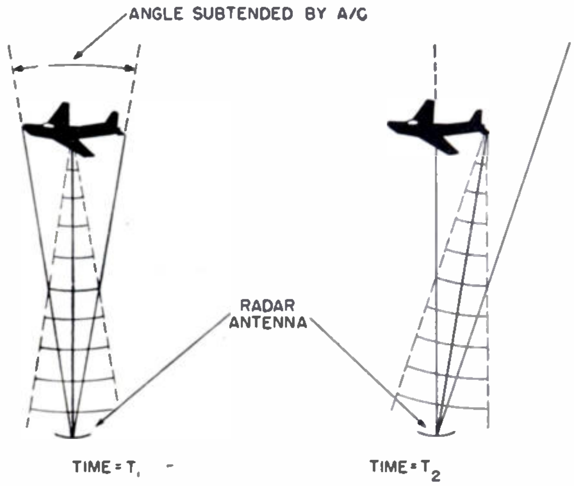

Fig. 4—One source of angular jitter is glint, or the random shifting of the target's effective center of radiation. In the left view the effective center is amidships. The right view shows it as the tail structure. The aircraft is shown near the radar antenna; thus, glint effect is heightened as an illustration aid.

Thus,

$$\theta_B = \frac{K_B \xi^{1/2}}{R} \quad \text{mils} \quad (2)$$

where

K_B = proportionality constant
ξ = servo gain-bandwidth product
R = range.

Receiver Noise

Every receiver has associated with it an equivalent thermal noise source that generates an amount of noise power which is essentially constant for that particular receiver. The angular signal is in competition with this noise for recognition in the angular sensing device and as this signal-to-noise ratio decreases one would expect the angular tracking jitter to increase. Now, since the angular sensing device in a conical-scanning radar operates on a voltage basis, and if it is assumed that the angular error voltage varies linearly with error angle, the angular jitter due to receiver noise θ_C is inversely proportional to the square root of the echo-signal-to-receiver-noise-power ratio.

That this is so can be shown with the help of Fig. 5. In Fig. 5 the angular-error-voltage-versus-error-angle curve is shown as V-shaped with the base submerged in receiver noise. The radar line-of-sight wanders randomly over an angle which is proportional to the width of the portion submerged in the noise. Now the action of the automatic gain control is such as to maintain the "V" ordinate fixed and to make the noise voltage proportional to the square root of the receiver-noise-to-echo-signal-power ratio and hence submerge the "V" to a depth, and for straight sides to a width, which is inversely proportional to the square root of the echo-signal-to-receiver-noise-power ratio.

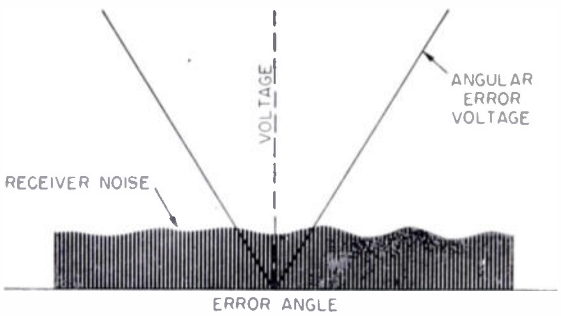

Fig. 5—Plot of error voltage versus associated angular error in the presence of receiver noise.

θ_C is also directly proportional to the system beamwidth and the square root of the servo gain-bandwidth product and inversely proportional to the slope of the normalized antenna *voltage* pattern at the spin axis.

Thus,

$$\theta_C = C_1 \frac{P_n^{1/2}\xi^{1/2}}{P_r^{1/2}S} W_B \quad \text{mils} \quad (3)$$

where

C_1 = proportionality constant
P_n = receiver noise power
P_r = echo signal power.

From radar considerations, the echo signal power varies with the several system parameters and range as follows:

$$P_r = \frac{C_2 P_t G_t^2 \sigma_t \lambda^2}{R^4} \quad (4)$$

where

C_2 = proportionality constant
P_t = transmitter power
G_t = antenna power gain along the spin axis
σ_t = target scattering cross section
R = range
λ = wavelength.

For automatic-tracking radars, where pencil beams are employed,

$$G_t = \frac{C_3 \eta}{W_B^2}\beta \quad (5)$$

where

C_3 = proportionality constant
η = antenna efficiency
β = a factor which allows for the loss in *power* gain along the spin axis as a result of the beam tilt ϕ (Fig. 3).

Thus, from (4) and (5)

$$P_r = \frac{C_4 P_t \eta^2 \sigma_t \beta^2 \lambda^2}{W_B^4 R^4} \quad (6)$$

where $C_4 = C_2 C_3$.

The receiver noise power may be expressed as follows if it is assumed to be of uniform density over the band

$$P_n = C_5 N \quad (7)$$

where

C_5 = proportionality constant
N = receiver noise figure.

Thus, from equations (6) and (7) equation (3) becomes

$$\theta_C = \frac{K_C N^{1/2} \xi^{1/2} W_B^3 R^2}{P_t^{1/2} \eta \sigma_t^{1/2} S \beta \lambda} \quad \text{mils} \quad (8)$$

where K_C = proportionality constant.

Residual Servo Jitter

Even when the line-of-sight is slaved to a reference, which in itself is noise-free, there is jitter about the reference. This jitter θ_D is due to the presence of electrical or mechanical backlash, sticky friction, and noise in the servo system. To a first approximation, θ_D in mils is assumed to be proportional to the square root of the servo-gain-bandwidth product and independent of range and other system parameters.

Thus,

$$\theta_D = K_D \xi^{1/2} \quad \text{mils} \quad (9)$$

where K_D = proportionality constant.

Conclusions

For a given system-target combination, (1), (2), (8), and (9) above reduce to

$$\theta_A = K_1 \quad \text{mils} \quad (10)$$

$$\theta_B = \frac{K_2}{R} \quad \text{mils} \quad (11)$$

$$\theta_C = K_3 R^2 \quad \text{mils} \quad (12)$$

$$\theta_D = K_4 \quad \text{mils.} \quad (13)$$

Admittedly, the state of the art has not advanced to a point that permits the evaluation of the K's in (10) through (13) for a particular set of system parameters and target. With this in mind, the plot of equations (10) through (13) on Fig. 6 must be interpreted as being qualitative rather than quantitative in nature. The total angular jitter θ_t is computed by assuming that the individual angular jitter sources are random; that is,

$$\theta_t = \sqrt{\theta_A^2 + \theta_B^2 + \theta_C^2 + \theta_D^2} \quad \text{mils.} \quad (14)$$

This curve is given on Fig. 6 and should also be interpreted as a qualitative plot.

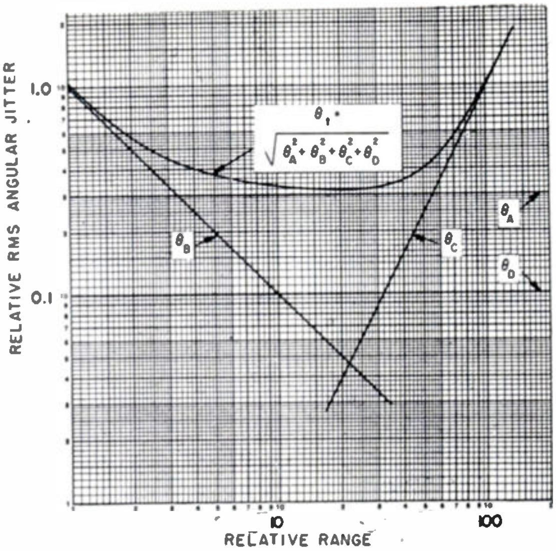

Fig. 6—Plot of sources of angular jitter.

From the curve of θ_t versus R of Fig. 6, it is seen that best tracking, i.e., minimum jitter, is obtained at an intermediate range where the contributions to angular jitter from glint and receiver noise are small compared with the contributions from residual servo jitter and random amplitude fluctuation in the echo signal. Such a range bracket, in general, exists for all systems. However, if there is no range interval where both θ_B and θ_C are negligible, the θ_t-versus-R curve will have no flat portion.

TABLE I

Summation of Effects of Jitter Sources on System Parameters

Jitter Source	System Parameter; Parameter Exponent							
	W_B	ξ	N	P_t	η	S	β	λ
θ_A	1	0.5	0	0	0	−1	0	0
θ_B	0	0.5	0	0	0	0	0	0
θ_C	3	0.5	0.5	−0.5	−1	−1	−1	−1
θ_D	0	0.5	0	0	0	0	0	0

In order to obtain the smallest amount of tracking jitter over an extended range, it is obvious that all four curves θ_A, θ_B, θ_C, and θ_D should be depressed as much as possible. To show better how the several system parameters determine the positions of these curves, Table I has been constructed from equations (1), (2), (8), and (9). Table I shows at what exponent each system parameter enters in determining the angular jitter from each source.

Reference to Table I shows that the servo gain-bandwidth product ξ enters each angular jitter component as a square root. It is therefore concluded that ξ should be made as small as possible consistent with the permissible servo lag and the target dynamics. Having minimized ξ, θ_B and θ_D are independent of the other system parameters although the type and size of the target affects θ_B and the care in the design of the servo system affects θ_D.

θ_A is seen to vary directly with the beamwidth W_B whereas θ_C is proportional to the cube of the beamwidth. It is reasonable, then, to employ the narrowest beam possible. Of course, care should be taken in design so that the receiver noise figure N is minimized and the antenna efficiency η is maximized since both these parameters are effective in determining θ_C. Increasing the transmitter power P_t and wavelength λ, also decrease θ_C, for θ_C is inversely proportional to the square root of P_t and inversely proportional to λ. Of course, it is implied that the aperture dimension is adjusted so as to maintain constant beamwidth if the wavelength is increased. It should further be noted that in applications where the maximum available power is required, P_t is nominally proportional to λ^2.

Both θ_A and θ_C are seen to vary inversely with S so that from this consideration ϕ (Fig. 3) should be chosen to maximize S. However, θ_C also varies inversely with β which decreases continuously as ϕ is increased. It is, therefore, necessary to choose an intermediate value for ϕ which will best satisfy the particular problem.

Of the several parameters, the beamwidth is the most potent factor. Low receiver-noise figure and high antenna efficiency can be obtained from careful design. The target dynamics establish the servo gain-bandwidth product in any given application. Increasing the transmitter power decreases the angular jitter due to receiver noise, but this is an expensive way, and other channels, especially beamwidth, should first be fully exploited.

Acknowledgment

The author makes appreciative acknowledgment to colleagues of the armament engineering departments of Sperry Gyroscope Company for many valuable suggestions made specifically for this paper or given earlier in the course of routine work.

The Statistical Properties of Noise Applied to Radar Range Performance*

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Summary—The advent of high-speed scanning radars which yield but a few pulses per scan element and displays which collapse the range co-ordinate has necessitated an analytical approach to the problem of radar range performance, as empirical methods have proven inadequate. An analysis is made which yields several graphs from which the radar detection range can be calculated for any desired detection probability, and from which computations of the probability of false target echoes can be made. Methods are also indicated which permit dark-tube operation and video pulse stretching. Statistical concepts involved in the effect of receiver noise are introduced.

I. Introduction

THE PERFORMANCE of radar systems possessing a display in which the range co-ordinate is not collapsed (PPI and B type) has been analyzed in several past papers.[1-3] In general, the calculations of performance are based on empirical results and apply quite well to slow scanning radars where a large number of pulses per target are available. The advent of high-speed scanning radars, yielding a small number of pulses per target and using displays in which the range co-ordinate is collapsed, has necessitated a somewhat different approach in calculating performance. It is the primary purpose of this paper to demonstrate a method of calculating the range performance of a radar on a statistical basis, that is, for a given range, the probability of detection of a target and the probability of occurrence of false target echoes may be stated. It is no longer necessary to use such quantities as maximum effective range or maximum useful range which are subject to wide variations in calculation and interpretation, or to ignore false target echoes and their consequences.

Displays which collapse the range co-ordinate (C and G types) have useful application in radar. However, since the collapse of the range co-ordinate is, in effect, a compression of the time co-ordinate, the spot on the cathode-ray tube remains essentially stationary for one or more repetition intervals. This results in noise pile-up and a consequent reduction in range; thus these displays have been looked upon with some disfavor. Actually, they possess a marked technical advantage in that short pulses (less than 2 microseconds) may be stretched to enhance the presentation. Such stretching in PPI and B displays simply enlarge the target without intensifying it. While some range resolution may be lost by pulse stretching, it is usually of little consequence. This paper presents a method of obviating this range reduction due to noise pile-up, and of obtaining a video output which allows a dark-tube display (screen illuminated by targets).

In the work that follows, clutter and jamming have been neglected. Anti-jam techniques have been widely discussed and the clutter problem is greatly dependent on the specific radar under consideration. In any event, if clutter and jamming appear as signals at the intensity grid of the cathode-ray tube, there is little left to do but present them.

II. Noise Probability Distributions and False Alarm Time

In order to obtain satisfactory radar operation, it is necessary to control the occurrence of noise pulses or false target indications. Consider the probability distribution of pure receiver noise which has passed through the detector and video amplifier. The probability distribution of this noise voltage can be assumed to be the distribution of the envelope of noise (not signal plus noise) in the intermediate-frequency section of the receiver. It has been shown[4,5] that this distribution is of the form

$$p(V)dV = V \exp(-V^2/2)dV, \tag{1}$$

where V is ratio of the envelope voltage to rms noise (prior to detection), and $p(V)dV$ is the probability of an instantaneous noise voltage lying between V and $V+dV$. A plot of equation (1) is shown in Fig. 1. The probability of exceeding a particular voltage V_c is

$$p_N = \int_{V_c}^{\infty} V \exp(-V^2/2)dV. \tag{2}$$

The integration is shown graphically in Fig. 1. Upon choosing a particular threshold level V_c, the fraction of time during which noise will exceed this level may be determined. Thus, the fraction of time may be controlled by the choice of V_c. The value of P_N required to produce on the average only one pulse in a specified time interval is

$$p_N = \frac{\text{time duration of noise pulse}}{\text{time interval during which noise can appear on the indicator}}. \tag{3}$$

* Decimal classification: R537×R161.6. Original manuscript received by the Institute, October 19, 1949.

† General Engineering and Consulting Laboratory, General Electric Company, Schenectady, N. Y.

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

[2] Kenneth A. Norton and Arthur C. Omberg, "The maximum range of a radar set," Proc. I.R.E., vol. 35, pp. 2–24; January, 1947.

[3] Ruby Payne-Scott, "The visibility of small echoes on radar PPI displays," Proc. I.R.E., vol. 36, pp. 180–196; February, 1948.

[4] S. O. Rice, "Mathematical analysis of random noise," *Bell. Sys. Tech. Jour.*, vol. 23, pp. 282–332, July, 1944; and vol. 24, pp. 46–156; January, 1945.

[5] S. Goldman, "Frequency Analysis Modulation and Noise," McGraw-Hill Book Co., Inc., New York, N.Y.; 1948.

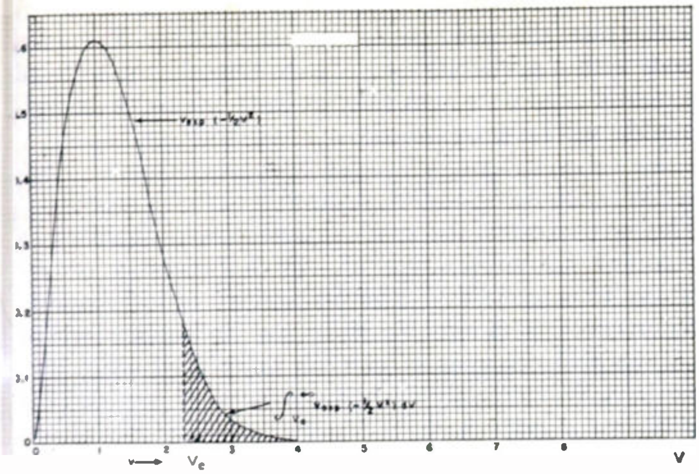

Fig. 1—Probability distribution of noise after detection.

The numerator of (3) is approximately equal to the reciprocal of the receiver bandwidth ($1/\Delta f$), while the denominator is equal to a quantity which will be called the false alarm time (τ_f) multiplied by the fraction of time f_n between successive pulses that the receiver is gated on. The false alarm time may be defined as the average time interval between false target indications. Equation (3) then becomes

$$p_N = \frac{1}{\Delta f \tau_f f_n}. \tag{4}$$

Equating (2) and (4) we obtain

$$\int_{V_c}^{\infty} V \exp(-V^2/2)dV = \exp(-V_c^2/2) = \frac{1}{\Delta f \tau_f f_n}. \tag{5}$$

A plot for (5) is shown in Fig. 2.

For any particular radar design Δf and f_n are determined, and consequently for any desired false alarm time (τ_f), an amplitude selection level V_c can be set. The manner in which the amplitude selection level affects the minimum detectable signal and consequently radar range, is taken up in the following section.

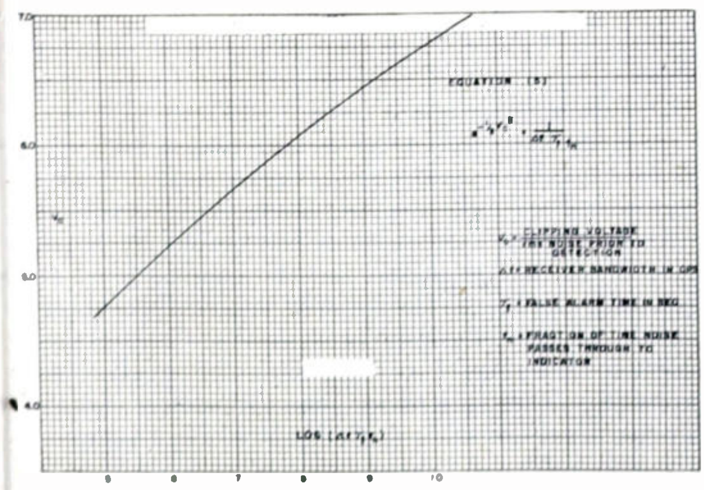

Fig. 2—Threshold value versus false alarm time and receiver characteristics.

III. Minimum Detectable Signal and Radar Range

The minimum detectable signal S_M is determined by the relation

$$S_M = S_N f_V = \overline{NF} kT\Delta f\, f_v, \tag{6}$$

where

NF = receiver noise figure, dimensionless ratio
k = Boltzmann's constant, Joules per degree Kelvin
T = Temperature, degrees Kelvin
Δf = receiver bandpass, cycles per second.

The visibility factor f_v is a factor by which the receiver noise power S_N must be multiplied in order that a signal be adequately identified on the indicator. If a linear detector is used

$$f_v = \left[\frac{\text{Minimum detectable rms signal voltage}}{\text{rms noise voltage}}\right]^2 = (V_s/\sqrt{2})^2 = 1/2V_s^2 \quad \text{where} \tag{7}$$

V_s = peak signal to rms noise voltage ratio required for adequate signal identification. It is convenient to define two terms, f_v and V_s, in that f_v applies on a power basis and V_s on a voltage basis.

The value of V_s depends primarily on three factors: (1) the value of V_c; (2) the probability distribution of signal plus noise; and (3) the required probability of signal plus noise, exceeding the threshold value (V_c).

The probability of detecting a single pulsed signal p_s is always less than certainty because the chance of noise subtracting from the signal amplitude exists. The minimum detectable signal, therefore, is a statistical quantity and is governed by the laws of probability. In Fig. 3 is shown the distribution of a sinusoidal signal

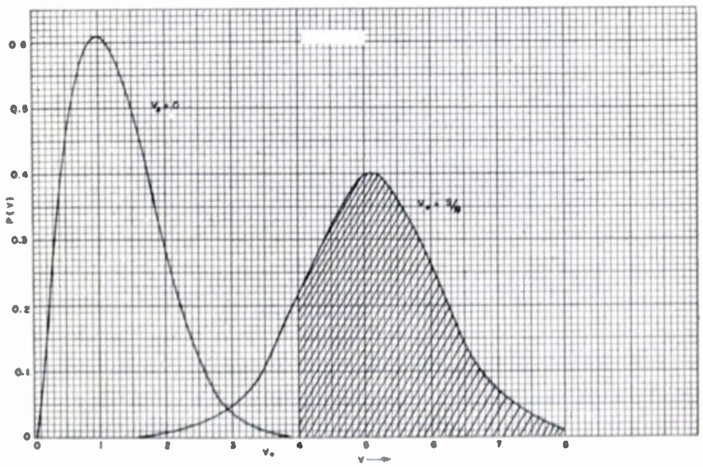

Fig. 3—Probability of distribution of noise and signal plus noise after detection.

plus noise and pure noise. Once V_c has been determined by choosing a false alarm time, a signal-to-noise ratio V_s must be found such that the shaded area equals the fraction of time signal plus noise is to exceed the threshold value. The distribution of signal plus noise is known to be[6]

$$p(V)dV = V \exp\left[-1/2(V_s^2 + V^2)\right] I_0(V_s \cdot V)dV, \quad (8)$$

where $I_0(V_s \cdot V)$ is the modified Bessel Function of order zero and argument $(V_s \cdot V)$.

Therefore V_s is determined by

$$\int_{V_c}^{\infty} V \exp\left[-1/2(V_s^2 + V^2)\right] I_0(V_s V)dV = p_s. \quad (9)$$

This integral cannot be evaluated in closed form. Rice[4] has shown a solution based on a series found by Bennett. Neglecting terms of magnitude less than V_s^{-3} the value of the integral becomes

$$p_s = 1/2 - 1/2 erf\left(\frac{V_c - V_s}{2}\right) + \frac{1}{2V_s\sqrt{2\pi}} \left[1 - \frac{V_c - V_s}{4V_s} - \frac{1 + (V_c - V_s)^2}{8V_s^2}\right] \exp\left[\frac{(V_c - V_s)^2}{2}\right]. \quad (10)$$

For a given p_s and V_c, solutions of this equation for V_s can only be found by trial and error, and consequently the most satisfactory method of employing the equation in a numerical problem is to plot it as shown in Fig. 4.

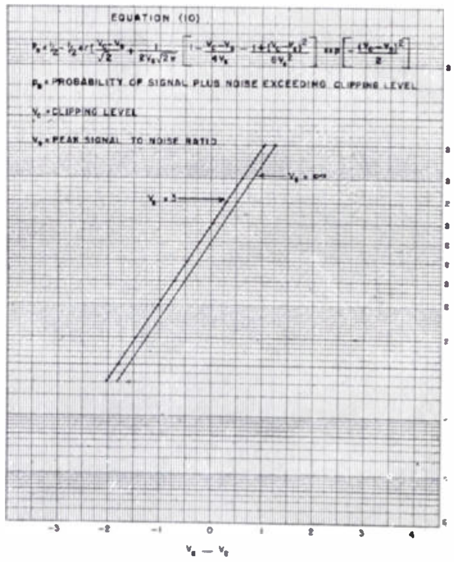

Fig. 4—Signal-to-noise ratio versus probability of exceeding threshold value.

The value of V_s, f_v and consequently the minimum detectable signal is a function of the false alarm time and the probability of signal plus noise exceeding V_c. Expressed in functional notation, we have

$$S_M = F(\tau_f, p_s). \quad (11)$$

It will now be shown how p depends on the number of pulses per scan element and the over-all probability of signal detection (p_d). Equation (11) may then be written as

$$S_M = G(\tau_f, N_e, p_d). \quad (12)$$

In many radar applications, scanning requirements are such that the number of pulses per element N_e varies over the explored volume. If there are N_e pulses, each with a probability p_S of exceeding V_c the over-all probability of detection (i.e., at least one pulse exceeding V_c) will be determined by

$$p_d = 1 - (1 - p_s)N_e. \quad (13)$$

This is true since the over-all detection probability is obtained by subtracting all the unfavorable occurrences from certainty. The probability of unfavorable occurrences is $(1-p_s)N_e$, since the probability of an unfavorable occurrence for each pulse is $(1-p_s)$, and we have N_e mutually exclusive events. A plot of (13) is shown in Fig. 5.

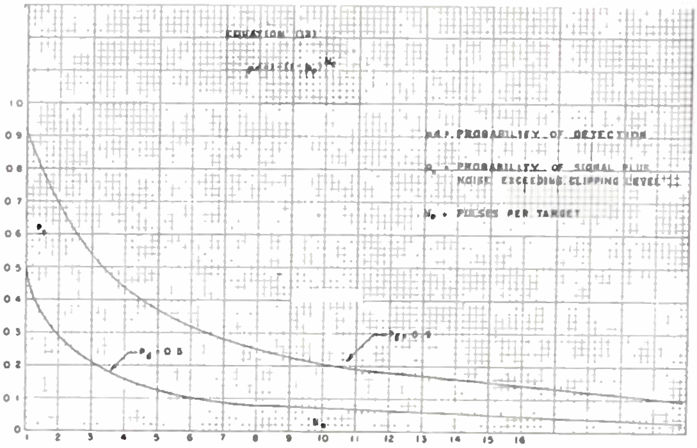

Fig. 5—Probability of signal plus noise exceeding threshold value versus pulses per target.

Suppose now that a fixed probability of detection p_d is required. By (13), p_s may be found for any value of N_e. Consequently, (12) follows from (11).

Once the minimum detectable signal for the required detection probability and false alarm time has been ascertained, the corresponding range for specific radar parameters can be found from the radar range equation

$$R = \left[\frac{P_t \sigma A_a^2}{4\pi S_M \lambda^2}\right]^{1/4} = \left[\frac{P_t \sigma A_a^2}{4\pi NFkT\Delta f \lambda^2 f_v}\right]^{1/4}, \quad (14)$$
$$= C\left(\frac{1}{f_v}\right)^{1/4}$$

where

P_t = transmitted power watts

σ = effective target cross section, square feet

A_a = effective aperture area, square feet.

It would be well now to summarize one possible method of using the equations and graphs. A desired false alarm time is chosen and the clipping level determined from Fig. 2. The required detection probability and the number of pulses per element are used in conjunction with Fig. 5 to determine the single pulse probability. Once the single pulse probability and the clipping level are known, the required peak signal to rms noise ratio and hence the visibility factor may be obtained from Fig. 4. The constant C of (14) may be calculated from known radar parameters. The actual range is found by substituting this value of visibility factor and predetermined constant C in (14). The numerical example of Section IV follows this procedure.

Alternative uses include the choice of range, probability of detection, and determining false alarm time, or the choice of range, false alarm time, and determining probability of detection.

IV. Numerical Example

$P_t = 40$ kw	$\Delta f = 1$ mc $= 10^6$ cps
$\sigma = 100$ ft^2	$\tau_f = 10$ min. $= 3.6 \times 10^4$ sec
$A = 1.61$ ft^2	$f_n = 0.1$
$f_a = 0.62 \quad A_a = A f_a$	$N_e = 6$
$\lambda = 3.0$ cm $= 0.1$ ft	$p_d = 0.9$
$NF = 12$ db $= 16.0$	$k = 1.38 \times 10^{-23}$ joules/°K.
$T = 300°$ Kelvin	

From equation (14)

$$C = \left[\frac{P_t \sigma A_a^2}{4\pi\gamma^2 NFkT\Delta f}\right]^{1/4} = 5 \times 10^4 \text{ yards.}$$

From Fig. 2,

for $\log_{10}(\Delta f \tau_f f_n) = \log_{10}(3.6 \times 10^9) = 9.556$, $V_c = 6.62$.

From Fig. 5 for $p_d = 0.9$ and $N_e = 6$

$$p_s = 0.32.$$

From Fig. 4 for $p_s = 0.32$ and $V_s = \infty$

$$V_s - V_c = -0.6,$$

then

$$V_s = 6.02.$$

From Fig. 6 for $V_s = 6.02$

$$f_v = 18$$

$$R = C(f_v)^{-1/4} = 5 \times 10^4 (18)^{-1/4}$$

$$= 2.43 \times 10^4 \text{ yards for 90 per cent detection}$$

probability and a 10-minute false alarm time.

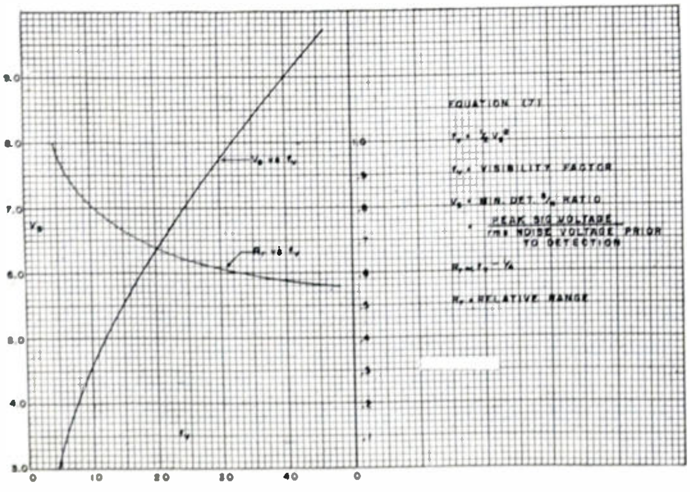

Fig. 6—Signal-to-noise ratio and range versus visibility factor.

V. Discussion

The relations between detection probability, the clipping level, the pulses per element, the false alarm time, and other factors are presented both in equation form and graphically. The numerical example of Section IV indicates one method of using these relationships. Another use, of equal importance, is to indicate the effect of various system parameters on the radar performance. For a selected false alarm time, it would be desirable to have a minimum value of clipping level (Fig. 2). Thus, the receiver bandwidth should be as narrow as other considerations, such as magnetron wander, permit. Small values of receiver bandwidth correspond to long pulses. It is also important that the interval during which the receiver is gated on be maintained short through the use of sensitivity time control (STC), and a range gate positioned near the maximum useful range.

This reasoning indicates a long pulse which, since the magnetron duty factor is a constant, indicates a low repetition rate and hence a low value of pulses per element. It would appear desirable from Fig. 5 to have as large a value of pulses per element as possible. However, a little thought will show that the receiver bandwidth not only affects the clipping level but also has the more important effect of increasing the receiver noise power. On this basis it is apparent that the pulses per element should not be increased at the expense of increasing the receiver bandwidth. Since, in general, physical scanning requirements affect the value of pulses per element, a value should be chosen in the range of three to six, which is the knee of the curve, provided the frame time thereby determined is reasonable. In many scans, the pulses per element vary throughout the scan and the range coverage is not uniform. Using the method of the numerical example, the exact range coverage may be calculated.

Since the visibility factor is proportional to the square of the peak signal to rms noise ratio on a voltage basis, it is desirable to keep this ratio and hence the single pulse detection probability a minimum. It can be seen from Fig. 5 that even for a 90-per cent detection probability, the single pulse detection probability may be as low as 20 per cent for ten pulses per element.

It is evident that the device which establishes the threshold level acts to integrate the additional information when more than one pulse is returned from a single target. This does not constitute integration in the conventional sense, that is, use of time-phase information, but the effect is similar.

Fig. 6 is included to show the fourth power relationship between range and visibility factor and further emphasizes the desirability of a low visibility factor. $R_r = 1$ was chosen for $f_v = 4.5$.

VI. Conclusions

The authors feel that the interpretation of range performance of a radar as a statistical quantity is a new and valuable one. Using the methods presented in this paper, the probability of detection of a given target at a particular range and the probability of occurrence of false echoes may be determined. This latter quantity is especially important in that a multiplicity of false echoes can reduce the effectiveness of any radar. The application of this statistical method will accomplish the following:

(1) *Result in more accurate comparisons between existing radars.* The nebulous term, maximum effective range, and misunderstandings concerning false target echoes may be eliminated.

(2) *Result in the systematic optimization of the radar system parameters for the required system performance.* Since performance can now be specified in a more exact manner, system parameters can be optimized to a higher degree of accuracy.

It will also be noted that if a clipping level be literally established, the noise and signal are separated with an average error of one false target per false alarm time interval. The resulting signals (both true and false) are then free to be processed as the designer wishes. One example is pulse stretching which greatly enhances range collapsed displays.

Thus for high-speed scanning radars with few pulses of short duration per target, a dark-tube presentation with the range co-ordinate collapsed and a high probability of target detection may be obtained with little or no sacrifice in range performance.

Principles of the Electrical Rating of High-Vacuum Power Tubes*

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Summary—**A rational system of ratings for high-vacuum power tubes has been arrived at, which permits the determination of ratings by calculation, once the radio-frequency power amplifier and oscillator ratings, class C, have been determined by operating test and life test. A summarized tabulation of the rating factors is given. A system for reducing ratings at high frequencies is also developed.**

Introduction

A RATING is defined by the American Standards Association as follows:[1] "A rating of a machine, apparatus, or device is a designated limit of operating characteristics based on definite conditions. *Note:* Such operating characteristics as load, voltage, frequency, etc., may be given in the rating."

Why are vacuum tubes (or electrical machines and the like) rated? Ratings show the user of a tube the conditions under which he can get satisfactory service and life. They also warn him that operation outside of ratings may result in premature failure or rejection of claims of unsatisfactory service made against the manufacturer.

Before ratings of vacuum tubes are considered, it is necessary to know how tubes may fail. This knowledge will give an insight into the ratings now used by industry.

Any component of a tube is subject to failure under certain conditions. For example, the glass bulb may crack or "suck in" because of excessive temperature. The excessive temperature may be due to high ambient temperature, high dielectric losses in the glass, losses in partially conducting coatings on the inner bulb wall, or bombardment of the bulb by electrons or ions.

Anodes may fail because of excessive temperature, also. Such failures may be indicated by melting, warping, or alloying. The temperature may be localized so that even a reduced dissipation may produce a dangerous condition.

Grids may fail because of excessive temperature evidenced by melting, warping, alloying, or grid emission. The high temperature may be due to excessive electron current to the grid, radio-frequency current flowing in the grid, radiant energy falling on the grid, or back bombardment caused by a high transit angle or emission from the anode. Grid emission may be due to evaporated materials deposited on the grid.

Cathodes or filaments may fail because of high operating voltages, high starting current, or loss of emission. The latter may be caused by high or low operating temperature, or gas bombardment with the gas coming from other electrodes or components which are at excessive temperature. Loss of emission may also be caused by impurities from other components which are evaporated onto the cathode, or excessive space current may cause excessive voltage drop in the cathode coating with attendant sputtering, sparking, and gas release.

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[1] Definition 05.50.040, American Standard Definitions of Electrical Terms (ASA C42-1941), published by American Institute of Electrical Engineers, New York, N. Y.

High voltage may cause failure due to external flash-over, or internal flasharcs, or it may cause excessive rf losses in bulbs or insulators.

The foregoing is only a partial list of how tubes may fail, but it serves to indicate the complexity of the problem of rating tubes.

If we attempt to rate tubes only in terms of the most fundamental parameters, such as maximum anode, grid, bulb, and insulator temperature, or peak voltages and peak currents, we immediately run into the difficulty of instructing the user how to measure such temperatures, voltages, and currents. These parameters are very difficult to measure, even with the best physical apparatus available. Because of this difficulty, the industry has developed indirect ratings, such as maximum dc voltages, dc currents, and dissipation, which have have been applied to various classes of service.

The simplest class of service for a power tube is as a radio-frequency power oscillator or amplifier—class C. The most widely used method for determining these ratings is first to life-test the tube as an oscillator operating at the maximum proposed values for frequency, voltages, currents, and dissipation; and then to operate a sufficiently large number of tubes under these same conditions until there is little doubt that the desired average life has been obtained. The ratings obtained in this manner follow in Table I.

TABLE I

RADIO-FREQUENCY POWER AMPLIFIER AND OSCILLATOR CLASS C TELEGRAPHY

(Key-Down Conditions Without Amplitude Modulation)

dc Plate Voltage (E_{bt})	dc Grid-No. 1 Current (I_{ct})
dc Grid-No. 3 Voltage	Plate Input (P_{it})
dc Grid-No. 2 Voltage	Grid-No. 3 Input
dc Grid-No. 1 Voltage	Grid-No. 2 Input
dc Plate Current (I_{bt})	Plate Dissipation
Maximum Frequency for Above Values	

After the ratings for this class of service have been determined, ratings for other classes of service can be calculated without the necessity of repeating tests. It is the purpose of this paper to give the principles which guide the determination of ratings for all other classes of service from these basic ratings.

NOMENCLATURE

d = cathode-to-grid spacing
e = instantaneous plate-supply voltage
E_b = dc plate voltage
E_{bt} = maximum rated dc plate-supply voltage for telegraph service
E_{bp} = maximum rated dc plate-supply voltage for telephone service with 100-per cent sine-wave modulation
E_r = rms value of the total transformer plate-supply voltage
E_{rm} = maximum rated rms total transformer plate-supply voltage
E_{br} = maximum rated dc plate voltage for unfiltered power supplies
E_{bi} = peak inverse plate voltage
E_{bim} = maximum safe peak inverse plate voltage
e_c = peak positive grid voltage
$E_{c1,2,3}$ = dc grid-No. 1, grid-No. 2, or grid-No. 3 voltage, respectively
E_{cp} = maximum rated dc grid voltage for telephone service
E_{cr} = maximum rated dc grid voltage for unfiltered plate supply operation
E_{ct} = maximum rated dc grid voltage for telegraph service
E_p = peak value of the rf component of plate voltage when the supply voltage is E_b
e_p = instantaneous value of the rf component of plate voltage
E_{peff} = rms value of all the rf components of plate voltage when the plate-supply voltage is not pure dc
f = frequency
f_1 = maximum frequency for maximum ratings
f_2 = an arbitrary frequency
i_p = instantaneous plate-supply current
I_b = dc plate current
I_{bt} = maximum rated dc plate-supply current in telegraph service
I_{bp} = maximum rated dc plate-supply current in telephone service with 100-per cent sine-wave modulation
I_{br} = maximum rated dc plate current for unfiltered plate-supply voltages
I_c = dc grid current
I_{ct} = maximum rated dc grid current for telegraph service
I_{cr} = maximum rated dc grid current for unfiltered plate-supply voltages
k = ratio E_p/E_b
K_1 = ratio E_{bt}/E_{bp}
K_2 = ratio I_{bt}/I_{bp}
K_3 = a constant
K_4 = ratio I_{ct}/I_{bt}
m = angular velocity of modulating voltage
η = efficiency factor
η_0 = efficiency factor at a very low frequency
η_1 = efficiency factor at f_1
η_2 = efficiency factor at f_2
p = angular velocity of supply voltage
P_d = plate dissipation
$P_{g2,3}$ = grid-No. 2, or grid-No. 3 input power, respectively
P_i = plate input
P_{it} = maximum rated plate input for telegraph service
ω = angular velocity of the rf voltage.

Plate-Modulated Radio-Frequency Power Amplifier Class-C Telephony

Carrier Conditions per Tube for Use with a Modulation Factor of 1.0

In this type of service, a modulating voltage is impressed on the plate in series with the dc plate voltage. Therefore, there is both ac and dc power input to the plate. The power input is greatest when the modulation factor is 1.0. The wave form of the modulation voltage under the most severe conditions is assumed to be a continuous sinusoid. In order to simplify the ratings, they are given for the carrier condition, that is, in the absence of the sinusoidal modulation.

First, the maximum rated dc plate voltage E_{bp} must be established. The rating may be determined by two different factors. In the case of high-voltage tubes, the determining factor may be external or internal flashover due to high peak voltages. In the case of small tubes, the rating is determined by rf losses in insulators and leads.

In the high-voltage case, it will be necessary to make some determination of the safe peak positive plate potential. E_b, could then be set equal to one-fourth of this value. During full modulation, the plate-supply voltage varies up to 2 times E_{bp} and the rf plate swing also approximately equals this value, resulting in a peak positive voltage on the anode of $4\times E_{bp}$.

In operation, there are rf losses in insulators and leads caused by the plate swing E_p and the capacitive currents which it produces. Both of these losses are proportional to E_p^2 and thus also proportional to e^2. In the case of full modulation,

$$e = E_{bp}(1 + \sin mt)$$

$$E_{peff} = \frac{kE_{bp}}{\sqrt{2}}\sqrt{\frac{1}{2\pi}\int_0^{2\pi}(1 + \sin mt)^2 d(mt)}$$

$$= \frac{k\sqrt{1.5}}{\sqrt{2}} E_{bp}.$$

In the case of telegraphy, the equivalent rms rf plate voltage is $kE_b/\sqrt{2}$. Thus, when full modulation is applied, the supply voltage E_{bp} produces losses equal to those produced by an unmodulated voltage of 1.225 E_{ip}. Then, the rating E_{bp} can be obtained from E_{bt} by dividing by 1.225. In practice, this factor is usually taken as 1.25 because it is slightly more conservative.

The dc grid-No. 3 voltage and dc grid-No. 1 voltage are generally not modulated, and their ratings for telegraphy and telephony can be the same. The dc grid-No. 2 voltage is generally unchanged too, as there is no rf voltage present at this grid, and even though the grid-No. 2 voltage is modulated, its value is generally so low that no difficulty is experienced even though the voltage is doubled on modulation peaks.

The dc plate-current I_{bp} rating will depend on the type of emitter used. If the filament is of pure tungsten, from which the emission usually is relatively low, then I_{bp} will have to be $\frac{1}{2}$ of I_{bt} to be sure that the crest value of plate current is equal in both cases. In the case of thoriated and oxide-coated filaments, the emission may be so high, relative to the demands in telegraph service, that the above rule need not apply. The rating can then be determined from power-input considerations.

With 100-per cent sinusoidal modulation the power input is 1.5 $E_{bp}\times I_{bp}$. If we equate this expression to the power input for telegraphy and further introduce $E_{bp}=(1/1.25)E_{bt}$

$$1.5 \times \frac{1}{1.25} E_{bt}I_{bp} = E_{bt}I_{bt}$$

$$I_{bp} = \frac{5}{6} I_{bt}.$$

We have thus determined both the plate input and I_{bp} ratings.

Grid-No. 3 input remains unchanged because there is no modulation. Grid-No. 2 input and dissipation are both 1.5 times higher in the modulated case than under carrier conditions, so both of these ratings must be two-thirds of the ratings in telegraphy.

Radio-Frequency Power Amplifier, Class B; Grid-Modulated Radio-Frequency Power Amplifier, Class C; Suppressor-Modulated Power Amplifier Class-C Telephony

These services can be grouped because of many points of similarity. Because the plate and screen voltage are unmodulated, their ratings can be the same as in telegraph service. Due to the requirements for linear modulation, grid-No. 1 voltage is never a limitation in grid modulation or class-B operation, so these ratings are usually omitted. Grid-No. 1 current is the same as in telegraphy except in class B, where the rating is omitted because it is never a limitation.

In all of these systems of modulation the efficiency varies linearly with modulation and at the carrier point, where the ratings are given, the efficiency is only one-half the peak value. With reasonable linearity of modulation, the crest efficiency is usually no more than two thirds, so that the carrier efficiency is only one third. In other words, at the carrier points, the plate loss is two thirds of the input or the input is three-halves the plate loss.

During modulation the plate loss decreases from the value at the carrier point. However, in rating the tubes we must assume long periods of operation with no modulation, so that the rating is determined by the carrier point. It is obvious that the plate dissipation ratings in these services can be the same as in telegraphy. Then the input ratings are 1.5 times the plate dissipation rating.

The plate current ratings can then be calculated from plate input and plate voltage. This calculated value is very low, and usually higher values are chosen to permit operation with full input at lower voltages. A rating equal to I_{bp} is logical and is often chosen.

Grid–No. 2 input rating usually has to be made somewhat lower than the telegraphy input rating because the screen current varies symmetrically over the modulation range. A value of two thirds the telegraph rating has been found suitable. Grid–No. 3 input rating is made the same as in telegraphy, except in suppressor modulation, where it is no limitation.

Radio-Frequency Power Amplifier, Class-B, and Grid-Modulated Radio-Frequency Power Amplifier, Class-C Television Service

This service differs from telephony service in that there is no carrier level. Three levels of rf voltage are defined. These are synchronizing peak level, corresponding to the maximum rf output voltage level, the black level, which is 75 ± 2.5 per cent of the synchronizing level, and white level, which is 0–15 per cent of synchronizing peak level. The synchronizing level is maintained very nearly 10 per cent of the time. The average picture level will, of course, vary with the material transmitted, but for rating purposes it must be assumed that black level can be maintained for indefinite periods of time.

Another difference in television service is the wide band of frequencies that must be transmitted. The tube output circuit must often be loaded heavily in order to secure the desired bandwidth. If a single tube operates into a parallel-tuned circuit of C farads and L henries and it is required to pass a band of Δf cycles per second, then the circuit must be loaded with an equivalent shunt resistance of

$$R = \frac{1}{2\pi \Delta f C} \text{ ohms.}$$

The circuit presents a load resistance of R to the tube at center of the band and $R/\sqrt{2}$ at each edge of the band, so that the amplitude response at the edges is 70 per cent of center response.

The foregoing load resistance is often well below the optimum for the tube, even when the circuit capacitance consists only of the tube output capacitance plus inevitable stray capacitance. When this is the case, limited rf plate voltage swing can be developed, and it becomes desirable to reduce the dc plate voltage or needless plate dissipation will be obtained.

With this discussion in mind we can proceed to ratings. Ratings are based on synchronizing level conditions. (The FCC rates transmitter power under these conditions.) Plate-voltage and screen-voltage ratings can be the same as in telegraphy. The grid bias rating should be defined for white level if grid modulation is used and can have the same value as in telegraphy. For class-B operation no grid bias rating is needed, as satisfactory operation is obtainable only with bias values near plate-current cutoff and, hence, far less in magnitude than the telegraph rating.

The plate-current rating generally can be the same as in telegraphy, as the conditions at synchronizing level are similar to telegraphy.

The plate-dissipation rating can clearly be the same as in telegraphy. The plate-input rating may have to be lower, depending on the plate-dissipation rating and the efficiency obtainable in television service. If the telegraph input rating, minus the power output, obtainable in television service with this input, does not exceed the plate dissipation rating, then the input rating for both classes of service can be the same.

At black level, and assuming a linear operating characteristic, the power input is down 25 per cent and the power output is down 44 per cent from synchronizing level. It can be readily shown that with efficiencies at synchronizing level, in the usual range of 50 to 60 per cent, the plate dissipation at black level is very close to the value at synchronizing level. Therefore, an increase in power-input rating at synchronizing level is not possible, even though this power input is obtained only 10 per cent of the time.

The grid–No. 2 input rating for this type of service could be higher than in telegraphy because of the short duration of synchronizing levels. However, since plate current and plate input are not higher, higher grid–No. 2 input would be of no advantage, so the rating is made the same as in telegraphy. The same reasoning applies to grid dissipation.

Audio-Frequency Power Amplifier and Modulator, Class B

The plate-voltage rating in this service is often made the same as for telegraphy. Higher values would be quite in order, however, because of the absence of rf voltages.

The plate input of a class-B amplifier varies with the signal input to the grid. If an ideal triode were available, that is, one in which plate characteristic curves were straight lines, and if the bias were adjusted exactly to cutoff, then the input would vary linearly with signal. At the same time, with ideal transformer coupling to a resistance load, the output would vary as the square of the signal up to the point where the plate swing has its maximum value. The plate dissipation, which is the difference between input and output, would then start as a straight line, as signal is increased, reach a crest at about two thirds of the full signal and then decrease up to full signal. In practical class-B amplifiers, considerable quiescent plate current (and thus also dissipation) is allowed in order to reduce distortion due to crossover from one of the push-pull tubes to the other. The peak of dissipation is thus less pronounced, although still present. Strictly, ratings should be based on this peak of dissipation. In practice, however, class-B amplifiers of this type have been used almost solely for amplification of speech and music where the average signal level is always far below the peak level, or the level at which maximum plate dissipation occurs. Because of this usage, the industry has largely adopted the practice of basing ratings on plate dissipation at full signal, which is the point of maximum plate efficiency.

The theoretical maximum efficiency is $\pi/4$ or 78.5 per cent. Actually, efficiencies are usually not over 70 per cent due to requirements for linearity. Thus, if a tube has a relatively low plate-dissipation rating, the input rating will be determined by the efficiency obtainable. The plate dissipation is made the same as in telegraphy because there are no factors which would require a different value. The maximum-signal dc plate-current rating can be the same as for telegraphy, but a somewhat lower value is often chosen because it is easier to get good linearity at lower values of current.

Audio-Frequency Power Amplifier Class A_1, Class AB_1, and Class AB_2

In high-voltage tubes, where peak voltages are important, the plate-voltage ratings for these services are usually the same as E_{bt}. In lower-voltage tubes, higher ratings can be allowed. Grid–No. 2 voltages equal to twice the rating for telephony are usually allowed in AB_1 and AB_2 services, but for A_1 service, where current flows continuously, the rating must be held down to the telegraph value to avoid overheating the grid–No. 2. Grid–No. 1 and grid–No. 3 ratings are often not given as they are not limitations.

One would expect that the plate dissipation could be the same as in telegraphy. However, under quiescent conditions, because the grid is negative and the plate voltage is at full supply value, focussing of the electron stream often takes place, with the result that localized heating of the plate occurs. If this effect is severe enough, a reduction of plate-dissipation rating in this class of service is necessary.

In AB_1 and AB_2 service, the plate-input rating is determined by efficiency and plate dissipation. The input can often be made equal to the telegraph input.

In A_1 service, the input rating is always made equal to the dissipation rating so that the tube will not be overheated in long quiescent periods.

In AB_1 and AB_2 service, the grid–No. 2 input rating is made equal to the telegraph rating so that the capabilities of this grid are not exceeded if maximum output is continuously demanded.

Oscillator and Amplifier Operation with Unfiltered Plate-Supply Voltage

If an oscillator or amplifier is operated with a dc plate voltage E_b and the plate and grid currents are I_b and I_c, respectively, then if the plate voltage is dropped to $(1/n)E_b$, the currents also drop by the factor $1/n$. Of course, if the oscillator is very poorly adjusted, such as with an abnormally low value of I_c, this relation may not be exactly followed. For all normal adjustments and with the grid bias derived from a grid resistor, it will be found that this relation holds true. The linearity of modulation obtained with plate-modulated amplifiers is largely due this phenomenon.[2]

If the plate-supply voltage is now a varying voltage, I_b, I_c, E_p all follow the plate-voltage variations linearity. This fact determines many of the steps which follow.

Case 1. Self-Rectifying Half-Wave Operation

A simplified oscillator circuit for such operation is shown in Fig. 1. The ac supply voltage rating E_{rm} can be determined first. It may be determined by one of

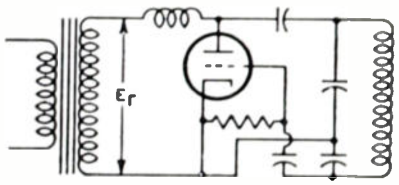

Fig. 1

three different considerations, the ability of tube to withstand positive plate potentials, the ability of the tube to withstand negative plate potentials, and the heating of insulators and conductors in the tube due to rf voltages and currents. Which one of these considerations determines the final rating depends on the tube design.

A. Peak Forward Voltage Consideration. The peak supply voltage is $\sqrt{2}E_r$. This voltage can be equated to the peak supply voltage in plate-modulated service when 100-per cent modulation is present, without creating any additional stresses on the tube. Hence,

$$\sqrt{2}\, E_{rm} = 2E_{bp}$$

$$\text{or } E_{rm} = \frac{\sqrt{2}}{K_1} E_{bt}. \qquad (1)$$

This consideration is likely to apply to high-voltage tubes in which internal or external flashover is an important factor.

B. Inverse Peak Voltage Consideration. The inverse peak plate voltage is $\sqrt{2}E_r$. Since this voltage is only about half the peak forward voltage, and since vacuum tubes usually withstand as much inverse as forward voltage, this consideration is not important.

C. Insulator and Conductor Heating. Both of these factors can be lumped together since both vary as the square of E_p. Note that rf currents are primarily capacitive charging currents caused by E_p. In this service, the tube oscillates only during alternate half cycles when the plate voltage is positive. During this half cycle

[2] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Co., Inc., 1st ed., New York, N. Y., pp. 533, 534, 539; 1943.

$$e_p = \sqrt{2}\, kE_r \sin \omega t \sin pt.$$

his expression follows directly from the opening paragaphs of this section. It can also be written

$$e_p = \frac{\sqrt{2}\, kE_r}{2} (\cos (\omega - p)t - \cos (\omega + p)t).$$

hus, e_p consists of two equal voltages of different freuencies. We are concerned with the rms value of these vo voltages and in accordance with well-known principles[3] the rms value is

$$\frac{\sqrt{2}\, kE_r}{2\sqrt{2}} \sqrt{1^2 + 1^2} = \frac{kE_r}{\sqrt{2}} \text{ (during alternate half cycles only).}$$

is to be noted that the above equations apply to positive half cycles only. The supply voltage is negative uring the intervening half cycle, and therefore the tube not generating rf voltage. The effective rf plate voltge over a whole cycle is then $1/\sqrt{2}$ the value above, or

$$E_{peff} = \frac{kE_r}{2}.$$

It is this voltage which produces rf heating of insuators, conductors, etc. This voltage must now be quated to the equivalent voltage obtained in teleraph service, at maximum ratings, thus establishing he value of E_{rm} at which equal loss occurs.

$$\frac{kE_{rm}}{2} = \frac{kE_{bt}}{\sqrt{2}}$$
$$\text{or } E_{rm} = \sqrt{2}\, E_{bt}. \tag{2}$$

D. Power Input Consideration. Plate-supply current lows only in half cycles, and due to the linearity beween supply voltage and current, the peak value of ach pulse is πI_b and

$$P_i = E_r \times \frac{\pi I_b}{\sqrt{2}} \times 1/2.$$

The $\sqrt{2}$ results from obtaining the effective value and he ½ from the fact that input power flows only one-half he time. Equating this input to the maximum rated nput for telegraph service, we get

$$P_{it} = \frac{\pi}{2\sqrt{2}} E_r I_{br}$$
$$\text{or } I_{br} = 0.9 \frac{P_{it}}{E_r}. \tag{3}$$

Equation (3) permits calculation of the maximum value of I_{br} for any value of E_r as determined from (1) or (2).

E. Peak Current Consideration. The peak supply current is πI_r. The highest peak supply current in normal service usually occurs in telephony, and $2I_{bp}$ at the peaks of modulation. Hence,

$$\pi I_{br} = 2I_{bp}$$
$$\text{or } I_{br} = \frac{0.636}{K_2} I_{bt}. \tag{4}$$

Peak instantaneous plate currents are proportional to these values. Thus, (4) permits choice of a value of I_{br} which results in peak currents no higher than obtained in telegraph service.

F. Grid Current Consideration. Due to linearity, the grid current varies with supply current and

$$\frac{I_{cr}}{I_{br}} = \frac{I_{ct}}{I_{bt}}$$
$$\text{or } I_{cr} = \frac{I_{ct}}{I_{bt}} \times I_{br}. \tag{5}$$

Thus, with I_{br} as determined from (3) or (4), the proper value of I_{cr} can be determined from the telegraph ratings of the tube.

G. Grid Voltage Consideration. Excessive rf grid voltage can cause overheating of the grid and rf leads due to capacitive currents and excessive dielectric loss in the region of grid seals. Since the rf grid voltage is proportional to the grid bias, a grid bias rating is a convenient way to limit the rf voltage. In self-rectifying half-wave operation, the grid bias voltage is a half sine wave which is at zero potential with respect to the cathode for half of each cycle. By means of the same reasoning as under 1-C, it can be shown that

$$E_{cr} = \frac{2}{\pi} E_{ct} = 0.636 E_{ct}, \tag{6}$$

for equal loss in both self-rectifying and telegraph services. There may be cases where peak grid voltages will be the limitation rather than loss considerations. In this case

$$E_{cr} = \frac{E_{ct}}{\pi} = 0.318 E_{ct}. \tag{7}$$

Case 2. Separately Rectified Half-Wave Operation

A simplified circuit as shown in Fig. 2. This case is different from Case 1 only in that the inverse plate voltage is zero. Nothing is gained by adding the rectifier tube, so this case is not of practical importance.

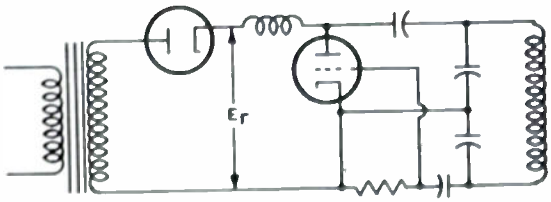

Fig. 2

[3] R. G. Hudson, "The Engineer's Manual," John Wiley and Sons, Inc., New York, N. Y., eq. 942; 1939.

Case 3. Separately Rectified Full-Wave Operation

A simplified circuit is shown in Fig. 3. This case follows very simply from Case 1, so a detailed derivation is not necessary. However, in this case it is more logical

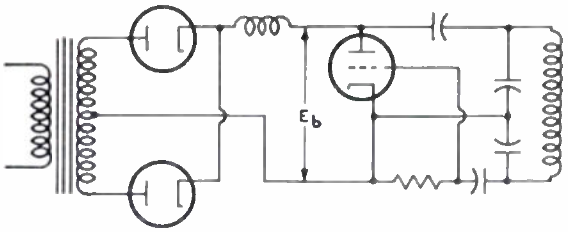

Fig. 3

to express equations in terms of the dc plate voltage E_b. If the rectifier tubes have negligible voltage drop then,

$$E_b = \frac{2}{\pi} \times \frac{\sqrt{2}\, E_r}{2}$$
$$= 0.45 E_r.$$

A. Peak Forward Voltage Consideration

$$E_{br} = \frac{4}{\pi} E_{bp} = \frac{1.27 E_{bt}}{K_1}. \qquad (8)$$

B. Insulator and Conductor Heating Consideration

$$E_{br} = 0.9 E_{bt}. \qquad (9)$$

C. Power Input Consideration

$$I_{br} = 0.81 \frac{P_{it}}{E_b}. \qquad (10)$$

D. Peak Current Consideration

$$I_{br} = \frac{4}{\pi} I_{bp} = \frac{1.27 I_{bt}}{K_2}. \qquad (11)$$

E. Grid Current Consideration

$$I_{cr} = \frac{I_{ci}}{I_{bt}} \times I_{br}. \qquad (12)$$

F. Grid Voltage Consideration

$$E_{cr} = \frac{2\sqrt{2}}{\pi} E_{ct} = 0.9 E_{ct}. \qquad (13)$$

Case 4. Self-Rectifying Full-Wave Operation

A simplified circuit is shown in Fig. 4.

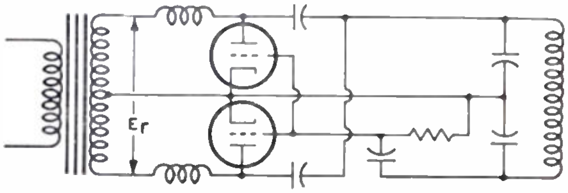

Fig. 4

A. Peak Voltage Consideration,

$$\text{Again,} \quad E_{rm} = 2\sqrt{2}\, E_{bp} = \frac{2.83}{K_1} E_{bt}. \qquad (14)$$

B. Inverse Voltage Consideration. The inverse voltage in this case is the sum of the negative supply voltage peak plus the rf plate-voltage swing present at the crest of the supply voltage. Thus, the inverse peak is

$$E_{bi} = \frac{\sqrt{2}\, E_r}{2} + k \frac{\sqrt{2}\, E_r}{2}$$

$$\text{or } E_{rm} = \frac{\sqrt{2}}{1+k} E_{bim}.$$

If the tube will withstand as much inverse voltage as forward voltage (which is usually the case), then this expression can be written in terms of the forward peak-voltage, which occurs on the crest of a plate-modulated cycle. The forward peak voltage is

$$2(E_{bp} + kE_{bp}) = 2(1+k)E_{bp}.$$

This expression can be substituted for E_{bim} and we obtain

$$E_{rm} = \frac{\sqrt{2}}{1+k} \times 2(1+k)(E_{bp}) = \frac{2.83}{K_1} E_{bt}. \qquad (15)$$

C. Insulator and Conductor Heating Consideration. It will be noted that this case is very much like Case 1, except that rf voltage is present on both grids and plates for the complete supply-voltage cycle. Hence,

$$E_{peff} = \frac{k\left(\dfrac{E_r}{2}\right)}{\sqrt{2}},$$

and this relation is equated to E_p. Thus

$$\frac{k\left(\dfrac{E_{rm}}{2}\right)}{\sqrt{2}} = \frac{kE_{bt}}{\sqrt{2}} \qquad (16)$$
$$\text{or } E_{rm} = 2E_{bt}.$$

D. Power Input Consideration. This relation is the same as in Case 1, except that the supply voltage per plate is $E_{r/2}$. Therefore,

$$I_{br} = 1.8 \frac{P_{it}}{E_r} \text{ per tube.} \qquad (17)$$

E. Peak Current Consideration. This relation is again the same as in Case 1, so that

$$I_{br} = \frac{0.636}{K_2} I_{bt}. \qquad (18)$$

F. Grid Current Consideration. This relation is also the same as in Case 1:

$$I_{cr} = \frac{I_{ct}}{I_{bt}} \times I_{br}. \qquad (19)$$

It should be noted that in this type of operation, each grid draws current on both half-cycles of supply voltage, because rf grid voltage is present on both half-cycles. The grid current which flows in a tube when its plate voltage is negative is wasted and it causes unnecessary grid heating. Excess heating will not occur if the above relation is used. However, the grid current available for excitation on the working half-cycle is reduced, and the oscillator may not operate as efficiently. Circuits have been devised to overcome this difficulty, but they are rather complex.

G. Grid Voltage Consideration

$$E_{cr} = \frac{2\sqrt{2}}{\pi} E_{ct} = 0.9E_{ct}.$$

Case 5. Self-Rectifying Full-Wave Operation **with a** *Smoothing Choke*

A simplified circuit is shown in Fig. 5 and an equivalent circuit is given in Fig. 6.

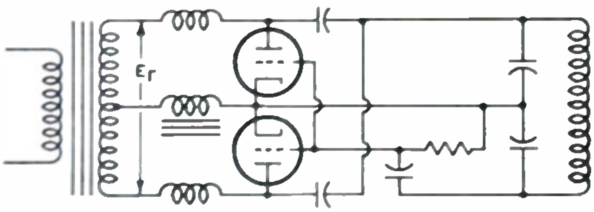

Fig. 5

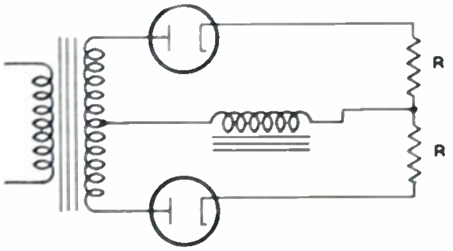

Fig. 6

A. Peak Voltage Consideration. The forward voltage supplied to a conducting tube is the average value of half the transformer voltage, assuming no dc resistance in the choke or transformer,

$$\frac{\sqrt{2}\, E_r}{2\pi}.$$

The conducting tube has this voltage applied during alternate half cycles. This condition is equivalent to 100 per cent modulation with a square wave. Therefore, it can be equated to the peak supply voltage in telephony

$$\frac{\sqrt{2}\, E_{rm}}{2\pi} = 2E_{bp} \tag{20}$$

$$\text{or } E_{rm} = \frac{4\pi}{\sqrt{2}} E_{bp} = \frac{8.9}{K_1} E_{bt}. \tag{21}$$

B. Inverse Peak Voltage Consideration. In this circuit R is the equivalent resistance of the oscillator of E_b/I_b. If the choke has a large reactance compared to R, then essentially constant current flows in the choke and each tube conducts a square pulse of current of 50 per cent duty factor. The amplitude of this pulse is $2\times I_b$ and the voltage drop from plate to cathode is E_b. At the peak of the transformer voltage ($\sqrt{2}E_r$), the tube which is not conducting has an inverse voltage of $\sqrt{2}E_r - E_b$ applied to its plate. However, since the plate of this tube is coupled to the plate of the other tube for rf frequencies, the true inverse peak voltage on the nonconducting tube is

$$E_{bi} = \sqrt{2}\, E_r - E_b + kE_b$$

$$E_{bi} \cong \sqrt{2}\, E_r \tag{22}$$

$$\text{or } E_{rm} = 0.7E_{bim}.$$

As this is a rather large voltage, the tube design must be checked to see if E_{rm} will be determined by this consideration.

C. Insulator and Conductor Heating Consideration. This loss is proportional to e^2. During the conducting cycle it is proportional to

$$\left(\frac{\sqrt{2}\, E_r}{2\pi}\right)^2.$$

However, since the two anodes are directly connected, so far as rf currents are concerned, this loss is also present over a full cycle.

In telegraph service the loss is proportional to E_{bt}^2. Equating losses,

$$\left(\frac{\sqrt{2}\, E_{rm}}{2\pi}\right)^2 = E_{bt}^2 \tag{23}$$

$$\text{or } E_{rm} = \sqrt{2}\,\pi E_{bt} = 4.44E_{bt}.$$

E_{rm} is thus determined by this equation rather than (21).

D. Power Input Consideration. The power input is

$$P_i = \frac{\sqrt{2}\, E_r}{2\pi} I_b.$$

Equating this expression to the maximum rated plate input for telegraph service

$$\frac{\sqrt{2}\, E_r}{2\pi} I_{br} = P_{it}$$

$$I_{br} = \frac{\sqrt{2}\,\pi P_{it}}{E_{rm}} = \frac{4.44P_{it}}{E_{rm}}. \tag{24}$$

E. Peak Current Consideration. The peak plate-supply current is $2I_b$ since the current is off half the time. This value can be equated to twice I_{bp} for equal peak currents. Then

$$I_{br} = I_{bp} = \frac{I_{bt}}{K_2}. \tag{25}$$

F. Grid Current Consideration. From what has gone before we can expect

$$I_{cr} = \frac{I_{ct}}{I_{bt}} \times I_{br}. \tag{26}$$

However, this equation does not take into account the fact that the nonconducting tube also draws current, so higher values of I_{cr} must be allowed unless a circuit is used which provides grid-current cutoff during the negative plate-voltage cycle.

G. Grid Voltage Consideration. The rf grid-voltage amplitude does not vary with time, hence the grid bias rating can be the same as for telegraphy

$$E_{cr} = E_{ct}. \tag{27}$$

Reduction of Ratings at High Frequencies

All ratings discussed so far have been independent of frequencies. As the frequency is increased, it is well known that the efficiency of tubes decreases because of transit-time effects. If the power input is held constant as the frequency is increased, a point will be reached beyond which the plate-dissipation rating of the tube is exceeded. It is obvious, therefore, that above the limiting frequency, plate input must be decreased. This practice is generally used in the industry. Our object here is to develop a rational basis for such de-rating.

The input can be decreased in two main ways: by a reduction of plate voltage, or a reduction in plate current. The first method is the best, because any reduction of plate current would have to be accomplished by reduction in the positive grid-voltage crest which would result in an increase in electron transit time and thus an unnecessary further reduction of plate-circuit efficiency. It is best, therefore, to maintain the plate current at its full value and reduce the plate input by means of a plate-voltage reduction. This reduction is commonly made by specifying the same percentage of reduction for both input and voltage.

Law[4] has found an empirical relationship for plate circuit efficiency η. Law's relationship is

$$\eta = \eta_0\left(1 - \frac{K_3 fd}{\sqrt{E_b}}\right).$$

Instead of E_b, it is even more accurate to write e_c, the peak positive grid voltage. Then $K_3d/\sqrt{e_c}$ becomes a constant. This relationship then shows that the curve of efficiency is a straight line starting at $\eta=\eta_0$ when $f=0$ and extending down to the zero axis as shown in Fig. 7. If we know the efficiency and corresponding frequency at two points, the straight line is completely determined. From an inspection of Fig. 7, it is obvious that

$$\frac{\eta_1 - \eta}{\eta_1 - \eta_2} = \frac{f - f_1}{f_2 - f_1}$$

or

$$\eta = \eta_1 - \left(\frac{\eta_1 - \eta_2}{f_2 - f_1}\right) \times (f - f_1). \tag{28}$$

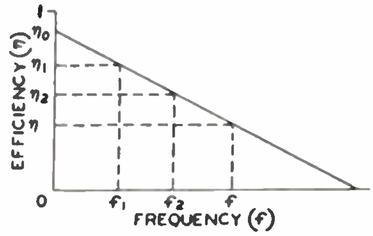

Fig. 7

In general, η_1 at the frequency f_1 must be found by test and therefore is known. If the second known efficiency η_2 is η_0 at $f=0$, then the expression for η is

$$\eta = \eta_1 - \frac{\eta_0 - \eta_1}{f_1}(f - f_1)$$
$$= \eta_0 - (\eta_0 - \eta_1)\frac{f}{f_1}. \tag{29}$$

If the plate dissipation rating is not to be exceeded, then

$$P_i = \frac{P_d}{1 - \eta},$$

and the ratio of this input to the maximum input rating at f_1 is

$$\frac{P_i}{P_{it}} = \frac{1 - \eta_1}{1 - \eta}. \tag{30}$$

From (29) and (30) the percentage of allowable input at frequencies above f_1 can be calculated. From earlier discussion, this percentage also applies to plate voltage, or

$$\frac{E_b}{E_{bt}} = \frac{1 - \eta_1}{1 - \eta}. \tag{31}$$

Although (30) and (31) have been developed for class-C telegraph service, they are equally applicable to other classes of service. It should be noted that since η_1 is usually the same in class-C telegraphy and class-C plate-modulated services, the same allowable percentages of input power and plate voltage is obtained for both services at a given frequency. This relationship is not true of grid-modulated and class-B linear amplifiers, since η_1 is then always of the order of 0.33 or less.

So far, it has been assumed that plate dissipation is the limiting factor in determining high-frequency ratings. There may be other factors and one of these will be discussed, namely, heating at the metal-to-glass seals due to capacitive currents. These currents are proportional to plate voltage and frequency. The hf resistance of the metal is proportional to the square root of frequency, if we assume that the depth of penetration is

[4] R. R. Law, "Electronics of ultra-high-frequency triodes," Proc. I.R.E., vol. 37, pp. 273–274; March, 1949.

small compared to metal thickness. With these factors in mind, it is readily demonstrated that to hold constant the power loss in the seals due to capacitive currents,

$$\frac{E_b}{E_{bt}} = \left(\frac{f_1}{f}\right)^{5/4}. \qquad (32)$$

In general, (32) calls for a much more rapid decrease in plate voltage than (31), and therefore tubes are usually designed so that seal heating is not a factor at f_1. Tests can then be made at f_1 and with constant plate dissipation to determine how high E_b can be when the maximum safe seal temperature is reached. This value can then be used in (32) in place of E_{bt}. If this value is sufficiently greater than E_{bt}, the allowable value of E_b will be higher than determined from (31), so that the ratings are determined by (31). There may also be cases where (31) applies over part of the frequency range and (32) beyond that range. If (31) cannot be applied at all because the seals are already up to maximum temperature at f_1, an unbalanced tube design is indicated. In some cases no change in design is necessary, if forced-air cooling can be introduced without undue expense.

From the foregoing discussion, it is clear that the limiting factor in high-frequency ratings should always be electron transit time. Then (29) and (31) determine the high-frequency ratings.

Conclusions

A rational system for rating high-vacuum power tubes has been described, based on knowledge of the safe ratings for the tubes in radio-frequency power amplifier and oscillator service. The rating factors for other classes of service are tabulated in Table II. A system of derating tubes at high frequencies is also developed.

TABLE II

Summarized Tabulation of Rating Factors

Maximum Ratings

Where a number is shown, the telegraphy value is to be multiplied by this factor to obtain the desired rating. Where the tabulated value is not a number but involves some other rating, the desired rating is given by the calculation indicated.

Type of service	E_b	I_b	P_i	P_d	E_{c3}	P_{g3}	E_{c2}	P_{g2}	E_{c1}	I_{c1}
Class-C Telegraphy	1	1	1	1	1	1	1	1	1	1
Class-C Plate Mod.	0.8*	0.833 or 0.5†	0.67‡	0.67	1	1	1	0.67	1	1
Low-Level Modulated Amplifiers	1§	0.833 or 0.5†	1.5 P_d	1	1	1	1	0.67	—	1
Television Amplifiers	1§	1	1§	1	1	1	1	1	—	1
Class-B Audio Amplifiers	1	1	1§	1	1	1	1	1	—	—
Class-A Audio Amplifiers	1	1	P_d	1‖	1	1	1	1	—	—
Class-AB Audio Amplifiers	1	1	1§	1	1	1	1	1	—	—
Self-Rectifying Oscillator (Case 1)	$\frac{E_{rm}}{\frac{1.41}{K_1}}$¶ or $\frac{E_{rm}}{1.41}$	$\frac{I_{br}}{\frac{0.9 P_{it}}{E_{rm}}}$ or $\frac{I_{br}}{\frac{0.636}{K_2}}$	1	1	—	—	—	—	$\frac{E_{cr}}{0.636}$¶ or $\frac{E_{cr}}{0.318}$	$\frac{I_{cr}}{K_4 I_{br}}$
Separately Rectified Oscillator (Case 3)	$\frac{E_{br}}{0.9}$ or $\frac{E_{br}}{\frac{1.27}{K_1}}$	$\frac{0.81 P_{it}}{E_{br}}$† or $\frac{1.27}{K_2}$	1	1	—	—	—	—	0.9	$K_4 I_{br}$
Self-Rectifying Oscillator Full-Wave (Case 4)	$\frac{E_{rm}}{\frac{2.83}{K_1}}$¶ or $\frac{E_{rm}}{2}$	$\frac{1.8 P_{it}}{E_{rm}}$† or $\frac{0.636}{K_2}$	1	1	—	—	—	—	0.9	$K_4 I_{br}$
Self-Rectifying Oscillator With Smoothing Choke (Case 5)	4.44¶ or 0.7 E_{btm}	$\frac{4.44 P_{it}}{E_{rm}}$† or $\frac{I_{bt}}{K_2}$	1	1	—	—	—	—	1	$K_4 I_{br}$

* In case of high voltage tubes a lower value may be necessary. In this case $E_{bp} = \frac{1}{2}$ max safe instantaneous plate voltage.
† Use the lower value only for tubes with limited emission capability, such as those using tungsten filaments.
‡ In case of tungsten filaments, the factor may be less, being determined by $E_{bp} \times I_{bp}$.
§ A lower factor may be necessary depending on the efficiency obtainable.
‖ If severe focussing is present a lower factor may be necessary.
¶ Use the lower value only in cases where maximum safe instantaneous voltage is the limiting factor.

Electron Tube Performance with Large Applied Voltages*

ABD EL-SAMIE MOSTAFA†

Summary—This paper contains an analytic treatment of vacuum-tube performance with large applied voltages. The method presented uses the analysis of the transconductance characteristic of electron tubes, instead of the usual analysis of the plate current-grid voltage characteristic. By choosing a certain function and its Fourier expansions over a suitable interval of the $g-e_g$ characteristic, the analysis of electron tubes with large applied voltages is simplified. Experimental verification is carried out for a triode.

I. Introduction

Electron tubes have become of increasing importance due to their extension into the fields of measurement and control. Therefore, a need is felt for a rigorous but simplified treatment for quantitative evaluation of electron tube performance. The method presented herein uses a certain function and its Fourier expansions over a suitable interval of the transconductance-grid voltage characteristic of the tube, instead of the usual analysis of the plate current-grid voltage characteristic. It is felt that the former leads to a better analysis of the tube performance.

In a paper by Barrow,[1] the analysis of vacuum-tube performance with large applied voltages was considered using Fourier expansions of the plate current-grid voltage characteristic. However, selecting the transconductance g rather than the plate current i_p for the expansion reduces the fundamental period to half its value, and consequently the number of terms required for the expansion will be highly reduced. In a recent paper,[2] the transconductance-grid voltage characteristic has been used in evaluating electron tube performance, but in a different and limited way as to the position and amplitude of the applied signal.

The method presented herein has been used by the author in analyzing class-C amplifiers and proved to be very useful, especially when the operating range is large. The method can be directly applied to the tube as a rectifier, modulator, oscillator, class-B and class-A amplifiers, as well as a frequency multiplier.

Although the method has been applied to a triode, we shall start with the simple case of a pentode where the internal impedance is high compared with the load impedance, and hence the plate current may be satisfactorily represented as a function of the grid voltage only.

* Decimal classification: R331.5×R262.5. Original manuscript received by the Institute, December 6, 1949; revised manuscript received, May 29, 1950.

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[1] W. L. Barrow, "Contribution to the theory of non-linear circuits with large applied voltages," Proc. I.R.E., vol. 22, p. 946; August, 1934.

[2] T. Slonczewski, "Transconductance as a criterion of electron tube performance," *Bell Sys. Tech. Jour.*, vol. 28, pp. 315–328; April, 1949.

II. Theoretical Analysis

Figs. 1(a) and 1(b) represent $i_p \to e_g$ as well as $g \to e_g$ characteristics, respectively, where,

e_g = total grid voltage $= v_g - E_c$
i_p = total plate current
E_c = grid bias voltage
v_g = ac input voltage applied to the grid
g = transconductance $= di_p/de_g$.

Let the operating range be extended from beyond cutoff, point b, to the end of the linear portion of the $i_p \to e_g$ characteristic, point d. The $g \to e_g$ characteristic can be represented by either of the following methods:

1. Power series: this will be cumbersome because of the required large number of terms. Even if the operating range is small, the work would be tedious, as was stated by Slonczewski in his paper.

2. Fourier series: this requires a large number of terms to get good approximation, due to the sharp jump in the $g \to e_g$ characteristic.

As the discontinuity is responsible for the rate of convergency of the Fourier series, the best thing to do, therefore, is to avoid this discontinuity by "an artificial extension" as shown in Fig. 1. The artificial extension is the dashed portion dfh in the $i_p \to e_g$ characteristic and the corresponding portion $d'f'h'$ in the $g \to e_g$ characteristic. $d'f'h'$ is constructed so that the resulting $g \to e_g$ characteristic which is represented by $a'b'c'd'f'h'$ is a symmetrical curve about the axis XY, passing through the middle point of the linear portion of the $i_p \to e_g$ characteristic (Fig. 1). In this way, the $g \to e_g$ character-

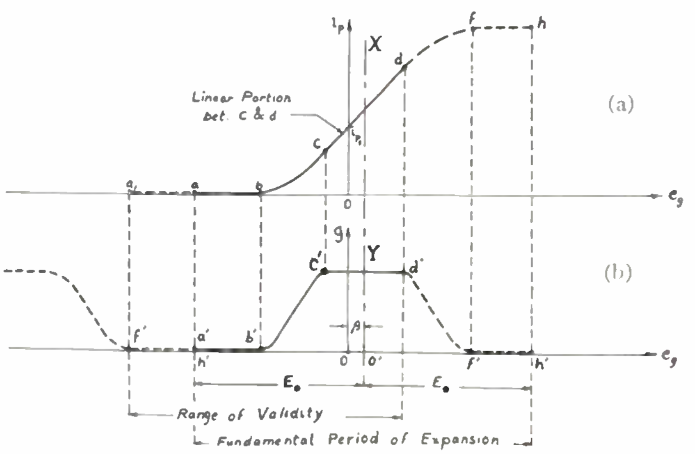

Fig. 1—Diagram showing $i_p \to e_g$ and $g \to e_g$ characteristics.

istic can be represented by a highly convergent trigonometric polynomial. The artificial extension will have another advantage, namely, the increase of the range of validity of the operating signal, i.e., the latter can cover the range a_1abcd of the $i_p \to e_g$ characteristic and not only the range $abcd$.

Now, if O' is chosen as the origin of harmonic analysis, cosine terms will only exist; the angle corresponding to OO' will be added as a constant, and therefore all voltages will be measured from zero value of e_g.

Let harmonic analysis give

$$g = di_p/de_g = a_0 + \sum_{n=1} a_n \cos \frac{n\pi}{E_0}(e_g - \beta). \quad (1)$$

The upper limit of the summation depends on the number of terms in the expansion required for satisfactory representation, but generally it is infinity, and where

β = voltage corresponding to OO',

$2E_0$ = voltage corresponding to the fundamental period of the harmonic analysis, and

a_0, a_1, a_2, a_3, etc., are constants to be determined from the harmonic analysis. Integration gives,

$$i_p = b_0 + a_0 \cdot e_g + \sum_{n=1} a_n \frac{E_0}{n\pi} \sin \frac{n\pi}{E_0}(e_g - \beta), \quad (2)$$

where b_0 = constant of integration to be determined from the value of i_p at $e_g = 0$. At $e_g = 0$, $i_p = i_{p0}$ (Fig. 1(a)), therefore,

$$b_0 = i_{p0} + \sum_{n=1} a_n \frac{E_0}{n\pi} \sin \frac{n\pi}{E_0}\beta. \quad (3)$$

Let $v_g = V \cos wt$, therefore,

$$e_g - \beta = V \cos wt - (E_c + \beta). \quad (4)$$

Substitution of equations (3) and (4) in (2) gives,

$$i_p = \left\{ i_{p0} + \sum_{n=1} a_n \frac{E_0}{n\pi} \sin \frac{n\pi}{E_0}\beta - a_0(E_c + \beta) \right\} + a_0 \cdot V \cos wt + \sum_{n=1} a_n \frac{E_0}{n\pi} \sin \left\{ \left(\frac{n\pi V}{E_0} \cos wt \right) - \frac{n\pi}{E_0}(E_c + \beta) \right\}. \quad (5)$$

But

$$\sin \{(V \cos wt) - E\} = \sin (V \cos ut) \cdot \cos E - \cos (V \cos wt) \cdot \sin E$$
$$= -J_0(V) \cdot \sin E + \left\{ 2(-1)^{k+1} \sum_{k=1} J_{2k-1}(V) \cdot \cos E \right\} \cos (2k - 1)wt - \left\{ 2(-1)^k \sum_{k=1} J_{2k}(V) \cdot \sin E \right\} \cos 2kwt,$$

where $J_k(V)$ is a Bessel function of the first kind of order k and and argument V, substitution in (5) gives

$$i_p = I_b + I_1 \cos wt + I_2 \cos 2wt + I_3 \cos 3wt + \cdots + I_r \cos rwt + \cdots, \quad (6)$$

where I_b = mean value of plate current

$$= i_{p0} - a_0(E_c + \beta) + \sum_{n=1} a_n \frac{E_0}{n\pi} \left\{ \text{rin} \frac{n\pi\beta}{E_0} - J_0\left(\frac{n\pi V}{E_0}\right) \cdot \sin \frac{n\pi}{E_0}(E_c + \beta) \right\}. \quad (7)$$

The value of the plate current with no applied signal can be obtained from (7) by putting $V = 0$, which is given by

$$i_{p0} - a_0(E_c + \beta) + \sum_{n=1} a_n \frac{E_0}{n\pi} \left\{ \sin \frac{n\pi\beta}{E_0} - \sin \frac{n\pi}{E_0}(E_c + \beta) \right\}. \quad (8)$$

Equations (7) and (8) gives the rectified current

$$= \sum_{n=1} a_n \frac{E_0}{n\pi} \left\{ 1 - J_0\left(\frac{n\pi V}{E_0}\right) \right\} \sin \frac{n\pi}{E_0}(E_c + \beta). \quad (9)$$

I_1 = fundamental amplitude of plate current

$$= V \left\{ a_0 + \sum_{n=1} a_n \frac{2J_1\left(\frac{n\pi V}{E_0}\right)}{\frac{n\pi V}{E_0}} \cdot \cos \frac{n\pi}{E_0}(E_c + \beta) \right\} \quad (10)$$

I_r = amplitude of the rth harmonic, where $r \neq 1$

$$= 2 \sum_{n=1} a_n \frac{E_0}{n\pi} J_r\left(\frac{n\pi V}{E_0}\right) \cos \frac{n\pi}{E_0}(E_c + \beta), \quad r \text{ being odd,} \quad (11a)$$

$$= -2 \sum_{n=1} a_n \frac{E_0}{n\pi} J_r\left(\frac{n\pi V}{E_0}\right) \sin \frac{n\pi}{E_0}(E_c + \beta), \quad r \text{ being even.} \quad (11b)$$

Although the operating range may be very large, due to the high convergency of the trigonometric polynomial (1) which is due to the introduced artificial extension, three or four terms are usually sufficient to give satisfactory results. Therefore, the calculations of (7), (8), (9), (10), and (11) are highly simplified. The Bessel coefficients $J_0(z)$, $J_1(z)$, $J_2(z)$, etc., and the recurrence relation $\{2J_1(z)/z\}$ can be found in any text book on Bessel functions.[3]

In the case of a triode, the composite characteristic which is the relation between the plate current i_p and the composite voltage e' ($= e_g + e_p/\mu$, referred to the grid side) can be assumed as one curve for most typical triodes, so long as the grid current is not considerable and secondary emission is much reduced, and unless the

[3] N. W. McLacklan, "Bessel Functions for Engineers," Geoffrey Cumberlege, London, p. 42; 1946.

minimum plate potential is equal to or smaller than the maximum grid potential. e_p and e_g are the total plate and grid potentials, respectively, and μ is the amplification factor. Therefore, it can be assumed that there is only one composite characteristic in the usual working range for a given tube. The analysis for a pentode can be directly applied to the case of a triode with the substitution of e' instead of e_g. The analysis will be simple if the output circuit is purely resistive.

In class-C amplifiers, the plate and grid circuits consist of tuned circuits, the load impedance in the plate circuit is very low at the harmonics and the output voltage can be satisfactorily assumed sinusoidal. If tuning and neutralization are perfect and if the operating frequency is not so high but that the transit time of electrons can be neglected, the output voltage can be assumed in antiphase with the exciting voltage. Therefore,

$$
\begin{aligned}
e' &= \text{lumped voltage, referred to grid side} = e_g + e_p/\mu \\
&= (V_g \cos wt - E_c) + \frac{1}{\mu}(E_b - V_p \cos wt) \\
&= (V_g - V_p/\mu)\cos wt + \left(\frac{E_b}{\mu} - E_c\right),
\end{aligned}
$$

where E_b and E_c are the polarizing plate and grid potentials, respectively. V_g and V_p are the amplitudes of the exciting (grid) and the output (plate) voltages. In the above equations, substitute $(E_b/\mu - E_c)$ and $(V_g - V_p/\mu)$ for E_c and V, respectively. The mean value I_b and the fundamental amplitude I_1 of the plate current can therefore be calculated from (7) and (10), respectively. Input power to plate circuit $= P_i = E_b \cdot I_b$, and output power $= \frac{1}{2} I_1^2 R_L$, where R_L = dynamic resistance of plate circuit at the fundamental frequency. It should be noted that in this case the solutions of (7), (8), (9), (10), and (11) will not be direct. For given values of E_b, E_c, V_g, and R_L, (10) can be solved by assuming several values of V_p and solving for I_1. The solution is that which satisfies the relation $V_p = I_1 \cdot R_L$, interpolation or otherwise is normally required.

III. Experimental Verification

A power supply frequency method is adopted where ordinary wattmeters, ammeters, and voltmeters can be used. The only correction at radio frequency is that of interelectrode capacities which can be included easily with the various circuit capacities. In this way, an equivalent way of introducing the load is effected by injecting in the plate circuit a sinusoidal voltage which is in antiphase with the exciting voltage.

An experimental small power tube (2-A-3 triode) was employed for test purposes and worked as a class-C amplifier. The $i_p \rightarrow e'$ characteristics were determined, following a method similar to that of Guiseppe Gramagli.[4] The curves are very approximately coincident,

[4] Guiseppe Gramagli, "The plotting of the characteristics of transmitting valves," *International Broadcast and Sound Engineer*, vol. 1, p. 134; 1937.

$$
\frac{di_p}{de'} = 3.03 + 4.60 \cos\frac{\pi}{E_0}(e'-\beta) + 1.564 \cos\frac{2\pi}{E_0}(e'-\beta) - 0.550 \cos\frac{3\pi}{E_0}(e'-\beta) - 0.793 \cos\frac{4\pi}{E_0}(e'-\beta) \ldots (a)
$$

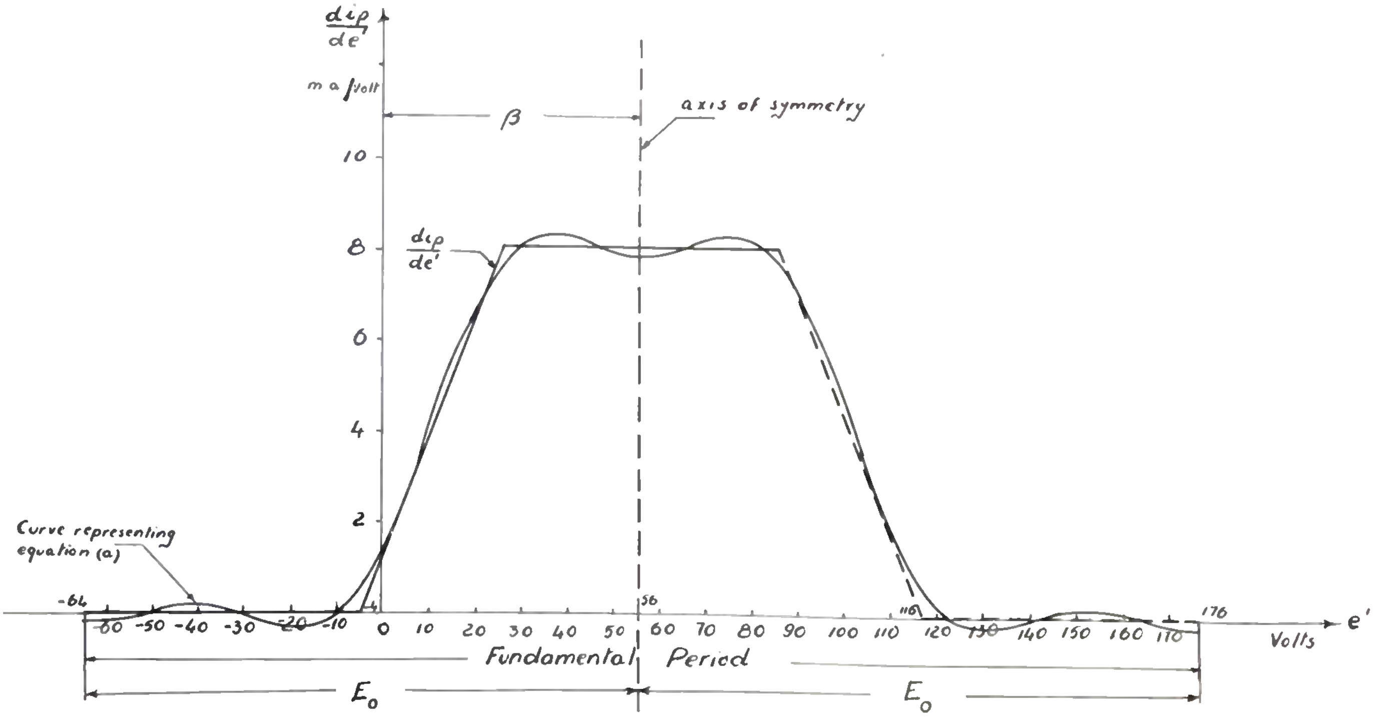

Fig. 2—Harmonic analysis.

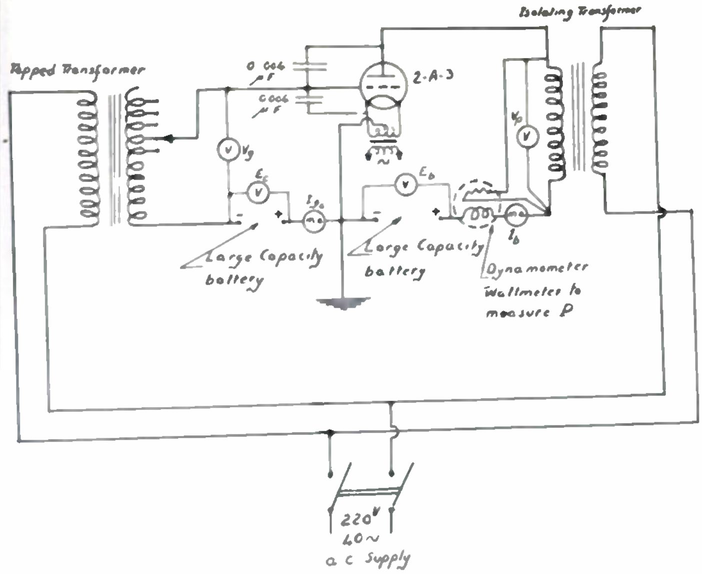

Fig. 3—Connection diagram showing class-C amplifier performance.

resulting in one composite curve, so long as the plate potential is not less than 100 volts. From the $i_p \rightarrow e'$ characteristic, di_p/de' is obtained and extended artificially as was stated before. Harmonic analysis is given in Fig. 2.

A connection diagram is shown in Fig. 3. The polarizing potentials E_b and E_c are obtained from large capacity batteries. Two transformers fed from the same supply are used to give two sinusoidal voltages in antiphase. Condensers of 0.004 μf are used between plate, grid, and filament to eliminate any possibility of high-frequency oscillation.

Fig. 4 shows what may be termed variable load characteristics for a constant output voltage. There is good agreement between experimental results and theoretical predictions. For the sake of comparison, theoretical curves based on treating the composite characteristic as a straight line[5] are also shown.

[5] This is the method adopted in W. L. Everitt, "Optimum operating conditions for class-C amplifiers," PROC. I.R.E., vol. 22, p. 152; February, 1934.

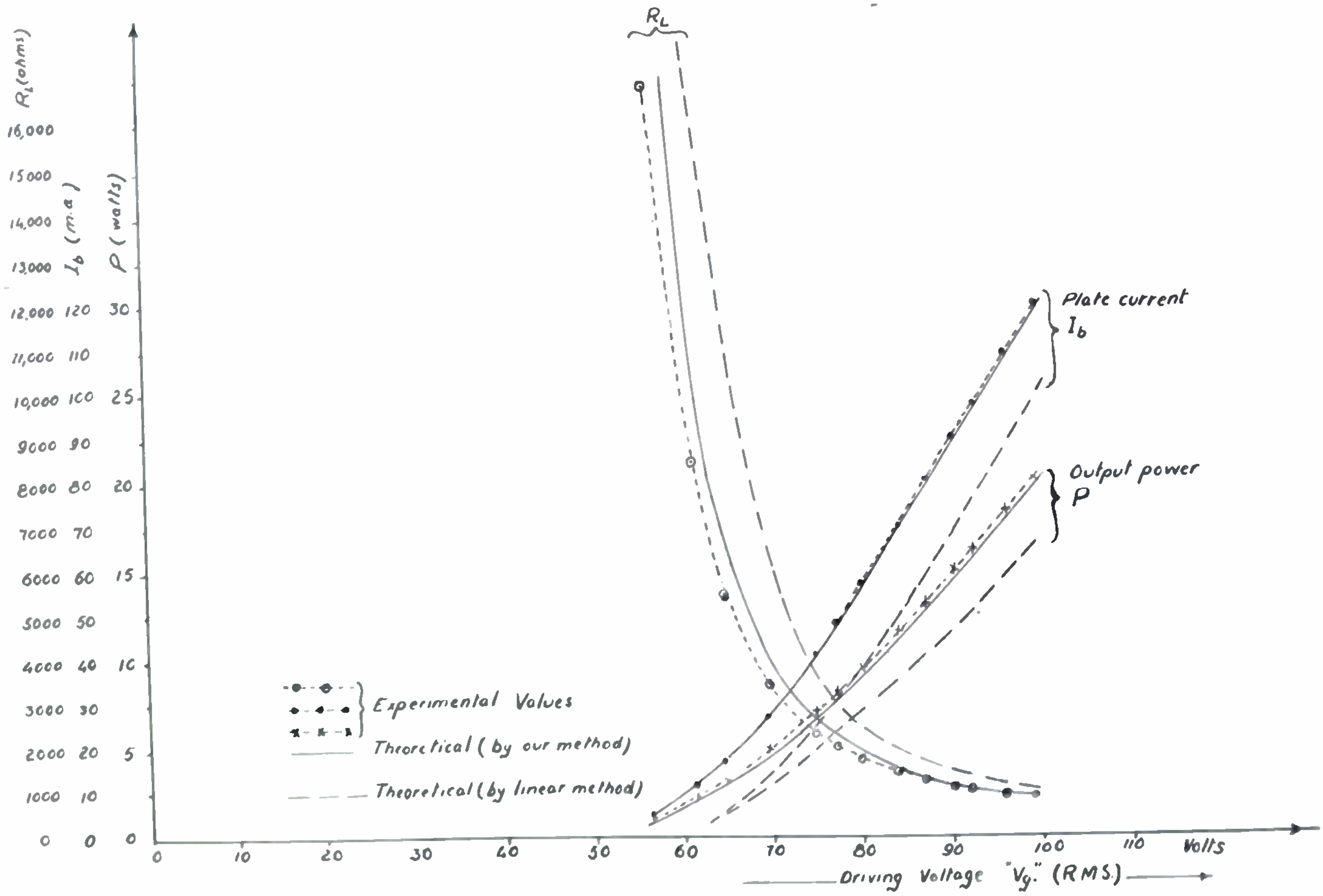

Fig. 4—Variable load characteristics for a constant output voltage.

A 300- to 4,000-Kilocycle Electrically Tuned Oscillator*

A. I. PRESSMAN†, MEMBER, IRE, AND J. P. BLEWETT†, ASSOCIATE, IRE

Summary—An oscillator is described whose frequency is electrically controllable over a 13-to-1 frequency range. Tuning is accomplished by saturation of a ferromagnetic core in the inductance of the tank circuit. The output is constant in amplitude to within 4 per cent and is reproducible in frequency to better than 0.1 per cent.

Introduction

In the three-billion-volt Cosmotron now under construction at the Brookhaven National Laboratory, protons will be accelerated in a race-track-shaped orbit from an energy of three million to about three billion electron volts. As the protons gain energy, there is an increase in their frequency of revolution and the value of magnetic field required to keep them on arcs of a circle of constant radius. The orbital frequency is related to the instantaneous value of magnetic field by

$$F = 4.304B[B^2 + (3{,}422)^2]^{-1/2}, \qquad (1)$$

where B is in gauss and F in Mc. This relationship is shown in Fig. 1. The anticipated B-versus-time cycle will be approximately a linear rise from zero to fourteen thousand gauss in one second (the acceleration time) with a five-second duty cycle. For a proton injection energy of 3 Mev, the orbital frequency will vary from 343 to 4,181 kc with a frequency-versus-time curve shaped much like the above F-versus-B curve and with, of course, the same five-second duty cycle. In order to avoid losing protons to the walls of the 36-inch wide vacuum chamber in which they revolve, the frequency of the source which accelerates them must track the magnetic field to within 0.3 per cent of the relation in (1). For 3-Mev injection, the frequency modulation cycle must start at 343 kc with a dF/dt of 20 Mc per second per second with a delay of not more than 50 microseconds after the magnetic field has reached 274 gauss.

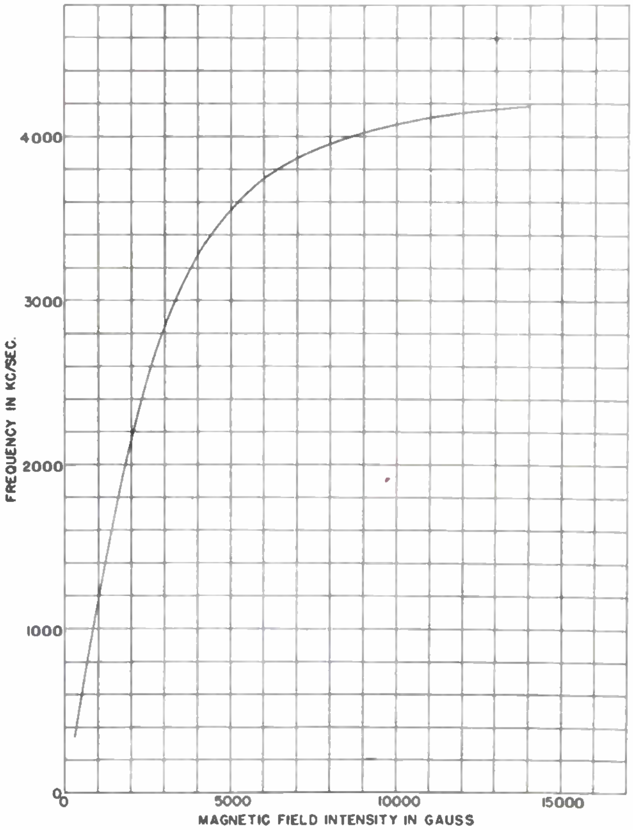

Fig. 1—Proton orbital frequency versus magnetic field intensity.

Early in the design of the machine, it was decided that these severe tuning, timing, and magnetic field tracking specifications could more easily be met by an electronically, rather than a mechanically, swept oscillator. Such an electronically tuned oscillator would be driven by an electronic analogue computer whose input signal would be a measure of magnetic-field intensity and whose output voltage would be determined by a properly shaped nonlinear impedance element. The output-input voltage characteristic of the nonlinear impedance element would be fixed by the F-versus-B relation of (1) and the frequency-versus-modulation voltage characteristic of the oscillator.

The development of an oscillator, electronically tunable over a 13-to-1 frequency range from 343 to 4,181 kc with 0.3-per cent stability and reproducibility presented a difficult problem. Tests were performed on a circuit which used beating reflex klystrons with a sweep voltage on one repeller. This type of oscillator covered the required range, but had only barely adequate frequency stability. A voltage-tuned multivibrator was also built which covered the range easily, but was not reproducible with respect to tube changes.

The oscillator finally chosen is a saturable inductance LC oscillator making use of the unique properties of Ferroxcube,[1,2] a ferromagnetic mixture of iron, man-

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† Brookhaven National Laboratory, Upton, N. Y.

[1] J. L. Snoek, "New Developments in Ferromagnetic Materials," Elsevier Publications, New York, N. Y.; 1946.
[2] V. D. Landon, "The use of ferrite cored coils as converters, amplifiers, and oscillators," *RCA Rev.*, vol. 10, p. 387; September, 1949.

ganese, and zinc oxides, manufactured by the Philips Company. This material has a high permeability and Q (Q being defined as the Q of a toroidal coil wound on a Ferroxcube core) well up into the radio frequencies. Its permeability can be reduced to 1/250 of maximum value by saturating fields as low as 50 ampere turns per cm. Curves of permeability and Q versus frequency are shown in Fig. 2. The variation of permeability with field as shown in Fig. 4 and throughout the entire range of the oscillator Q never falls below 25. The material has a fairly high temperature coefficient of permeability. The coefficient is a function of saturating field but is never more than -1.3 per cent per degree centigrade between 20° and 50°C. This has necessitated temperature controlling the inductance to achieve the required oscillator stability and reproducibility.

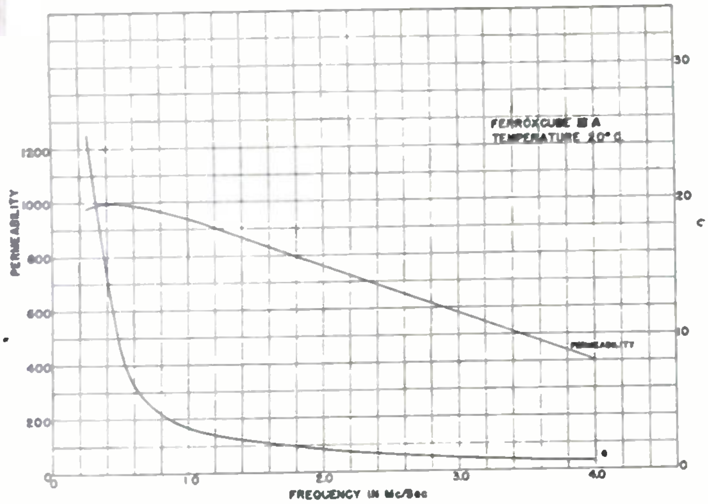

Fig. 2—Permeability and Q versus frequency.

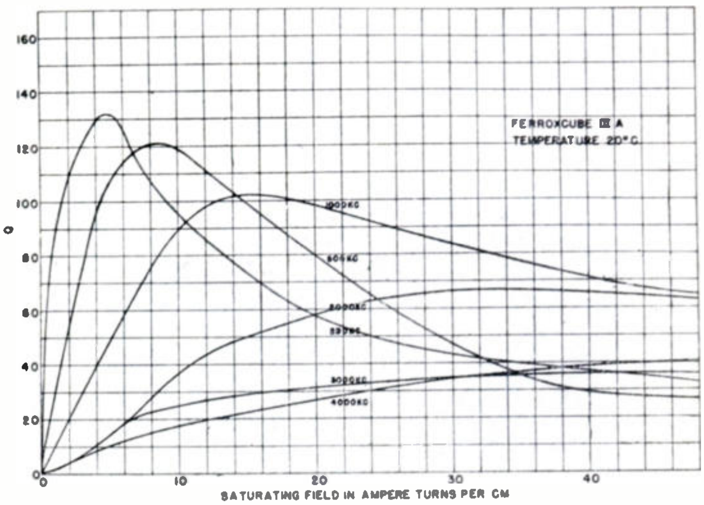

Fig. 4—Q versus saturating field.

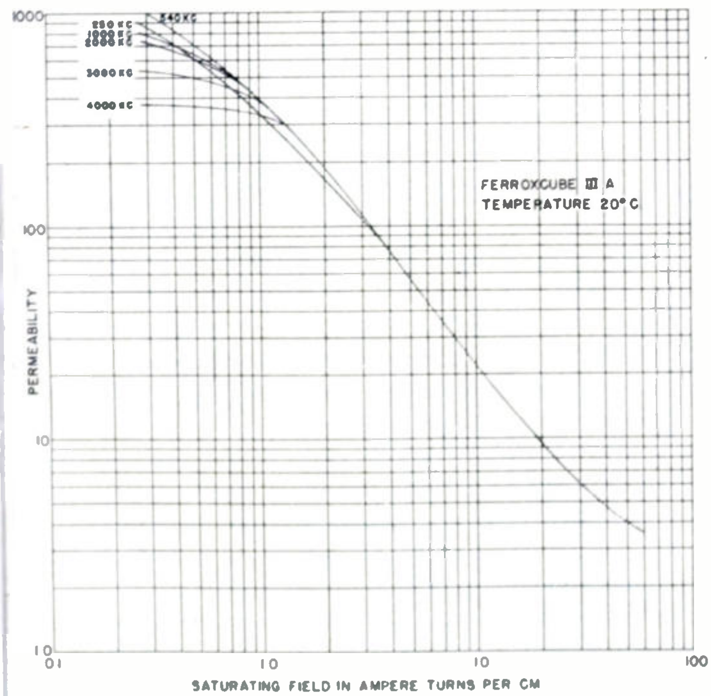

Fig. 3—Permeability versus saturating field.

saturating field is shown in Fig. 3 for various frequencies. With zero saturating field, Q falls off so rapidly with frequency that oscillations could not be maintained much above 700 kc. But Q rises with saturating

Design of Saturable Inductance

The core used in the saturable inductance is a toroid of $1\frac{3}{8}$ inches inside diameter, $1\frac{3}{4}$ inches outside diameter and $\frac{5}{8}$ inch height. It is saturated by a current through an auxiliary winding. It is essential that there be no coupling between the radio-frequency and saturating windings. Such coupling would result in radio-frequency currents in the saturating winding which would reflect impedance back into the radio-frequency winding and would make the saturating winding, its distributed capacity, and current source effectively part of the radio-frequency circuit and lead to generally unpredictable results.

The scheme we have adopted for decoupling the radio-frequency and saturation windings is shown in Figs. 5 and 6. A rectangular hole 8×16 mm is cut into the side of the toroid, the plane of the hole being normal to the radius of the toroid. Two 24-turn radio-frequency windings are threaded through the hole and connected as shown in Fig. 5. The direction of winding is such that the radio-frequency flux returns around the hole and cancels out in the back leg of the toroid. The toroid is now completely enclosed in a doughnut-shaped metallic jacket which leaves about a quarter-inch of clearance between the inside of the jacket and the toroid itself. A saturating winding of 380 turns is threaded through the doughnut-shaped jacket. This winding subtends an angle of about 45° and is centered over the radio-frequency winding. None of the radio-frequency flux lines links the saturation winding and, therefore, there is no magnetic coupling between the two coils. Winding the saturating coil on the outside of the metallic jacket also decreases the electrostatic coupling between the two coils. The jacket serves as a liquid tight container through which oil at a temperature of 35°C is pumped at a speed of ½ gallon per minute from a reservoir whose temperature is maintained constant to 0.01°C. The jacket is made in two sections separated by a neoprene gasket. This prevents the flow of eddy currents which would be induced in the shield by the slowly varying modulation current in the saturation winding.

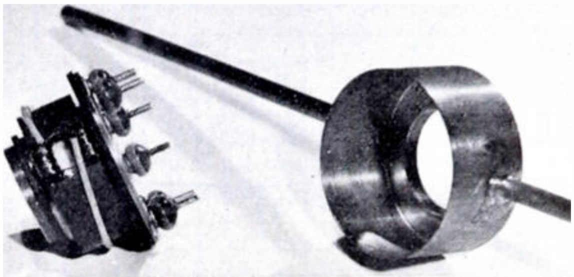

Fig. 5—Saturable inductance disassembled.

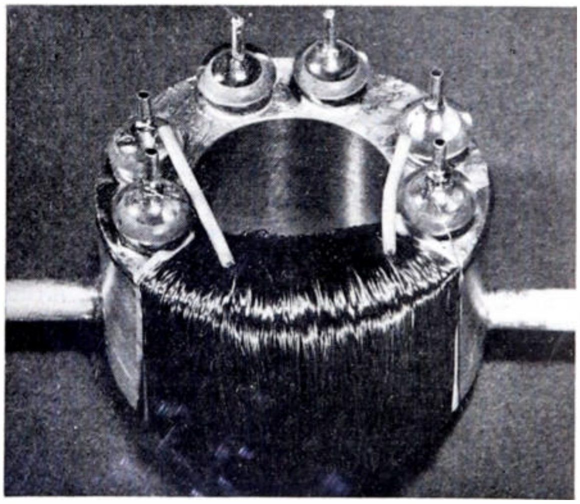

Fig. 6—Saturable inductance assembled.

The saturating magnetomotive force appears for the most part in the legs of the radio-frequency coil since the effective cross section of the magnetic path is smaller there. By altering the thickness of the legs of the radio-frequency coil, the shape of the frequency-versus-saturating current characteristic may be changed. The inductance used in our oscillator had a zero saturating-current inductance of 1,680 μh and dropped to 10.4 μh at 385 ma through the 380-turn saturating winding. This corresponded to a frequency change of 330 to 4,181 kc.

Oscillator Characteristics

The oscillator circuit diagram is shown in Fig. 7 and its frequency-versus-saturating current characteristic in Fig. 8. The oscillating circuit is of the Clapp[3] type with

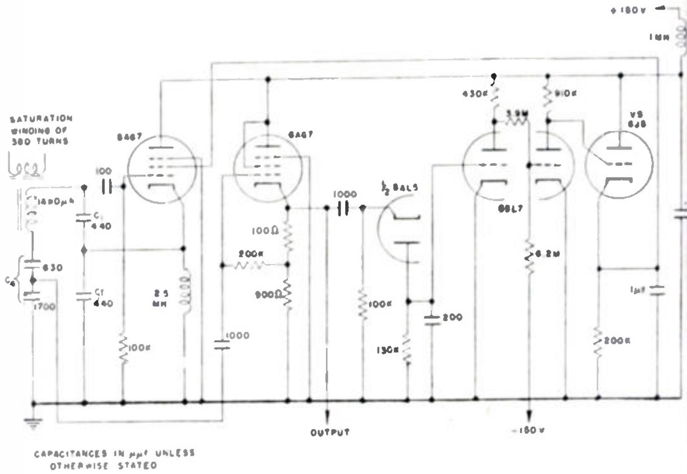

Fig. 7—Permeability tuned oscillator.

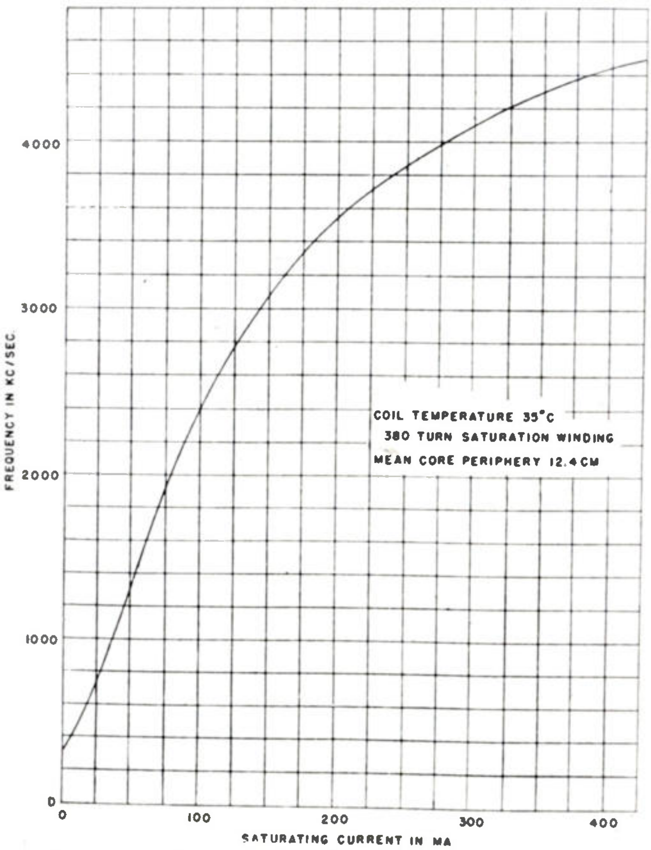

Fig. 8—Oscillator frequency versus saturating current.

[3] J. K. Clapp, "An inductance capacitance oscillator of unusual frequency stability," Proc. I.R.E., vol. 36, p. 356–358; March, 1948.

an amplified automatic-volume-control signal fed back to the screen of the oscillator tube. Without automatic volume control, the signal amplitude falls off by a factor of 6 from 330 to 4,181 kc because of the very large change in L/C ratio throughout the frequency range. With automatic volume control, the amplitude is constant to within 4 per cent. In the Clapp circuit, isolation from the effects of tube interlectrode capacity changes is achieved by making C_1 large compared to C_{gk} and C_{gp} and by making the radio of C_1 to C_e as large as possible. But a large C_1/C_e ratio causes a large relative signal-amplitude ratio at the two ends of the band. An appreciably larger ratio than the one used would kill the oscillations at the high-frequency end. The output has an amplitude of 2.8 volts rms and a sinusoidal wave form which shows no distortion throughout its entire frequency range when viewed on a high-frequency oscilloscope.

The oscillator easily meets its stability and repeatability requirements. At zero saturating field, at about 330 kc, of ten different tubes used as the oscillator tube, the greatest frequency difference between any two was 0.06 per cent. The stability with respect to plate voltage is 0.001 per cent per volt over the range 140 to 150 volts. The stability with respect to filament voltage is 0.006 per cent per volt over the range 6.0 to 6.6 volts. With the oil in the reservoir maintained at 35°C within 0.01°C and with the oil being pumped through the jacket at ½ gallon per minute, the frequency-versus-saturating current curve repeats to within 0.05 per cent for readings taken every 2 hours over a period of five days. At a fixed current at any point within the frequency range, the oscillator will stay within 0.01 per cent of zero beat with a crystal-controlled source over a two-hour period. A frequency-versus-saturation current curve at 36°C departs a maximum of 0.67 per cent from the one taken at 35°C. In the present application of the oscillator, a dc amplifier whose gain without feedback is 650 drives five paralleled 6BG6 tubes whose plate circuits include the saturation winding. The circuit is shown in Fig. 9. Three fourths of the output voltage across a common cathode load resistor is fed back (negative feedback) into the input of the dc amplifier. A voltage swing of 55 volts at the input grid drives the oscillator through its entire frequency range. The frequency-versus-grid voltage characteristic of the amplifier and oscillator in cascade is also reproducible to 0.05 per cent over a period of a week. Because of the negative feedback proportional to current output, the effective source impedance of the generator driving the saturation winding appears as 390,000 ohms. Thus, even with an initial saturation-winding inductance of 0.3 henries, the time constant is sufficiently small that the lag of output current behind input voltage is insignificant in our application.

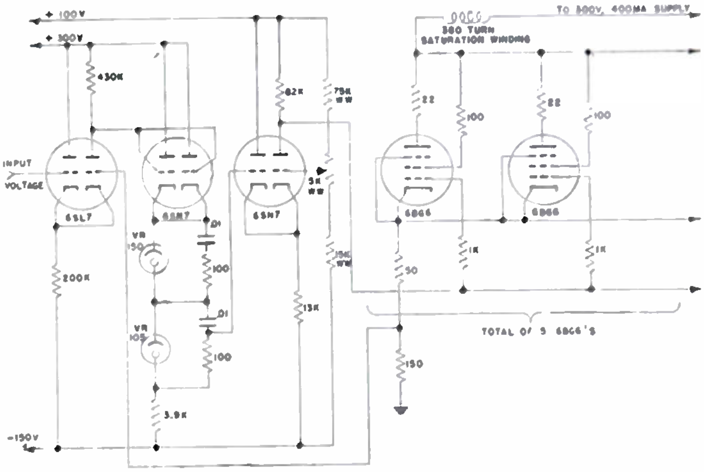

Fig. 9—Control amplifier.

There is a hysteresis effect in the oscillator. At low saturating fields, frequency readings taken at the same current approached from two directions may differ by as much as 6 per cent. Above 1 Mc, the effect is less than 1 per cent. This is of no consequence for the present use of the oscillator, as in each proton acceleration cycle, the saturating current will be moving along the same hysteresis loop in the same direction from the same initial bias.

Saturable inductors, as described above, have also been used in phase shift and Wien Bridge inductance-resistance oscillators. In such oscillators, frequency is proportional to $1/L$ rather than $1/\sqrt{L}$, and much greater frequency ratios for given inductance ratios are obtainable. A Wien Bridge LR oscillator built at this laboratory by W. H. Surber of Princeton University tunes over a 95-to-1 frequency range with a maximum saturating field of about 50 ampere turns per cm. The frequency range covered is 100 kc to 9.5 Mc.

Acknowledgment

The authors are indebted to D. Griffin for his invaluable assistance in the construction and testing of the oscillator and for many helpful suggestions.

Sinusoidal Variation of Inductance in a Linear Series *RLC* Circuit*

EDWARD I. HAWTHORNE†

Summary—Sinusoidal variation of inductance in a series *RLC* circuit is considered in a manner which generalizes previous treatments by placing the variation inside the differentiation sign, and by inclusion of dissipation. A simple relationship is established with solutions for capacitance variation. Dissipation is shown to lead to time-variable damping in a specified manner. Typical differential analyzer solutions are presented.

I. Introduction

The BROAD field of parametric excitation has been receiving considerable attention in recent years due to its wide applicability to many phases of power, communication, and electronic engineering.[1–19] The greatest effort has been directed to the study of simple linear circuits subjected to a sinusoidal time variation of one of the parameters about a mean value. The above represents the simplest case of parametric excitation and also the widest range of its application. Further restrictions on absence of driving forces other than those required for the variation of the parameter lead to the homogeneous differential equation with time-variable coefficients. This implies the study of the oscillatory modes of the circuits and brings the paramount problem of stability to prominence. Once the fundamental solutions of the homogeneous equation are known, response to additional driving forces can be determined by conventional means.[20,21]

Sinusoidal variation of either L or C in a simple series LC circuit, the most basic problem of all, gives rise to a combination of frequency- and amplitude-modulated wave. Various aspects of capacitance variation have been treated rather exhaustively.[1–9,12–15,17–19] Effects of dissipation R have also been considered.[15,17,18] The purpose of this paper is to extend previous treatments to the case of sinusoidally varying *inductance*, including effects of dissipation.

II. Nature of the Problem of Inductance Variation

The problem at hand in the light of previous investigations is summarized in Table I on the basis of a series RLC circuit.[22]

Case 1 has been solved[13,17] by tabulating fairly complete fundamental sets[20] of solutions of Mathieu's equation, to which it leads, using the Eniac built by the Moore School of the University of Pennsylvania, for a wide range of parameters over the interval $0 \leqq \omega\tau \leqq 2\pi$. Application of Floquet's theorem[13,20,21] then enables solutions for $\omega\tau > 2\pi$ to be determined from the tabulated solutions. Regions of stability in the parametric plane have also been determined.[13,15,17] Dissipation was accounted for by multiplying the solutions by the simple damping function, $\epsilon(-R\tau/2L)$, and a suitable modification of the parameters.[17]

Case 2 has been solved[18] by a series expansion method utilizing the techniques of continued fractions for quick and accurate computation of the "characteristic exponent" in the theory of Floquet,[13,15,17,20,21] which also applies to this case, to determine the trigonometric components of the solution.[23] Regions of stability in the

* Decimal classification: R140. Original manuscript received by the Institute, May 2, 1950.

† Moore School of Electrical Engineering of the University of Pennsylvania, Philadelphia, Pa.

[1] John R. Carson, "Notes on the theory of modulation," Proc. I.R.E., vol. 10, pp. 57–64; February, 1922.

[2] B. Van der Pol and M. J. O. Strutt, "On the stability of the solutions of Mathieu's equation," *Phil. Mag.*, vol. 5, pp. 18–38; January, 1928. (Gives numerous references to applications of Mathieu's equation.)

[3] B. Van der Pol, "Frequency modulation," Proc. I.R.E., vol. 18, pp. 1194–1206; July, 1930.

[4] W. L. Barrow, "A new electrical method of frequency analysis and its application to frequency modulation," Proc. I.R.E., vol. 20, pp. 1626–1640; October, 1932.

[5] M. J. O. Strutt, "Lamesche-Mathieusche und Verwandte Funktionen in Physik und Technik," Springer, Berlin; 1932.

[6] W. L. Barrow, "Frequency modulation and the effects of a periodic capacity variation in a nondissipative oscillatory circuit," Proc. I.R.E., vol. 21, pp. 1182–1203; August, 1933.

[7] W. L. Barrow, "On the oscillations of a circuit having a periodically varying capacitance," Proc. I.R.E., vol. 22, pp. 201–213; February, 1934.

[8] H. Ataka, "On superregeneration of an ultra-short-wave receiver," Proc. I.R.E., vol. 23, pp. 841–884; August, 1935.

[9] W. L. Barrow, D. B. Smith, and F. W. Bauman, "Further study of oscillatory circuits having periodically varying parameters," *Jour. Frank. Inst.*, vol. 221, pp. 403–416, March, 1936; and pp. 509–529, April, 1936.

[10] S. B. Crary, "Two-reaction theory of synchronous machines," *Elec. Eng.*, vol. 56, pp. 27–31; January, 1937.

[11] C. F. Wagner, "Unsymmetrical short circuits on water-wheel generators under capacitive loading," *Elec. Eng.*, vol. 56, pp. 1385–1395; November, 1937.

[12] J. G. Brainerd, "Notes on modulation," Proc. I.R.E., vol. 28, pp. 136–139; March, 1940.

[13] J. G. Brainerd and C. N. Weygandt, "Solutions of Mathieu's equation," *Phil. Mag.*, vol. 30, pp. 458–477; December, 1940.

[14] F. J. Maginniss, "Sinusoidal variation of a parameter in a simple series circuit," Proc. I.R.E., vol. 29, pp. 25–28; January, 1941.

[15] J. G. Brainerd, "Stability of oscillations in systems obeying Mathieu's equation," *Jour. Frank. Inst.*, vol. 233, pp. 135–142; February, 1942.

[16] N. Minorsky, "On parametric excitation," *Jour. Frank. Inst.*, vol. 240, pp. 25–46; July, 1945.

[17] H. J. Gray, R. Merwin, and J. G. Brainerd, "Solution of the Mathieu equation," *Trans. AIEE*, vol. 67, pp. 429–441; 1948.

[18] E. Cambi, "Trigonometric components of a frequency-modulated wave," Proc. I.R.E., vol. 36, pp. 42–50; January, 1948.

[19] E. M. Williams and L. Vallese, "Wide deviation frequency-modulated oscillators," Proc. I.R.E., vol. 36, pp. 1282–1284; October, 1948.

[20] E. T. Whittaker and G. N. Watson, "Modern Analysis," Cambridge University Press; 1920.

[21] E. L. Ince, "Ordinary Differential Equations," Dover Publications; 1944.

[22] The principle of duality permits the extension of this to an "equivalent" parallel G-C-L circuit.

[23] This method could, of course, be also applied to Case 1.

TABLE I—THREE CASES OF SINUSOIDAL VARIATION OF L OR $C(K<1)$

	Variable element	Differential equation to be solved	Effect of dissipation R	Most recently analyzed by:
Case 1	$\frac{1}{C}=\frac{(1+K\cos\omega\tau)}{C_0}$	Mathieu's Equation	simple damping	J. G. Brainerd[13,16,17]
Case 2	$C=C_0(1+K\cos\omega\tau)$	A form of Hill's Equation	simple damping	E. Cambi[18]
Case 3*	$L=L_0(1+K\cos\omega\tau)$	More complex form of Hill's Equation	variable damping	this paper

* Inverse L variation is as yet of little practical significance.

ırametric plane were also determined.[18] Dissipation ·as again accounted for by the factor $\epsilon(-R\tau/2L)$ and a ıitable modification of the "characteristic exponent."[18]

Case 3 represents a more complicated situation. If ıe variation could be taken outside of the differentia-on sign, then the equation

$$(1+K\cos\omega\tau)\frac{d^2\hat{q}}{d\tau^2}+R\frac{d\hat{q}}{d\tau}+\frac{\hat{q}}{C}=0;\quad K<1 \qquad (1)$$

ould result, which is exactly equivalent to Case 2 if .ssipation be zero in both cases. This was actually con-.dered by Cambi[18] as the case of variable "inductance." ıductance variation, however, requires that total flux nkages be differentiated in summing up emf's around ıe circuit, and gives rise to the basic equation for ase 3:

$$\frac{d}{d\tau}\left[(1+K\cos\omega\tau)\frac{d\hat{q}}{d\tau}\right]+R\frac{d\hat{q}}{d\tau}+\frac{\hat{q}}{C}=0; K<1. \qquad (2)$$

he solution of equation (2) will now be discussed.

III. SEPARATION OF THE EQUATION

The substitution:

$$t=\omega\tau;\qquad Q=\frac{\omega L_0}{R};\qquad \epsilon=\frac{1}{\omega^2L_0C} \qquad (3)$$

ːduces (2) to the nondimensional form

$$\frac{d}{dt}\left[(1+K\cos t)\frac{d\hat{q}}{dt}\right]+\frac{1}{Q}\frac{d\hat{q}}{dt}+\epsilon\hat{q}=0. \qquad (4)$$

:limination of the first derivative term can be obtained y the transformation

$$(t,Q,K,\epsilon)=Y(t,Q,K,\epsilon)\cdot W(t,K)\cdot E(t,K,Q), \qquad (5)$$

ʼhere Y is a solution of

$$\ddot{Y}+Y\left[\frac{1}{4}+\frac{\epsilon}{1+K\cos t}-\frac{(1-K^2)+\frac{1}{Q^2}}{4(1+K\cos t)^2}\right]=0, \qquad (6)^{24}$$

nd where

$$W=\sqrt{\frac{1+K}{1+K\cos t}} \qquad (7)$$

[24] Dots denote differentiation with respect to t.

$$E=\exp\left[\frac{1}{Q}\cdot\frac{\tan^{-1}\left(\sqrt{\frac{1-K}{1+K}}\tan\frac{t}{2}\right)}{\sqrt{1-K^2}}\right]. \qquad (8)$$

IV. EFFECT OF DISSIPATION

The predominant effect of dissipation is included in the damping function E, equation (8). The factor $1/Q^2$ in (6) has a small, second-order effect[25] for Q larger than about 10. Fig. 1 indicates when the $1/Q^2$ term can surely be neglected for 1 per cent accuracy.

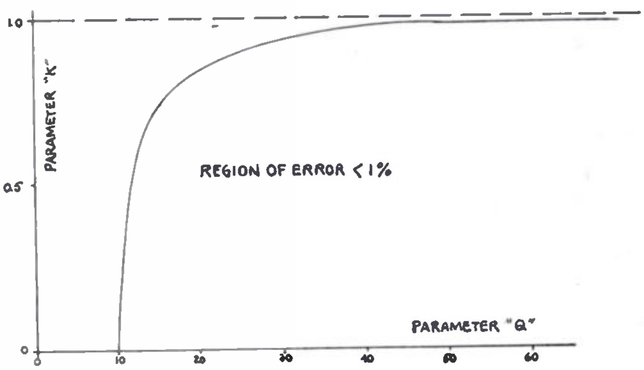

Fig. 1—Effect of neglecting the $1/Q^2$ term in (6).

The sinusoidal variation of inductance has an interesting effect on the damping function (8), which in absence of variation ($K=0$) properly reduces to the $\epsilon-(t/2Q)=\epsilon-(R\tau/2L_0)$ expression. Fig. 2 indicates how the numerator in the exponent of E varies with time, depending on K, in a manner which may be termed "damping modulation."

V. SOLUTION OF THE DIFFERENTIAL EQUATION

If the factor $1/Q^2$ be neglected in (6) then $Y\cdot W$ becomes independent of Q and may be shown to be the solution of the original equation (2) without dissipation, which then becomes

[25] An analogy exists in the case of wave propagation where the primary effect of dissipation lies in the damping of the amplitude function, and where dissipation has only a second-order effect on the phase of the wave. It should also be pointed out that neglecting the $1/Q^2$ term in (6) leads to pessimistic, hence safe, results with respect to stability consideration, since dissipation cannot make an otherwise stable system unstable.

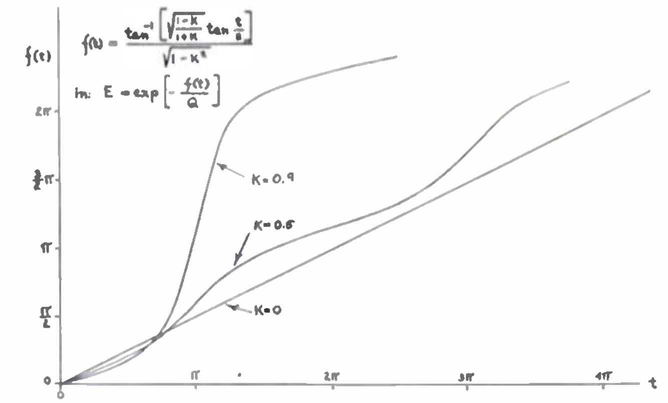

Fig. 2—Variation of the exponent $f(t)$, of damping function, $E(t)$, with t.

$$\frac{d}{dt}\left[(1 + K\cos t)\frac{dq}{dt}\right] + \epsilon q = 0. \qquad (9)$$

The complete approximate solution of (2) then is

$$\hat{q}(t, K, \epsilon\, Q) = q(t, K, \epsilon)\cdot E(t, K, Q). \qquad (10)$$

Equation (9) represents a form of Hill's equation,[16,18–21] as evident from (6), and its complete solution may be sought by the methods listed under Cases 1 and 2.

A very close correlation, however, exists between (9) which represents Case 3 without dissipation and the differential equation of Case 2 without dissipation, treated by Cambi.[18] In the terminology of this paper the latter becomes:

$$(1 + K\cos t)\frac{d^2r}{dt^2} + \epsilon r = 0. \qquad (11)$$

This relationship can be stated as follows: Let the fundamental set[13,21,22] of (9) be $q_1(t)$ and $q_2(t)$ defined

$$q_1(0) = 1; \quad \dot{q}_1(0) = 0; \quad q_2(0) = 0; \quad \dot{q}_2(0) = 1 \qquad (12)$$

and let the fundamental set of (11) be, correspondingly, $r_1(t)$ and $r_2(t)$, where

$$r_1(0) = 1;\ \dot{r}_1(0) = 0;\ r_2(0) = 0;\ \dot{r}_2(0) = 1. \qquad (13)$$

It can then be shown that

$$\left.\begin{aligned} q_1(t) &= \dot{r}_2(t); & \dot{q}_1(t) &= -\frac{\epsilon}{1+K\cos t}\, r_2(t) \\ q_2(t) &= -\frac{1+K}{\epsilon}\,\dot{r}_1(t); & \dot{q}_2(t) &= \left(\frac{1+K}{1+K\cos t}\right) r_1(t) \end{aligned}\right\}. \qquad (14)$$

Since the solutions q's and r's represent the respective charges on the capacitors, while their derivatives represent the currents in the circuit (for zero dissipation), the above equations indicate that currents in the capacitance-varying circuits correspond to the capacitor charges in the inductance varying circuit, and conversely, except for constants or for simple time functions. With this important modification then, summarized in (14), the work of Cambi[18] may be applied to the inductance variation of Case 3.

VI. Stability Considerations and Typical Solutions

It is evident that stable regions in the parametric ϵ, K plane are coincident for solutions of (9) and (11). Consequently results of Cambi[18] may be generalized to

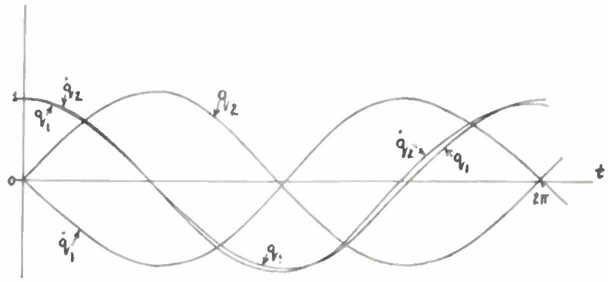

Fig. 3—Differential analyzer solutions of (9) for $K=0.1$, $\epsilon=1.0$.

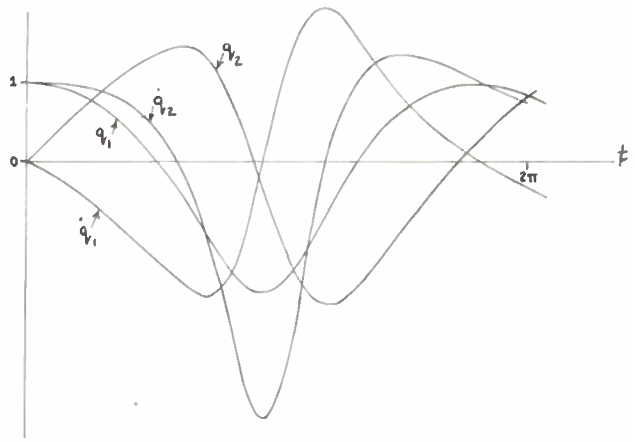

Fig. 4—Differential analyzer solutions of (9) for $K=0.7$, $\epsilon=1.0$.

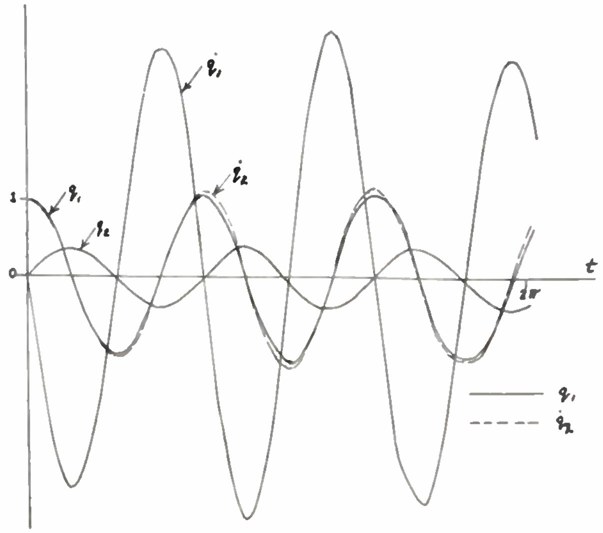

Fig. 5—Differential analyzer solutions of (9) for $K=0.1$, $\epsilon=8.0$.

:lude Case 3 as well as Case 2. Studies of (9), made c the Moore School Differential Analyzer at the niversity of Pennsylvania, have served to substan- ute the above development. Typical fundamental ts of solutions of (9) for a representative range of rameters ϵ and K are shown on Figs. 3 to 6.

VII. Conclusions

Treatment of inductance variation inside the differ- tiation sign, and including dissipation, gives rise to vo distinct effects. One is the direct relationship with e previously studied[18] solutions of the capacitance var- tion in the sense that the functional expressions for pacitor charges and circuit currents are, roughly eaking, interchanged. The other, caused by dissipa- on, is the time varying damping, illustrated on Fig. 2.

VIII. Acknowledgments

The author wishes to express his gratitude to J. G. rainerd for his help in preparation of this paper. The uthor is indebted to the Moore School of Electrical ngineering of the University of Pennsylvania for the se of the differential analyzer.

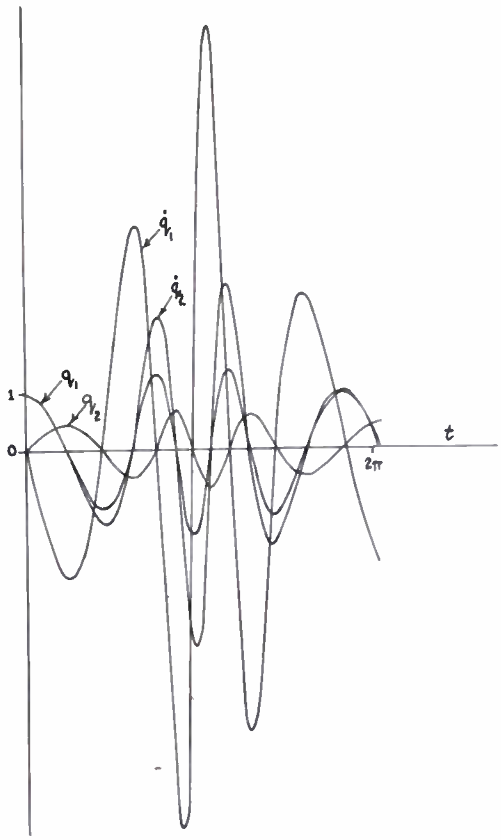

Fig. 6—Differential analyzer solutions of (9) for $K=0.7$, $\epsilon=8.0$

Some Studies of Pulse Transformer Equivalent Circuits*

C. K. HADLOCK†, ASSOCIATE, IRE, AND D. LEBELL‡, ASSOCIATE, IRE

Summary—**A pulse transformer equivalent circuit is synthesized om data obtained by direct test on a typical pulse transformer. he differential equations for the equivalent circuit are solved by eans of a mechanical differential analyzer for a variety of circuit onstant values. Wave forms of transformer output voltage as ob- ined from the differential analyzer are of assistance in the design pulse transformers.**

I. Introduction

PULSE TRANSFORMERS are currently finding important application in computer and radar cir- cuits, as well as in many other devices utilizing hort duration pulses. The method of analysis to be escribed here is based on an equivalent circuit which as been idealized to the extent that the pulse trans- ormer which contains nonlinear, distributed parameters s represented by a circuit containing linear, lumped ircuit constants. These assumptions permit a rela- ively simple procedure for synthesis and subsequent nalysis of the equivalent circuit.

* Decimal classification: 143.5. Original manuscript received by he Institute, December 1, 1949; revised manuscript received, June 2, 1950.
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‡ The University of California, Los Angeles, Calif.

II. Synthesis of Equivalent Circuit[1]

The particular pulse transformer to be studied is treated as a four-terminal network whose basic arrange- ment of impedances (Z_A, Z_B, Z_C) is the T of Fig. 1. The assumption of the T configuration fixes the frequency responses of Z_A, Z_B, and Z_C for a particular transformer. These responses are obtained by measurement of the

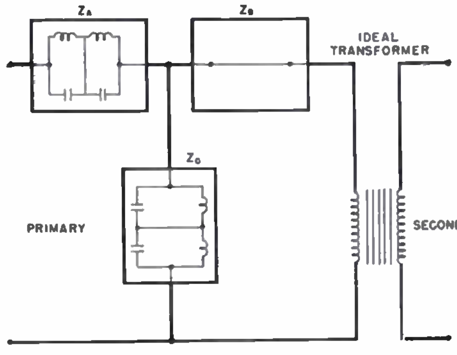


Fig. 1—Four-terminal network representing pulse transformer ($Z_B=0$).

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

primary and then the secondary frequency characteristics for both open- and short-circuit condition on the transformer.

The test data obtained from the transformer (Fig. 2) indicate that Z_B is very nearly zero. Therefore, for equivalence, Z_A should display the same response as that seen from the primary winding with the secondary

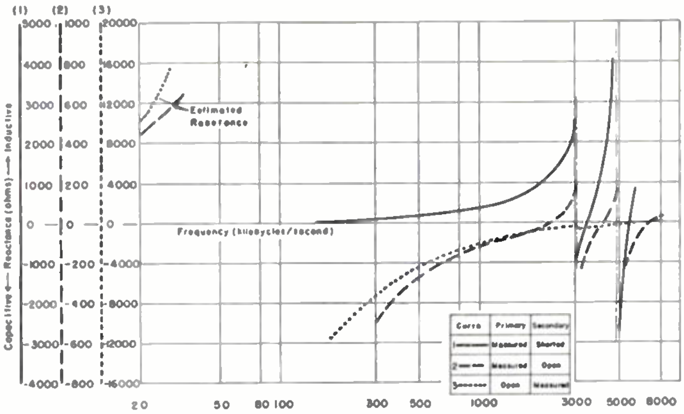


Fig. 2—Frequency characteristics of a typical pulse transformer (test data).

short circuited (i.e., the two anti-resonant and two series resonant points of curve 1). The network representing Z_A in Fig. 1 is capable of producing this response. Correspondingly, the network representing Z_C in Fig. 1 is obtained from consideration of Fig. 2, curve 3 which is the transformer's response as seen from the secondary with the primary open. Hence the circuit of Fig. 1 will exhibit the frequency response characteristics of the actual transformer.

The approximation is made that the circuits representing Z_A and Z_C could be replaced by single LC parallel combinations. The validity of this step is questionable where frequencies above 4.0 Mc appear (see Fig. 2). The Fourier frequency spectrum of a single pulse exhibits an appreciable amplitude over a frequency interval extending from $-(10/2\epsilon)$ to $+(10/2\epsilon)$ where ϵ is the pulse duration. For a 1-μ sec pulse duration then, $10/2\epsilon = 5$ Mc. Hence neglecting all components above 4.0 Mc might introduce serious error. Assuming, however, that this simplification is permissible, the equivalent L (T with the impedance $Z_B=0$) is shown in Fig. 3. An equivalent pulse generator and load (referred to the primary) has been added. Capacitor C_2 is assumed to contain any load capacitance; values of generator and load resistance (R_1 and R_2) may be modified to include transformer copper loss and core loss, respectively.

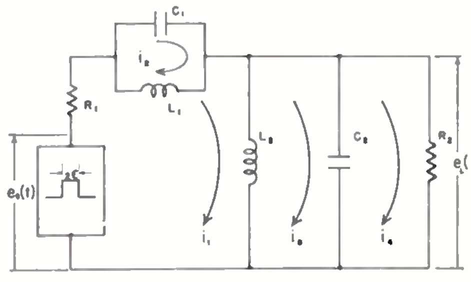


Fig. 3—Assumed equivalent circuit of pulse transformer.

III. Analysis of Equivalent Circuit[2]

Analysis of this equivalent circuit is based on its conventional loop equations which follow:

$$e_{in}(t) = R_1 i_1 + L_1 \frac{d}{dt}(i_1 - i_2) + L_2 \frac{d}{dt}(i_1 - i_3) \quad (1)$$

$$0 = L_1 \frac{d}{dt}(i_2 - i_1) + \frac{1}{C_1}\int i_2 dt \quad (2)$$

$$0 = L_2 \frac{d}{dt}(i_3 - i_1) + \frac{1}{C_2}\int (i_3 - i_4)dt \quad (3)$$

$$0 = R_2 i_4 + \frac{1}{C_2}\int (i_4 - i_3)dt \quad (4)$$

where $e_{in}(t)$ is a voltage pulse of magnitude E_0 and duration ϵ.

In order to explore the effects on the output wave shape of changing the circuit parameter values, a number of solutions corresponding to several sets of values of circuit constants were desired. Solutions obtained by techniques of operational calculus would have required an excessive expenditure of man hours of computing. For this reason the problem was solved by means of the University of California mechanical differential analyzer. This computer was selected because of its ability to produce many solutions rapidly, inexpensively, and with the desired accuracy. Equations (1) to (4) were solved in dimensionless form in order to facilitate the generalization of results. A selection of representative solutions (output curves) obtained from the differential analyzer appears in Figs. 4 and 5.

IV. Conclusions

The various effects on the output wave form for changes in the circuit parameters (see Figs. 4 and 5) are of special interest in determining design criteria. Existing discrepancies between the output wave shape of the transformer tested, and the wave shape obtained from analysis of the equivalent circuit are directly traceable to the simplifying assumptions made in arriving at the equivalent circuit of Fig. 3. It is difficult to obtain simple, accurate, equivalent circuits of devices which contain distributed parameters when consideration of a wide frequency range is necessary. In particular, the representation of Z_A and Z_C by single LC combinations is valid only if frequencies above the first resonant point are negligible in the circuit. Further, the assumed linearity of all circuit elements may produce discrepancies under certain conditions.

It is over these considerations that a compromise must always be made between ease of analysis and accuracy of results to determine the most suitable choice

[2] "The Differential Analyzer of the University of California," University Press, University of California, Los Angeles, Calif., 1947.

equivalent circuit. Greater precision is obtainable y selecting more complicated impedance structures for $_A$, Z_B, and Z_C and/or by treating certain elements (pri- arily L_2) as nonlinear. However, the output wave apes of the actual transformer and those obtained om the equivalent circuit of Fig. 3 (employing meas- red values of circuit constants) show sufficiently good greement for most practical purposes.

The desirability of employing a computer is immedi- tely evident from a comparison of the time required to lve the circuit equations by means of the Fourier trans- rm and differential analyzer methods.

Twenty-five hours were required to obtain a solution r a single set of parameters by the Fourier transform ethod whereas the average time required by the dif- erential analyzer was less than twenty minutes per olution. More complicated equations which would of necessity result from a more complicated equivalent circuit and the introduction of nonlinear elements would greatly increase the advantage of a computer solution over other methods.

Acknowledgments

Acknowledgments for the contents of this paper are due the Engineering Pulse Transformer Research Project, sponsored by the Air Matériel Command under the direction of L. M. K. Boelter. Special acknowledgments are due J. W. Miles and L. L. Grandi for their contributions to this research project.

Bibliography

1. G. N. Glascoe and J. V. Lebacqz, "Pulse Generators," Radiation Laboratory Series #5, Part III, chap. 12, McGraw-Hill Book Co., New York, N. Y.; 1948.
2. Reports I, II, and III, Pulse Transformer Contract w-28-099-ac-195, University of California, Los Angeles, Calif.

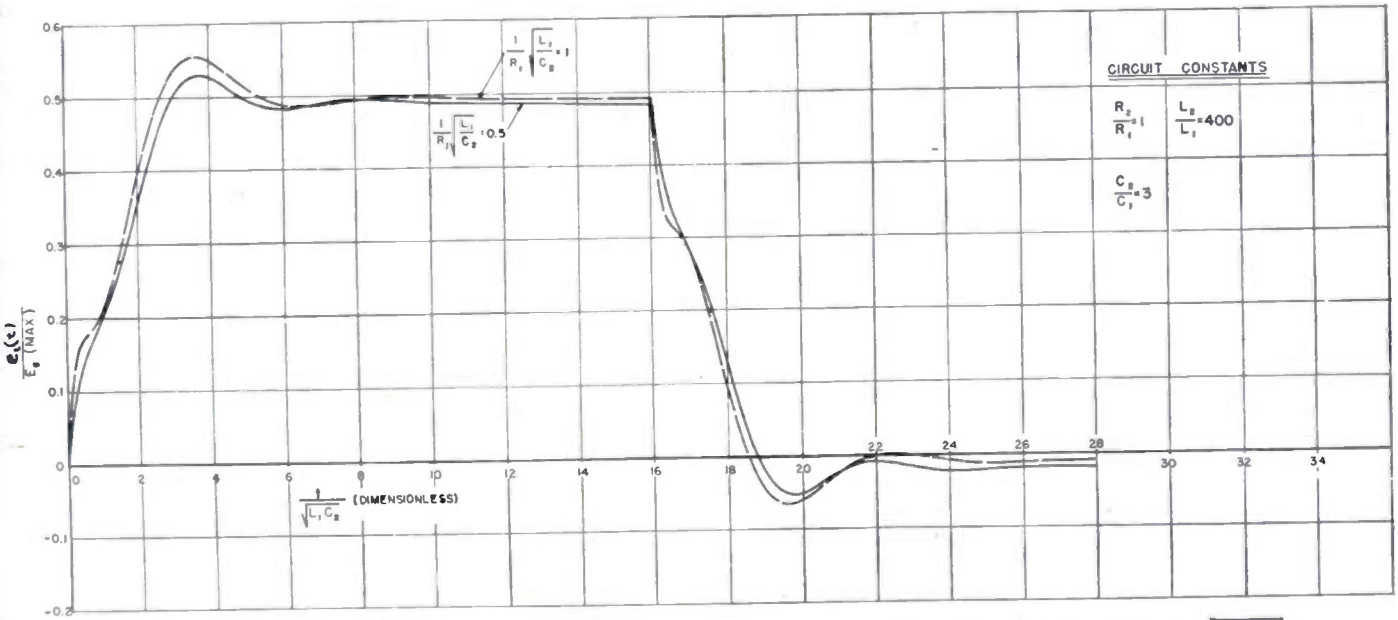


Fig. 4—Typical output curves from mechanical differential analyzer, showing the effect of a varying $1/R_1\sqrt{(L_1/C_2)}$.

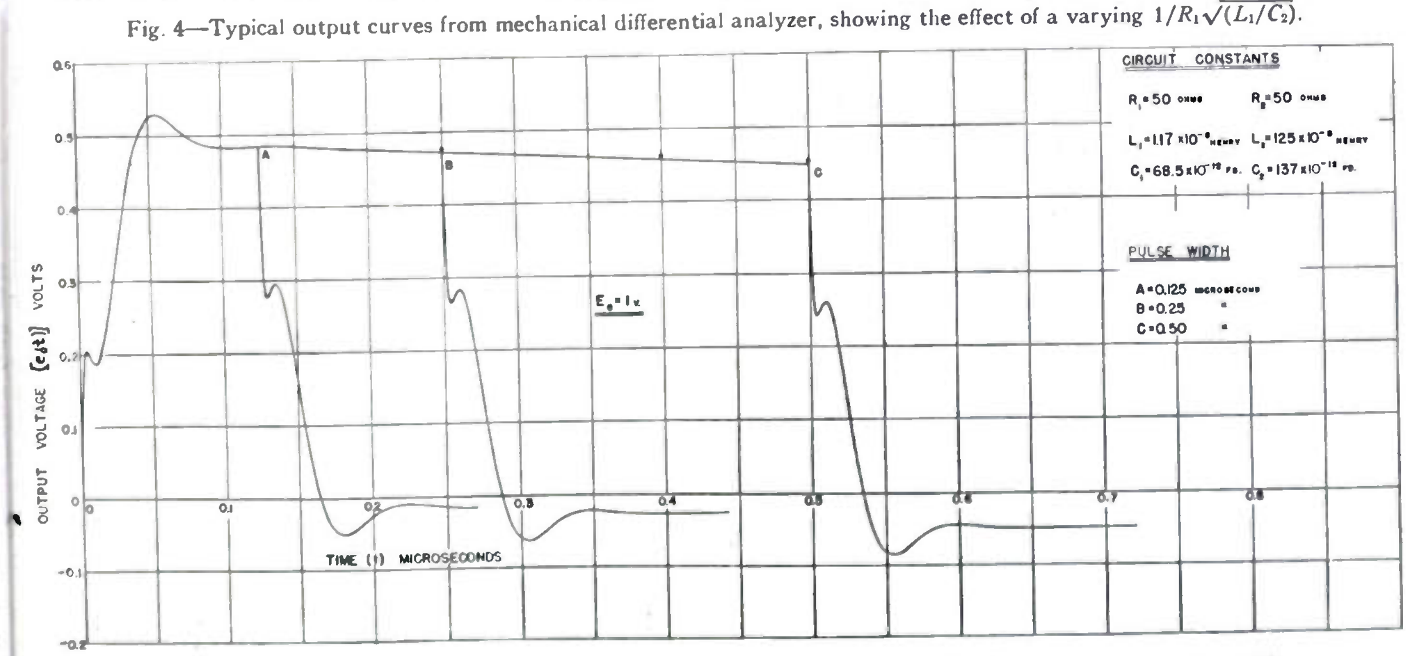


Fig. 5—Typical output curves from mechanical differential analyzer, showing the effect of a varying pulse width.

Design of Exponential-Line Pulse Transformers*

EVERARD M. WILLIAMS†, SENIOR MEMBER, IRE, AND EDWARD R. SCHATZ†, ASSOCIATE, IRE

***Summary*—Design procedures for exponential transmission-line-section pulse transformers are described together with data on an experimental unit.**

THE GENERAL THEORY of pulse transformers comprising sections of exponential transmission line has been described previously.[1] This paper summarizes design procedures for such transformers and describes an experimental unit.

An exponential-line pulse transformer is constructed from a section of transmission line in which conductor spacing is tapered so that the inductance and capacitance per unit, l_x and c_x, vary exponentially with the length x according to the relations

$$l_x = L_0\epsilon^{\gamma x}$$

$$c_x = C_0\epsilon^{-\gamma x}$$

where L_0 and C_0 are the inductance and capacitance per unit length at $x=0$, the input.

The impedance level at the input end is

$$Z_{\text{in}} = \sqrt{\frac{L_0}{C_0}}$$

while that at the output end is

$$Z_{\text{out}} = \sqrt{\frac{L_0}{C_0}}\,\epsilon^{\gamma x}.$$

It seems probable from the results of the earlier study that maximum nominal efficiency (in the case of the very short pulses for which the exponential-line pulse transformer is particularly suitable) results from a selection of output impedance level equal to load resistance since this condition brings about a minimum initial reflection at the load.

The insertion of filter sections, somewhat as outlined by Wheeler[2] for the steady-state exponential-line transformer, could be used to reduce or even substantially eliminate reflections; this alternative, however substitutes energy losses in terminating resistors and energy stored in reactive elements for energy otherwise reflected. Compensating filters could undoubtedly be used advantageously for shaping output pulses in cases in which efficiency is not of primary importance.

Either coaxial or open-wire lines may be employed. Christiansen[3] has developed for steady-state use a particularly convenient form of open-wire exponential line which may be adapted for pulse-transformer purposes. The authors have found it more convenient to work with coaxial-line structures, because of the ease with which powder or liquid dielectrics may be employed to increase electrical length and the adaptability of this form of line to extremely high power levels. These coaxial-line structures employ cylindrical outer conductors and tapered inner conductors. The shape required in the tapered conductor is given by the expression for the ratio of radii (in terms of sending end ratio of radii R_2/R_1) as

$$\frac{r_2}{r_1} = \left(\frac{R_2}{R_1}\right)^{\epsilon^{\gamma x}}. \tag{1}$$

A sketch of typical curves of this function, which imposes fairly evident limitations on realizable impedance levels and ratios, appears in Fig. 1. In this figure K is the dielectric constant.

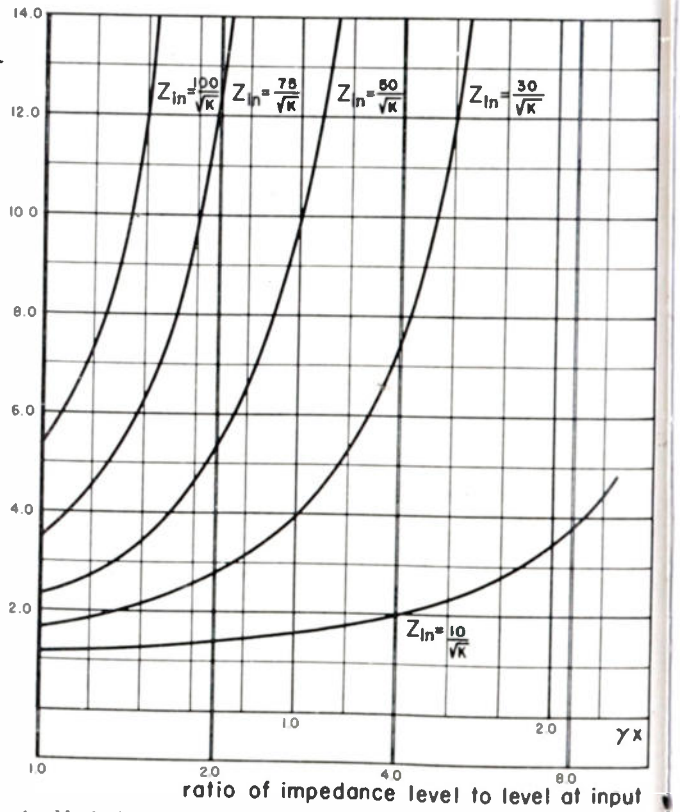


Fig. 1—Variation of ratio of conductor radii with length parameter γx in coaxial exponential-line pulse transformers as a function of input impedance level.

* Decimal classification: R143.5. Original manuscript received by the Institute, May 22, 1950; revised manuscript received, August 14, 1950. Presented as part of the paper, "The Exponential-Line Pulse Transformer," 1950 IRE National Convention, New York, N. Y., March 9, 1950.

This work was supported in part by the joint ONR-AEC program.

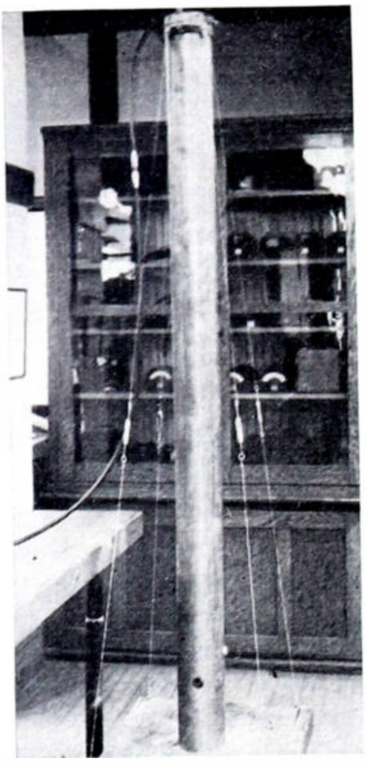
† Carnegie Institute of Technology, Pittsburgh 13, Pa.

[1] E. R. Schatz and E. M. Williams, "Pulse transients in exponential transmission lines," PROC. I.R.E., vol. 38, pp. 1208–1212; October, 1950.

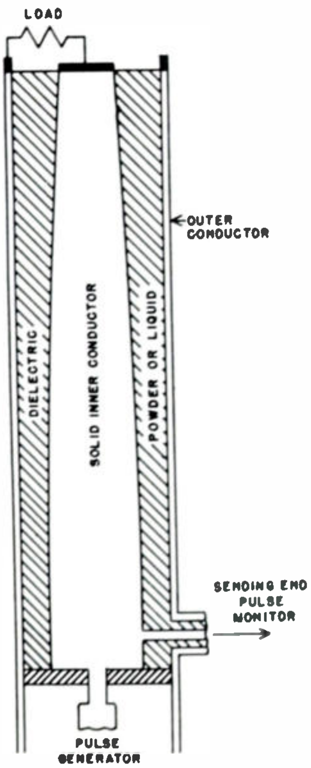
[2] Harold A. Wheeler, "Transmission lines with exponential taper," PROC. I.R.E., vol. 27, pp. 65–71; January, 1939.

[3] Wilbur N. Christiansen, "An exponential transmission line employing straight conductors," PROC. I.R.E., vol. 35, pp. 576–581; June, 1947.

Fig. 2 shows a photograph and sketch of an experimental pulse transformer. The center conductor was machined from a solid rod by a series of linear taper cuts which approximate the radii given by (1) to about 0.0005 inch. No data are available at present as to the accuracy with which it is necessary to approximate exponential tapers; however, tests with this experimental unit show no distortion attributable to discontinuities in its taper.



(a)



(b)

Fig. 2—(a) Photograph and (b) sketch of experimental transformer. 1.5 meters in length. The pulse generator is incorporated in the base of the unit to simplify lead problems with the short pulses employed.

The conductor radii are determined by the load impedance, transformation ratio, and dielectric used. The outer cylinder radius may be chosen sufficiently large to reduce attenuation and avoid pulse distortion due to this cause; in many instances, however, the principal factors in attenuation will be those due to the line dielectric and the outer conductor radius will be determined by mechanical considerations relating to size, weight, support, and ease of construction. The most significant electrical design choices are:

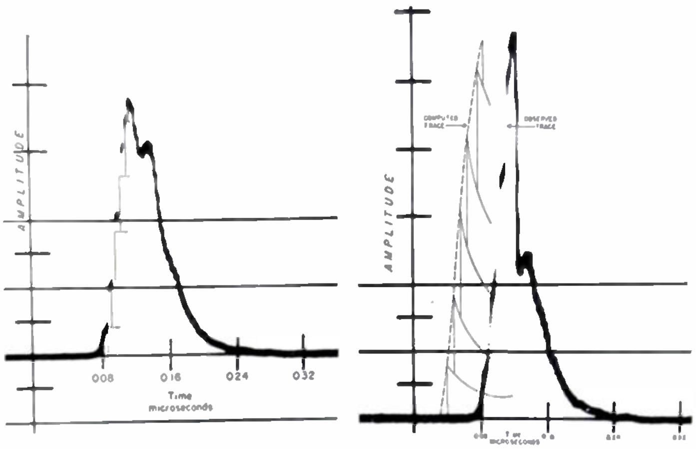
(1) the length of the line section and its associated taper, and

(2) the dielectric material.

These will be considered in some detail.

I. Choice of Line Section Length

With a given impedance ratio the pulse distortion and transformer efficiency depend on the length of the line section or, more precisely, on the ratio of line length x to velocity of propagation c in the dielectric between conductors. Curves given in the earlier paper[1] provide a ready means for computing response to a rectangular input pulse and thus for selection of x/c to



(a) (b)

Fig. 3—Typical input and output pulses for the transformer of Fig. 2, showing step-by-step method for calculation of response to pulses with finite rise time. Computed output is intentionally displaced to left to avoid confusion. (a) Input pulse. (b) Pulse delivered to load.

provide an output pulse with a specified minimum distortion. When the input pulse is not rectangular (as, for instance, in the common case of finite rise and decay times) it may be necessary to take the actual input pulse shape into account. The transient response of an exponential-line transformer to transients of finite rise time may be computed from the step-function response of the earlier paper by means of a step-by-step analysis. In this analysis the input pulse is resolved into the sum

TABLE I

PROPERTIES OF SOME DIELECTRICS SUGGESTED FOR USE IN EXPONENTIAL-LINE PULSE TRANSFORMERS

Material	Approximate Dielectric Constant	Approximate Power Factor in Frequency Range of Interest	Remarks
Barium-Strontium-Titanate (solid)	7,500	1–2 per cent	Losses too high except for very short lines, poor dielectric strength.
*Barium-Strontium-Titanate (sintered-powdered)	20	0.05–0.1 per cent	High leakage due to surface effects, poor dielectric strength, but easy to handle.
Rutile (TiO_2) (solid)	99	.01–0.07 per cent	Quite satisfactory.
*Nitrobenzene, C. P. $C_6H_5NO_2$ (liquid)	36	0.25 per cent	Satisfactory where losses can be tolerated.
*Mixture of nitrobenzene and powdered barium-strontium-titanate (liquid)	50	0.35 per cent	Satisfactory where losses can be tolerated.

* Data based on measurements of commercial samples.

of a number of step-impulse functions, delayed one with respect to another to form an approximation to the input pulse. The transformer output response is determined from the superposition of its responses to each of the input steps. An analysis of this type is illustrated in Fig. 3 for the transformer of Fig. 2. It is evident that quite precise results follow in this instance from a relatively small number of steps.

A similar process may be applied to the determination of input current and the calculation of efficiency.

II. CHOICE OF DIELECTRIC MATERIAL

The length and, therefore, the physical size of an exponential-line pulse transformer may be reduced in proportion to the square root of the dielectric constant k of the dielectric material, providing that material with sufficiently low losses is used. The effect of losses on output wave form may be estimated from the attenuation of high-frequency components according to the expression, derived from the general attenuation expression of the earlier paper

$$\frac{e_{out}}{e_{in}} = \epsilon^{-\alpha x} = \epsilon^{-(\pi f\delta/c)x}$$

in which δ and c are the power factor and propagation velocity, respectively, in the dielectric medium. For a specified transformer ratio Z^{out}/Z^{in} and nominal efficiency N, with which the line-section length with air dielectric is x_0, the actual length is

$$\frac{x_0}{\sqrt{k}}$$

and the velocity $c = C_{air}/\sqrt{k}$ so that the attenuation factor αx is

$$\frac{\pi f \delta x_0}{C_{air}}$$

and the attenuation is directly proportional to dielectric power factor, regardless of the dielectric constant and the consequent length of the transformer section.

Some typical dielectrics that have been used or considered for use by the authors are tabulated in Table I. Construction of lines is simplified when liquid or powder dielectrics are used since the lines may be machined and assembled and the dielectric poured into place. Use of a material such as rutile would probably require firing of large tapered dielectric sections on which conductors would be sprayed or plated.

The experimental transformer of Fig. 2 was designed for use with either liquid (nitrobenzene) or powder (powdered barium-strontium-titanate) dielectric. It has a nominal impedance ratio of one to four and would develop its full voltage ratio of one to two with a pulse of about 3 millimicroseconds duration. With the pulse of Fig. 3 its ratio is about one to one and one-half. Its peak power capacity with powdered barium-strontium-titanate dielectric is about 12.5 megawatts. With nitrobenzene dielectric, a better insulator, its peak power capacity is about 500 megawatts.

CORRECTION

E. A. Gerber, author of the paper, "High-Frequency Vibrations of Plates Made from Isometric and Tetragonal Crystals," which appeared on pages 1073–1079, of the September, 1950, issue of the PROCEEDINGS OF THE I.R.E., has called to the attention of the editors the following error:

On page 1075, the third line of formula (12) should read

$$-\frac{1}{\epsilon_3^S}\frac{\partial\epsilon_3^S}{\partial T}\bigg)\bigg\}\cos^2\phi\bigg]$$

instead of

$$+\frac{1}{e_3^S}\frac{\partial e_3^S}{\partial T}\bigg)\bigg\}\cos^2\phi\bigg].$$

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John P. Blewett (A'43) was born in Toronto, Canada, on April 12, 1910. He received the B.A. and M.A. degrees from the University of Toronto in 1932 and 1933, respectively, and the Ph.D. degree in physics from Princeton University in 1936.

J. P. Blewett

After a year at the Cavendish Laboratory in Cambridge, England, as a Royal Society of Canada Fellow, he joined the staff of the research laboratory of the General Electric Company in Schenectady, N. Y. While at General Electric, he was engaged in studies of oxide-coated cathodes, microwave power generation and propagation, and high-energy electron accelerator design and operation.

In 1947, Dr. Blewett joined the accelerator project at the Brookhaven National Laboratory, Upton, L. I., N. Y. At present he is associated with the three-billion-volt proton synchrotron project at Brookhaven.

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Charles E. Brockner (A'43–SM'50) was born in New York, N. Y., on August 12, 1917. He received the B.S. degree in electrical engineering from Union College in 1940 and was awarded the M.S. degree at Columbia in 1942, also in electrical engineering. While at Columbia, he served as a teaching fellow in electrical engineering and also undertook subminiature tube investigations under sponsorship of the Department of Terrestrial Magnetism, Carnegie Institute of Washington.

C. E. Brockner

From 1942 to 1946 Mr. Brockner was a member of the electrical engineering staff at Union College, attaining the rank of associate professor before leaving. During his final year at that College he served in a consultant capacity with the General Crystal Corporation. Since 1946 Mr. Brockner has worked mainly on automatic fire control systems while with Sperry Gyroscope Company, in Great Neck, L. I., N. Y. He is now a senior project engineer in the armament radar engineering department.

Mr. Brockner is a member of the Society of Sigma Xi and of the American Institute of Electrical Engineers.

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For a photograph and biography of Edward R. Schatz, see page 1219 of the October, 1950, issue of the Proceedings of the I.R.E.

Elery F. Buckley (S'42–A'43–M'50) was born on December 31, 1920, at Bright, Ont., Canada. He received the B.A.Sc. degree in electrical engineering from the University of Toronto in 1943, and the S.M. degree from the Massachusetts Institute of Technology in 1949.

E. F. Buckley

From 1943 to 1946, he was a member of the junior teaching staff in the department of electrical engineering at the University of Toronto. During this period he was also a member of a geophysical survey party searching for radioactive ore deposits in the Canadian northwest, and spent a brief time with the radiology section of the National Research Council of Canada. Since 1946 he has been on the teaching staff of the Massachusetts Institute of Technology. He is now an instructor in electrical engineering, spending a considerable portion of his time in the Research Laboratory of Electronics.

Mr. Buckley is a member of the American Institute of Electrical Engineers and the American Society for Engineering Education.

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John P. Chisholm was born in Poughkeepsie, N. Y., on December 1, 1923. The first three years of his undergraduate study were obtained from St. Francis Xavier University in Antigonish, Nova Scotia. He receive the B.S. degree in electrical engineering from the Massachusetts Institute of Technology in 1944.

J. P. Chisholm

From 1944 to 1945 he served as an instructor in the Radar School conducted by MIT. From 1945 to 1947 he was an assistant in the department of electrical engineering while pursuing studies leading to the M.S. degree in electrical engineering in 1947. From 1947 to 1950 he served as an instructor in the same department, spending approximately one-half of this time in the Research Laboratory of Electronics working on the design of airborne telemetering equipment for the United States Navy. He is now employed as an electrical engineer by the Bell Aircraft Corporation, Buffalo, N. Y.

Mr. Chisholm is a member of Sigma Xi.

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William Dite was born in New York, N. Y., on March 2, 1919. He received the B.E.E. degree from the College of the City of New York in 1940. In 1950, he was awarded the M.E.E. degree from the Brooklyn Polytechnic Institute. From 1940 to 1943 he was associated with the Signal Corps Laboratories, at Fort Monmouth, N. J., where he worked on sound-ranging and radar equipment. Since 1943, Mr. Dite has been employed by the Federal Telecommunication Laboratories, in Nutley, N. J., where his work is largely concerned with communication systems employing pulses.

William Dite

Mr. Dite is a member of the American Institute of Electrical Engineers.

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Gerald W. Farnell was born in Toronto, Canada, on August 31, 1925. He received the B.A.Sc. degree in electrical engineering from the University of Toronto in 1948. From 1943 to 1945, he served as an instructor in the Canadian Army, returning to the University of Toronto in October, 1945, to complete his undergraduate studies.

G. W. Farnell

From September, 1948, to 1950 he was employed as a research assistant in the Research Laboratory of Electronics at the Massachusetts Institute of Technology, where he worked on the design of airborne telemetering equipment for the United States Navy. During this period, he did graduate work at the Massachusetts Institute of Technology and received the S.M. degree in electrical engineering in September, 1950. He is now employed as a lecturer at McGill University, Montreal, Canada.

Mr. Farnell is a member of the American Institute of Electrical Engineers.

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R. L. Garman (A'39) was born in Stevens, Pa., on July 21, 1907. In 1929 he received the B.S. degree from Franklin & Marshall College, and in 1931 and 1932 the M.Sc. and Ph.D. degrees, respectively from New York University. During the years 1929 to 1932 he served as a Fellow at New York University and was appointed to the teaching staff in 1932, serving progressively as assistant instruc-

R. L. Garman

tor and assistant professor in the department of physical chemistry. In 1942 he joined the staff of Radiation Laboratory at the Massachusetts Institute of Technology. At the termination of the recent war period, he assumed his present duties as director of research of General Precision Laboratory, Pleasantville, N. Y.

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Clay K. Hadlock (A'45) was born on November 26, 1919, in Lost Hills, Calif. He received the B.S. degree in 1942 from the University of California, after which he joined the staff of the Philco Corporation, as an electronic engineer. In 1946 he left Philco and became associated with the University of California at Los Angeles, Calif., both as a teacher of engineering and as a research associate.

Clay K. Hadlock

In 1950 Mr. Hadlock was made a research engineer at Hughes Aircraft Company, in Culver City, Calif., where he is now engaged in systems analysis.

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Edward I. Hawthorne was born on September 25, 1921, in Poland. He came to the United States in November, 1937, and became a naturalized citizen in 1943.

E. I. Hawthorne

After graduation from the Cooper Union School of Engineering in February, 1943, he joined the RCA Victor Division in Camden, N. J., as a test equipment design engineer. He served in the United States Army from September, 1944, to October, 1946. After a short stay with the RCA Victor Division as a transmitter design engineer, Mr. Hawthorne was appointed as an instructor in electrical engineering at the Moore School of Electrical Engineering of the University of Pennsylvania in February, 1947, where he has since been engaged in teaching, research, and graduate work toward the Ph.D. degree. He received the M.S. degree in electrical engineering in 1948.

Mr. Hawthorne is an associate of the AIEE, and a member of Tau Beta Pi and Sigma Xi.

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For a photograph and biography of Ronold King, see page 1218 of the October, 1950, issue of the Proceedings of the I.R.E.

S. M. Kaplan (S'48–A'50) was born in Baltimore, Md., on July 25, 1921. Before entering the armed service, he attended the Johns Hopkins University. During the war, he served overseas with the U. S. Army Signal Corps. Following his discharge, he entered the University of Maryland, from which he obtained the B.S. degree in electrical engineering in 1948. He joined the General Electric Company in 1948 on the Rotating Engineering Program, and after a short while became a member of the General Engineering and Consulting Laboratory, where he has been engaged in radar electronic development. He is, at present, doing graduate work at Union College.

S. M. Kaplan

Mr. Kaplan is a member of Tau Beta Pi and Phi Kappa Phi.

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Lawrence F. Koerner (A'29) was born in Niles, Mich., on February 25, 1897. He received the B.S. degree from Colorado College in 1923, and and in the following year the M.S. degree from Harvard University. He joined the staff of Bell Telephone Laboratories, Inc., in 1924, working first on the development of vacuum-tube oscillators, detectors, and frequency-measuring equipment. Since 1928 he has concentrated on developing crystal units and crystal oscillator circuits for transmitters and receivers. Mr. Koerner holds several patents on crystal equipment and is the author of articles on the testing and calibration of quartz crystals.

L. F. Koerner

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Don Lebell (A'49) was born on May 8, 1926, in California. During the period from 1944 to 1946, he was on active duty in the United States Naval Reserve, after which he received the B.A.S. degree from the University of California, in Berkeley, Calif. In 1947 he was awarded the B.S. degree, also from the University of California.

Don Lebell

Mr. Lebell has been employed by that University, at Los Angeles, as an instructor in engineering and then as an assistant professor. He is currently in charge of the operation of the University's mechanical differential analyzer. In 1949 Mr. Lebell received the M.S. degree from the University of California at Los Angeles.

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R. W. McFall was born in Moline, Ill., on July 2, 1921. He was graduated from the University of Maryland in 1943 with the B.S. degree in electrical engineering, after which he joined the General Electric Company. During the period 1943–1945, he was associated with the Company's test program. In 1945 he joined the General Engineering and Consulting Laboratory, where he is now a section engineer in the electromechanical division.

R. W. McFall

Mr. McFall is a graduate of the GE Advanced Engineering Course. He is also a member of Tau Beta Pi and Phi Kappa Phi, and of the AIEE.

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Philip F. Panter (A'43–SM'48) was born in 1908 in Poland. After early schooling in Tel-Aviv, Palestine, he later received at McGill University, Montreal, P.Q., Canada, the following degrees: B.Sc. in 1933; B.Eng. in electrical engineering, in 1935; and Ph.D. in physics, in 1936. He continued research in spectroscopy at McGill for an additional year.

P. F. Panter

After teaching mathematics and physics in Palestine for a year, Dr. Panter returned to Canada as an assistant professor of mathematics and physics in the evening division of Sir George Williams College in Montreal. He served also on the staff of the physics department of McGill University until late 1945.

Early in 1941 Dr. Panter joined the transmitter department of the Canadian Marconi Company in Montreal. In October, 1945, he was appointed senior engineer, responsible for the development of FM broadcast equipment, at Federal Telephone and Radio Corporation. He later transferred to Federal Telecommunication Laboratories, and was in charge of the theoretical group of the communication division. He is now on an to International Standard Electric ɔrporation as a technical adviser on some dliotelephone installations to be made in rael. Dr. Panter is a member of the Radio lub of America.

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Abd El-Samie Mostafa was born in airo, Egypt, on April 27, 1917. He re- ɛived the B.Sc. degree in June, 1937, and the Ph.D. degree in 1946, both in electrical engineering from Fouad I University, in Egypt. He has held the posts of teaching assistant and lecturer at Fouad University, and was promoted to associate professor in 1948. He is now professor of radio engineering, associated with the esearch section in radio and telecommunication work. During the period 1946 to 1948, Dr. Mostafa also constructed and supervised he electrical laboratories, and assisted in the esearch activities of the Fouad I University lectrical department.

ı E.-Samie Mostafa

Dr. Mostafa was elected an associate nember of the IEE (London) in June, 1948, ınd was awarded the Fouad El-Awal prize or science in Egypt for 1950.

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A. I. Pressman (M'47) was born in Brooklyn, N. Y., on August 30, 1916. He received the B.S. degree in physics from Brooklyn College in 1938, and the M.S. degree in physics from Columbia University in 1939.

A. I. Pressman

From 1940 to 1941 he was a junior physicist in the Corps of Engineers and from 1941 to 1946 he was a Signal Corps Radar Officer in charge of radar repair and maintenance depots and teams in various commands in the ETO. During the period 1946 to 1947 Mr. Pressman was an associate engineer at the Sperry Gyroscope Co. He has been at the Brookhaven National Laboratory working on frequency control problems for a three-billion-volt proton synchrotron since 1947.

Charles H. Papas (S'41–A'42) was born in Troy, N. Y., on March 29, 1918. He received the B.S. degree in electrical engineering from the Massachusetts Institute of Technology in 1941. He attended Harvard University on a Gordon McKay Scholarship from 1945 to 1948, receiving the M.S. degree in 1946 and the Ph.D. in 1948.

C. H. Papas

From 1941 to 1945 Dr. Papas was associated with the Naval Ordnance Laboratory and the Bureau of Ships, Washington, D. C.

Since 1948 Dr. Papas has been a research fellow at Harvard University, working on boundary value problems in electrodynamics.

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Jenny E. Rosenthal was born in Moscow, Russia, on July 31, 1909. She did her undergraduate work in France, receiving the Sc.B. degree from the University of Paris in 1926. After graduating from New York University with the Ph.D. degree in physics in 1929, she was for two years National Research Fellow at Johns Hopkins University. Subsequently she was research associate at New York University, Fellow of the American Association of University Women at Columbia, and instructor in the Graduate Division at Brooklyn College.

J. E. Rosenthal

Until the war her research work was in the fields of applied mathematics and spectroscopy, particularly the spectra of polyatomic molecules. Since joining the staff of the Signal Corps Engineering Laboratories in 1942, she has worked in electronics, particularly pulse networks and vacuum-tube design.

Although all her publications appear under her maiden name, Civil Service regulations require her to be known in the Signal Corps Laboratories under her married name of Bramley.

Dr. Rosenthal is a member of the American Mathematical Society and a Fellow of the American Physical Society.

Hubert J. Schlafly (S'41–A'42–SM'47), was born in St. Louis, Mo., on August 14, 1919. He received the B.S. degree in electrical engineering from the University of Notre Dame in 1941.

H. J. Schlafly

Following graduation he was employed by the General Electric Company and became associated with the Advance Development Laboratory of their electronics department. In February, 1944, he was assigned by GE to the Radiation Laboratory at the Massachusetts Institute of Technology to assist in design and production engineering of a gunfire control system. After the war, in the marine engineering section of GE, he was concerned with the commercial design of navigational radar equipment. In 1947 he joined the research division of Twentieth Century-Fox Film Corporation as director of their newly formed television laboratory.

Mr. Schlafly has served on the IRE Subcommittee 23.2 of the Video Recording Committee since 1948. He is a member of the Society of Motion Picture and Television Engineers and is active in the Films in Television Committee of that Society.

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Edwin E. Spitzer (A'28–M'38–SM'43) was born at Fitchburg, Mass., on February 22, 1905. He received the B.S. and M.S. degrees in electrical engineering from the Massachusetts Institute of Technology in 1927. Mr. Spitzer was associated with the research laboratory and the vacuum-tube engineering department of the General Electric Company from 1927 to 1933.

Edwin E. Spitzer

From 1933 to date he has been with the engineering section, Tube Department, RCA Victor Division of the Radio Corporation of America, and is now manager of the power and gas tube development group at the Lancaster, Pa., plant.

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For a photograph and biography of E. M. Williams, see page 1220 of the October, 1950 issue of the Proceedings of the I.R.E.

Institute News and Radio Notes

Institute of Radio Engineers Elects New Officers for 1951

Ivan S. Coggeshall, general traffic manager of Western Union Telegraph Company's overseas communications, has been elected President of The Institute of Radio Engineers for 1951. He succeeds Raymond F. Guy, manager of radio and allocation engineering of the National Broadcasting Company.

Jorgen C. F. Rybner of Copenhagen, Denmark, professor of telecommunications at the Royal Technical University of Denmark, will serve as Vice-President. The author of Danish and English textbooks on network theory, he will succeed Sir Robert Watson-Watt of England.

Election of the 1951 Directors has been announced as follows: *Directors-at-Large*, William H. Doherty, director of electronic and television research for Bell Laboratories, Murray Hill, N. J., and George R. Town, associate director of the engineering experiment station at Iowa State College, Ames, Iowa; *Region 2*, Harry F. Dart, office manager of the electronics engineering department of Westinghouse Electric Corporation, Bloomfield, N. J.; *Region 4*, Paul L. Hoover, head of the department of electrical engineering, Case Institute of Technology, Cleveland, Ohio; *Region 6*, William M. Rust, Jr., head of geophysics research for Humble Oil and Refining Company, Houston, Texas; and *Region 8*, Allan B. Oxley, chief engineer of RCA Victor Company, Montreal, Quebec, Canada.

Calendar of COMING EVENTS

AIEE-IRE-NBS High-Frequency Measurements Conference, Hotel Statler, Washington, D. C., January 10–12

1951 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 19–22

IRE Southwestern Conference, Dallas, Texas, April 20–21

1951 Convention of SMPTE, April 30–May 4, Hotel Statler, N. Y.

1951 Annual Meeting of the Engineering Institute of Canada, Mount Royal Hotel, Montreal, May 9–11

1951 IRE Technical Conference on Airborne Electronics, Biltmore Hotel, Dayton, Ohio, May 23–25

1951 Summer General Meeting of AIEE, June 25–29, Royal York Hotel, Toronto, Canada

1951 IRE West Coast Convention, San Francisco, Calif., August 29–31

New Booklet Is Available on Engineering College Research

The Engineering College Research Council of the American Society for Engineering Education has published a booklet, "Research Is Learning," compiled by representatives of 83 leading engineering schools. The booklet contains a comprehensive definition of the role of research in engineering education, together with twelve brief summaries of universities' research activities in a wide variety of technical fields.

Copies of "Research Is Learning" are available without cost on request to the Secretary of the Engineering College Research Council, Room 7-204, 77 Massachusetts Ave., Cambridge 39, Mass.

Technical Committee Notes

The **Standards Committee,** under the Chairmanship of Professor J. G. Brainerd, held a meeting on October 19. The following Standards have been approved and appear in this issue of the PROCEEDINGS: **Standards on Circuits: Definitions of Terms in Network Topology;** and **Standards on Television: Methods of Measurement of Electronically Regulated Power Supplies.** . . . A new Committee to be known as the **IRE Coordination Committee with International Organizations** has been formed. This Committee will be the means of liaison with CCIR and other international standardizing bodies with which IRE is concerned. Axel G. Jensen will serve as Chairman. The duties of this Committee will be to report directly to the Executive Committee; to work through the U. S. Department of State, rather than with international bodies; and to ascertain the needs of CCIR in which the IRE may have responsibilities and assist in solving CCIR problems. When CCIR or any other body turns to the IRE for assistance, the new Committee will then set up an ad hoc group to study the problem and provide the required solution. When the Standards Committee feels that a matter should be brought to the attention of an international standardizing body, it will be referred to this Committee. . . . Work is progressing towards the revision of the **Master Index of IRE Definitions.** Copies of this publication will be available in late January. The Standards Committee has drawn up a new procedure to be employed when submitting proposed definitions for standardization. Proposed definitions are to be sent through the Definitions Co-ordinator to members of the Standards Committee and to chairmen of all technical subcommittees concerned with definitions. In the case of chairmen of subcommittees, several copies will be sent to each chairman. A period of one month will be allowed for comments which will be mailed to the chairman of the sponsoring committee. An additional period of from two to four weeks will be allotted to the sponsoring committee, if necessary to review the comments and incorporate them in definitions. If the comments are ignored, the sponsoring committee should be prepared to explain its action. The definitions will then go to the Standards Committee for action at its next meeting. The goal of this procedure is to eliminate most of the technical comments before the Standards Committee meeting. It has been suggested that when definitions are submitted to the Standards Committee, controversial ones must be accompanied by an explanation stating the sponsoring committee's reasons for deciding on the specific proposal.

A meeting of the **Circuits Committee** was held on October 27, W. N. Tuttle, Chairman, presiding. Comprehensive reports of the activities of the subcommittees were given by the chairmen.

A meeting of the **Antennas and Waveguides Committee,** under the Chairmanship of A. G. Fox, was held on October 24. Work is progressing on the definitions of waveguide terms and transmission line terms.

The Electron Tubes and Solid-State Devices Committee held a meeting on November 2, under the Chairmanship of L. S. Nergaard. The Committee has drafted outlines for **Proposed Standards on High-Frequency Tubes: Definitions of Terms and Proposed Standards on High-Frequency Tubes: Methods of Testing.**

The Industrial Electronics Committee, under the Chairmanship of D. E. Watts, held a meeting on November 2. Reports on the progress of work in the various subcommittees were given by the chairmen.

The IRE will participate in the **Annual Convention of the Institute of Aeronautical Sciences,** to be held January 29, 30, 31, and February 1. The fourth day of the Annual IAS Convention (February 1), will be devoted to Electronics in Aviation. The morning session of February 1 will present the electronic problems of the aviation people, and the aviation problems of the electronics group will be presented in the afternoon.

A successful **Joint IRE/AIEE Third Annual Conference on Electronic Instrumentation in Nucleonics and Medicine** was held on October 23, 24, and 25, 1950, at the Park Sheraton Hotel, New York, N. Y. One of the highlights of the three-day Conference was the round-table discussion on "The Effects of Atomic Weapons" by a five-man panel of experts, at which W. R. G. Baker presided. The discussion was particularly interesting and enlightening, because it was specifically intended to acquaint the layman with the simple facts concerning protection against atomic attack. Simultaneously with the Conference, the Second Annual Nucleonics Manufacturers' Exhibit was held on October 24 and 25, at the Park Sheraton Hotel. Scalers, Geiger tubes, survey meters, scintillation counters, and industrial instruments sensitive to alpha, beta, and gamma radiation were featured.

The Joint Technical Advisory Committee held a meeting on October 24 under the Chairmanship of J. V. L. Hogan. A supplemental report prepared by JTAC has been presented to the Federal Communications Commission.

Professional Group Notes

The IRE Professional Group, under the Chairmanship of R. F. Rollman, has divided the Sections of the Institute in three Regions for the purpose of carrying on its activities with greater effectiveness. Each Group Region is represented on the Administrative Committee of the Group by a Regional Chairman. The functions of Regional Chairmen will be as follows: Facilitate the exchange of ideas between the Administrative Committee and the local Sections, maintain an active program in the Region of preparing and delivering papers, co-operate with Standing committees, and appoint "sparkplugs" as representatives of the Group in each Section within the Region. . . . **The Broadcast Transmission Systems Group** has mailed 2,231 letters together with IRE membership application blanks to chief engineers of broadcast stations and chief engineers of manufacturers of broadcast equipment. Chairman Lewis Winner reports that the response, both as to applications for Group membership and IRE membership, has been exceptionally good. During a meeting of the Group Administrative Committee on October 27, plans were discussed for the formation of a local Group in Cleveland. Plans were also discussed for "Broadcast Day" to be held on Tuesday, March 20, during the 1951 IRE National Convention. Papers on the following tentative subjects were suggested for presentation on "Broadcast Day": Maintenance of TV Broadcast Transmission Equipment; The Manpower Problem in Broadcasting; Electronic Instrumentation Covering AM, FM, and TV Instrumentation at the Broadcast Station Through Use of the Oscillograph; Substitution in Transmitters; Color Television or UHF; Radio Silencing During Wartime; Phase Modulation; and TV Recording. In addition to the papers, an effort will be made to arrange for an exhibit of broadcast equipment produced during the last thirty years to be displayed at Grand Central Palace during the Convention. . . . **The Professional Group on Instrumentations** plans to sponsor a program of technical papers during the 1951 IRE National Convention. H. L. Byerly was appointed to represent the Group on the Convention Technical Program Committee. The Administrative Committee, under the Chairmanship of Ernst Weber, discussed at its September 25 meeting plans for mailing informal news bulletins to all members of the Group. . . . **A National meeting of the IRE Professional Group on Vehicular Communications** was held on November 3, in Detroit, Mich. Austin Bailey is Chairman of this Group. Inspector E. C. Denstaedt of the Detroit Police served as Chairman of the sessions, which were well attended and enthusiastic. As a result of this meeting much interest has been aroused in potential new members for IRE, this Group, and other IRE Professional Groups. It has been tentatively decided to hold the 1951 national conference of the Group in Chicago, if a Section Group is operating there. A two- or three-page newsletter containing summaries of the eight papers presented at the 1950 National Meeting will be sent to all Group members. The Group is planning to sponsor a symposium at the 1951 IRE National Convention. . . . The **IRE Professional Group on Nuclear Science,** co-sponsors of the Joint IRE AIEE Third Annual Conference on Electronic Instrumentation in Nucleonics and Medicine, held a national business meeting in conjunction with the Conference at the Park Sheraton Hotel on October 24. M. M. Hubbard, Chairman, presided at the meeting. . . . **The Audio Group** is sending out periodic newsletters to all of its membership. It recently arranged for the three audio papers which were presented at the 1950 National Electronics Conference, and is planning to sponsor one or more symposia at the 1951 IRE National Convention to include a panel on "Loudspeakers." An Audio Center is being planned for the third floor of Grand Central Palace during the Convention which will include both exhibit booths and special sound demonstration rooms. . . . The IRE/RTMA Radio Fall Meeting was held October 30, 31, and November 1, 1950, at the Hotel Syracuse, N. Y. J. S. Steen of the **IRE Professional Group on Quality Control,** presided over a session sponsored by this Group. A television session was sponsored by the **IRE Professional Group on Broadcast and Television Receivers.** R. A. Hackbusch of this Group presided. O. L. Angevine, Jr., presided at the Audio session, which was sponsored by the IRE Professional Group on Audio. All of these sessions proved to be of great benefit and stimulated interest in many other IRE Professional Groups. Application for Group membership may be obtained from the Technical Secretary, at IRE Headquarters.

Signal Corps Laboratories Report Over 700 Vacancies

The Signal Corps Engineering Laboratories and The Signal Schools comprising the Signal Center at Fort Monmouth, N. J., have over 700 vacancies for civilians who qualify as instructors, technical writers, and professional engineers in the field of communications and electronics. These positions must be filled as soon as possible. Applications will be accepted from either veterans or nonveterans, regardless of whether the applicants have taken Civil Service examinations, according to John D. Sullivan, Chief of Civilian Personnel.

Although veterans are accorded preference, there are a sufficient number of openings to provide for the appointment of nonveterans who can qualify. The salaries for these positions range from $3,825 to $6,400 a year.

Other job openings at Fort Monmouth include physicist, mechanical engineer, photographic engineer, and engineering draftsman at salaries ranging from $2,650 to $5,400 a year. No clerical or administrative personnel are needed at this time.

The principal research and developmental laboratories for the Signal Corps, as well as the training center for Signal Corps military personnel, are located at Fort Monmouth.

Applicants for electronic engineering positions should have at least one year of professional experience in addition to an appropriate bachelor's degree or a master's degree for the salary of $3,825 a year. Additional experience may be qualifying for higher salaries up to $6,400.

Applicants with four or more years of practical experience in the construction, maintenance, or repair of radio and radar are especially needed for instructor positions. Men with teaching experience in addition to their technical backgrounds in these fields may be employed at salaries ranging from $3,825 to $4,600 a year. Openings are in microwave radio relay, radar, radio electronics, fixed station radio, central office techniques, teletype installation and maintenance, repeater and carrier, and dial central office techniques.

Technical writers with 3½ to 5½ years of experience in preparing reports, manuscripts, or manuals dealing with electronics, radio, and radar or communications may qualify for positions at salaries of $3,825 to $5,400 a year.

Applicants should complete Standard Form 57, which is available at any first or second class post office, and submit it to the Civilian Personnel Branch, Building T-530, Fort Monmouth, N. J. For applicants who can report in person, interviews will be conducted between 8:00 a.m. and 3:00 p.m., Tuesday through Friday.

NYU Students Demonstrate 1928 Television Broadcast

A public demonstration of how a television picture looked in 1928, when station WRNY began the first regular daily broadcasting of "live" images, was held by the IRE and AIEE Student Branches at New York University on November 8. Because much of the transmitting and receiving equipment no longer exists, Student Branch members exhibited replicas and employed substitute methods to produce images.

Col. Robert Hertzberg, one of the technicians who helped with the 1928 broadcast, acted as master of ceremonies during the presentation of scrolls to Isidore Goldberg, president of Pilot Radio Corp., and Hugo Gernsback, publisher of *Radio-Electronics* magazine. Pilot Radio Corporation designed and developed the equipment used in 1928 and station WRNY, owned by Mr. Gernsback, furnished the radio broadcast channel.

The televising equipment used in the 1928 broadcasts included an electric arc light, photoelectric cells, and a 2-foot scanning disk pierced by a spiral of 48 holes which was rotated at 450 rpm. The television signals were broadcast simultaneously over WRNY on 920 kc and W2XAL on 9.7 Mc, using a bandwidth of only 5 kc. The WRNY transmitter was located at Coytesville, N. J., atop the Palisades across the Hudson River from New York City. An estimated 2,000 standard radio sets in the New York City area were equipped at that time for television, employing a neon-gas "glow lamp" and a scanning disk identical to the one used at the transmitter and operated in synchronism with it. The resulting 48-line 7.5-frame image, when originally demonstrated at NYU in 1928, was viewed on a screen 1½ inches square and was magnified by a lens to twice that size.

The Professional Group Activity of the IRE

The growing importance of the IRE Professional Groups has led the Chairman of the Committee on Professional Groups, Dr. W. R. G. Baker, to prepare the following constructive suggestions to those desiring to form a new Group and also to those already active within a Group. Members of the Institute are urged to familiarize themselves with the following so that they may derive maximum benefit from the Group activities.—*The Editor.*

Because the Professional Group activity of the IRE has proven so eminently successful and because of the wide interest in this activity, it has been suggested that an abstract of the "Manual for IRE Professional Groups" be published in the PROCEEDINGS.

One of the greatest values of the Professional Group has been to bring together, under the sponsorship of the IRE, those members having interest and enthusiasm in a specific phase of electronics. The activities of these Groups have stimulated IRE members to increase their knowledge in these fields and have inspired a wider dissemination and communication of ideas.

Through the Professional Groups, membership in the IRE gains even greater value.

The establishment of Professional Groups within The Institute of Radio Engineers for those members who have interests in specialized fields has proven extremely successful. These Professional Groups are voluntary associations of Institute members who have formed the equivalent of a national "society within a society" around a subject of mutual interest.

Ten of these Professional Groups have been formed in the IRE. These are: Antennas and Propagation, Audio, Broadcast and Television Receivers, Broadcast Transmission Systems, Circuit Theory, Instrumentation, Nuclear Science, Quality Control, Radio Telemetry and Remote Control, and Vehicular Communications. In addition, a petition has been received for the formation of an IRE Professional Group on Airborne Electronics.

As can be seen, the subjects around which the Professional Groups may be formed may be broadly functional, narrowly restricted, or may fall in a "field of use" category. Since the Group system permits development of the IRE as an integrated technical society with a wide variety of interests, the Board of Directors of the IRE has adopted the Group principle to enable its members to organize and still make use of the machinery, experience, and publication channels of the IRE.

For the IRE member who may be interested in forming a new Professional Group, the following steps are suggested:

1. Communicate with one or more colleagues to consider the field of interest in relation to those already organized.

2. Communicate, through the Technical Secretary at Institute Headquarters, with the Chairman of the Committee on Professional Groups who will provide the promoter with advice, copies of the "Manual for Professional Groups," and possibly names of other members who might be interested. These communications should be in triplicate and should outline the prospective scope of interest in sufficient detail to enable the Committee to co-ordinate it with the fields of interest of existing groups.

3. The promoter should secure members to join with him in signing a petition for establishment of the new Group. (The form of the petition is included in the Manual.) The petition which must be signed by at least 25 members of Associate or higher grade-and is then forwarded through the Technical Secretary for approval by the Committee on Professional Groups and the Executive Committee of the Institute. As a part of the petition, a list of 6, 9, 12, or 15 names, including the promoter's, must be submitted for appointment to the Group's initial Administrative Committee.

4. Unless the promoter requests otherwise, he will be appointed temporary chairman of the Administrative Committee of the Group. His first duty is to call a meeting of the Committee to elect a regular chairman and draw up a constitution and by-laws, the form of which is specified in the Manual. It should particularly be noted that, although each Group is permitted to delineate qualifications for memberships in the Group beyond those for Institute membership, the Institute favors admission of all IRE members except Students into any Group on an equal basis. Group membership may be national and cut across Sections and Regions. In a typical case an Institute member might belong to one Section, and to one or more Groups.

5. Headquarters will assist the Groups by circularizing members, maintaining membership lists, and handling a moderate amount of correspondence. Although it is believed that most Groups will be self-supporting eventually, it is expected the Board of Directors may in special cases provide funds for Groups not yet firmly established.

6. One of the first and important objectives of a new Group should be to study the publication needs of its members. Equal in importance to the holding of meetings should be the solicitation from Group members of papers to be sent in to the PROCEEDINGS OF THE I.R.E. for possible publication. The Group's committee system should include a committee set up specifically to study this field of activity, solicit papers on newer developments, and arrange lectures or talks if formal written papers are not presented. The PROCEEDINGS will be the standard avenue of publication for all Groups, although certain additional forms of publication may be used for conferences, group symposia, and the like.

Because of the importance of symposia to the Professional Groups, some suggestions for their planning and execution are listed below:

The Group sponsoring a symposium should appoint representatives to a Steering Committee to guide the symposium. If the symposium is to be sponsored jointly with another organization or organizations, each sponsor should nominate representatives.

The Steering Committee would be responsible for the organization and management of all other committees necessary for administration of the symposium, and an IRE Headquarters representative would be named to serve on the Committee upon request to the Professional Group Co-ordinator.

The Steering Committee would be responsible for policy and finances to each of the sponsor organizations, including IRE Headquarters. The Committee should submit to the Executive Committee its proposals for sponsorship and financing, and disposition of profits or losses. It must render a complete report covering the results of the symposium within 90 days after the meeting. This Committee would repay any outstanding loans and divide any profit from the operations among the sponsors.

The Technical Program Committee would have the responsibility of securing speakers, establishing dates for submission of abstracts of technical papers, and selecting and designating chairmen of the various sessions, subject to the approval of the Steering Committee.

The Public Relations and Publicity Committee would assist in obtaining prominent speakers for special functions such as luncheons, dinners, banquets, or key meetings; assist in the suggestion of topics for key talks, and offer editorial assistance. This committee also would select such technical papers which would insure audience and press interest and prepare appropriate pre-conference releases for the press.

The Finance Committee would handle financial matters in accordance with the Manual for Professional Groups and the regulations of the IRE. The Local Arrangements Committee would be responsible for all arrangements involving facilities, registrations, printing of programs, and entertainment.

More detailed suggestions and instructions for the formation and operation of professional groups is included in the Manual For Professional Groups. Copies of this manual may be obtained by writing to the Technical Secretary at Institute Headquarters.

If the Group desires to publish Transactions containing the papers delivered before any Group Conference or Symposium, it should promptly communicate with the Editorial Department of the Institute which will arrange for such publication, provided the text of the corresponding papers is made available in time, and provided the Executive Secretary approves the financial arrangements, required number of copies of the Transactions, cost per copy to the Group members and others, and any contractual arrangements proposed by the Group to defray some or all of the cost of printing and distributing such Transactions.

7. National meetings and conferences within the field of activity of a Group may be promoted without formal permission, but approval of IRE Headquarters must be obtained to prevent conflict with other IRE activities. Through co-operative action with the Technical Program Committee of the IRE National Convention, a Group may arrange symposia, meetings, or sessions at the Annual or Regional Conventions. Joint meetings may be held by a Group with Sections, Group divisions, or other societies. Approval for such joint meetings must receive the approval of the IRE Executive Committee.

1951 IRE National Convention News

THE WORLD-FAMOUS Waldorf-Astoria Hotel in New York City will share its facilities with those of the Grand Central Palace to accommodate the 1951 IRE National Convention on March 19 through 22, an event of such importance that more than 18,000 engineers and scientists from all parts of the world are planning to attend. Plans for the convention are rapidly nearing completion, and all indications show that, under the guidance of the Convention Committee, the theme of the convention, "Advance with Radio-Electronics," will be faithfully carried out by the outstanding program of technical papers and exhibits.

Technical Program

Under the chairmanship of Ernst Weber, the Technical Program Committee in co-operation with IRE Professional Groups, is planning a papers program of particular interest and significance this year, consisting of some 36 sessions and about 175 papers. Particularly noteworthy is the role that the Professional Groups are playing in organizing portions of the program. At last count, seven Professional Groups are planning eleven special symposia on subjects of particular and timely interest in their respective fields. In addition, all Groups are assisting the Technical Program Committee in scheduling regular technical sessions comprising voluntarily contributed papers in their special fields.

The Technical Program Committee is planning to include again the special Tuesday night symposium on Television, which has become such a popular feature of past conventions. This year's symposium will feature a panel of several leading engineers from companies currently engaged in color television research and development who will discuss the latest technical advances which have been achieved in their respective laboratories, a session which promises to be of extreme timeliness and importance.

Technical Sessions will be held in the Grand Ballroom, the Jade Room, and the Astor Gallery on the third floor, and the Wedgwood Room on the main floor of the Waldorf-Astoria Hotel; and in the Maroon and Blue Halls on the third floor of the Grand Central Palace, a convenient two blocks from the Hotel.

Exhibits

The Grand Central Palace will be the scene of what has become the most comprehensive technical display of radio-electronic apparatus and their applications in the United States. Exhibits Manager William C. Copp reports that this year's Radio Engineering Show will be larger than ever, with 263 exhibitors completely filling three floors of the Grand Central Palace. The exhibits will run the entire gauntlet of products from subminiature circuits to complete communications systems, affording engineers an unparalleled opportunity for visual examination of the latest products in every field and consultation with the manufacturers' representatives.

Several noteworthy features are being planned for the third floor of the Palace. All of the exhibits pertaining to the audio and nuclear fields are to be consolidated in an audio center and a nuclear center, respectively, on the third floor. The audio center will include 8 sound theaters for the demonstration of the audible features of the audio equipment. Technical sessions covering the audio and nuclear fields will be held in the nearby Maroon and Blue Halls for the convenience of those interested in these fields. The Halls themselves have been redesigned with a view to improving the seating arrangement. Also on the third floor will be the Government-sponsored exhibits which have been allotted a larger area this year, occupying the entire Park Avenue side of the Palace.

Social Events

A "get-together" Cocktail Party will be held on the first evening of the convention, March 19, on the Waldorf-Astoria's Starlight Roof, famous the world over. In this setting of world renown, members and guests will have an excellent opportunity to renew old acquaintances and make new ones.

The traditional President's Luncheon on March 20 will also be held on the Starlight Roof honoring the retiring president, Raymond F. Guy, and the newly elected president, Ivan S. Coggeshall. In addition, a popular entertainment feature is being planned, details of which will be announced later.

The Annual IRE Banquet will be held in the Grand Ballroom of the Waldorf on Wednesday evening, March 21, at which time the Institute awards for 1951 will be presented by President Coggeshall. A speaker of national prominence will give the keynote address.

Women's Activities

Mrs. Raymond F. Guy, Chairman of the Women's Activities Committee, reports that a particularly attractive program for the ladies is being planned, including an intriguing "technical session for women only." Other activities will include a fashion show and behind-the-scenes tour at the Waldorf, a trip to the United Nations, a television broadcast, and a matinee performance at a leading Broadway show.

Further Details

As convention plans progress they will be reported in these pages in subsequent issues, and the Convention (March) issue of the Proceedings will contain a complete program of events, together with 100-word abstracts of all papers to be delivered at the convention.

IRE People

George W. Bailey (A'38–SM'46), Executive Secretary of The Institute of Radio Engineers, has been elected a national director of the Armed Forces Communications Association for a term extending through the year 1954.

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Harry Stockman (A'42–M'44–SM'45) has been appointed director of research of the Tobe Deutschmann Corporation in Norwood, Mass. As a scientific leader, Dr. Stockman is responsible for the theoretical phases of contract work for the government and the electronic industry.

Harry Stockman

Previously, Dr. Stockman has served as consulting engineer and scientist to several industrial firms in the greater Boston area, contributing to the development of new devices and weapons on government contracts. During the period 1945–1948, he was associated with USAF Research in Cambridge, and taught pre-radar and other causes at Harvard University from 1941–1945. A graduate of the Royal Institute of Technology, Stockholm, Sweden, and of Harvard University, Dr. Stockman has conducted fundamental research in the fields of nonlinear circuit theory and basic communication. Active in electronics research and development for more than twenty years, Dr. Stockman served on the Communications Panel of RDB in Washington, D. C., during 1947–1948, and has presented and published numerous scientific, and experimental papers, reports, patent communications, and text books.

He is a member of Harvard Sigma Xi, and has been awarded the decoration of the Liberty Cross for service in Finland.

❖

David S. Rau (A'22–SM'45) has been appointed assistant vice-president and chief engineer of RCA Communications, Inc. Mr. Rau joined RCA as a student engineer when he was graduated from the United States Naval Academy at Annapolis, Md., in 1922.

David S. Rau

He served in various engineering capacities with the company in New York, California, and the Philippines prior to World War II, in which he rose from the rank of lieutenant commander to captain on the staff of the Director of Naval Communications.

Upon his return to the company from the service, he was promoted to the post of assistant to the vice-president in charge of engineering.

Norman L. Winter (A'47–M'47) has been named the director of special electronic sales for Sperry Gyroscope Company. He will be responsible for all sales of products and engineering developments in electronic tubes and microwave components to military and nonmilitary customers in the United States and Canada. Mr. Winter will continue to coordinate Sperry activity in the air navigation and traffic control field.

Norman L. Winter

He was with the Research and Development Board as an executive director of the committee on electronics until he joined the Sperry organization in 1949. After five years of wartime service with the Signal Corps and Air Force, he now holds the rank of colonel in the U. S. Air Force reserve.

Mr. Winter is also a member of the Institute of Aeronautical Sciences and the American Association for the Advancement of Science.

❖

John N. Dyer (J'30–A'32–SM'45–F'49), supervisor of radar and air navigation research and development for Airborne Instruments Laboratory, Inc., Mineola, L. I., N. Y., has been named director of research and engineering of the company.

John N. Dyer

He succeeds **John F. Byrne** (SM'45–F'50) who recently resigned to take a position with Motorola Corporation, Chicago, Ill.

Mr. Dyer, who joined the Laboratory in 1945, was graduated from the Massachusetts Institute of Technology in 1931. He was in charge of radio engineering from 1933 to 1935 with Byrd Antarctic Expedition II. He joined Columbia Broadcasting System upon his return and worked on CBS television until 1942, when he became leader of the group developing vhf transmitters at the Radio Research Laboratory of Harvard University.

Early in 1944, Mr. Dyer became a director of the American British Laboratory in England. He was appointed head of the Radio Research Laboratory's field division after VE day. He joined Airborne Instruments Laboratory five years ago.

Mr. Byrne, who also served AIL as a vice-president, and was graduated from Ohio State University, was also associated with the Bell Telephone Laboratories and Collins Radio Company. He worked with Mr. Dyer at the Radio Research Laboratory at Harvard, where he became associate director early in 1945. In September of that year he joined Airborne Instruments Laboratory.

❖

Thomas H. Briggs (A'45–SM'45) has joined the research division of the Burroughs Adding Machine Company as research engineer in the special devices department.

Thomas H. Briggs

Mr. Briggs received the B.S. degree in physics from Wesleyan University in 1927, and the M.S. degree in physics from the California Institute of Technology in 1928. Upon his graduation, he became a member of the staff of the Westinghouse Lamp Company and later joined the Raytheon Manufacturing Company.

In 1939 he became chief engineer of the special purpose tube company as head of the electronics research and development laboratory.

He is chairman of Committee B4-VIII-A of the American Society for Testing Materials.

❖

Philips B. Patton (A'46), former field engineer, has become manager of the sales engineering department of Lenkurt Electric Co., Inc., carrier equipment and component manufacturer, San Carlos, Calif. Before joining the company, he was chief of FCC's Radio Telephone-Telegraph Section, Common Carrier Branch; flight radio officer with Pan American World Airways; and a field engineer and technical co-ordinator with Farnsworth Mobile Radio. He was previously associated for many years with Western Union Telegraph Company.

Philips B. Patton

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Raymond C. Miles (S'41–A'44–M'49), has been appointed deputy head of the design and test section of the Engineering and Production Division of Airborne Instruments Laboratory, Mineola, L. I., N. Y. He will assume full responsibility of the design and test section.

He comes to AIL after serving with Haller, Raymond and Brown, Inc., as chief engineer, and prior to that, with Federal Telephone and Radio Corporation as a design and development engineer.

Earl H. Schoenfeld (A'35–SM'45), project manager and electronic engineer for the W. L. Maxon Corporation of New York, N. Y., died on October 9 at Memorial Hospital in Manhattan. A resident of Mamaroneck, he was extremely active in the IRE New York Section in recent years, serving as Secretary during 1949–1950.

Mr. Schoenfeld, who was born in Berkeley, Calif., on October 23, 1909, was a graduate of Stanford University, from which he received the A.B. degree in 1932, and the degree of electronic engineering in 1934. He was a member of Sigma Xi and Phi Beta Kappa fraternities.

Upon his graduation he was employed by RCA Communications for two years and then became associated with Heintz and Kaufmann Ltd. of Canada. He was affiliated with U. S. Forest Radio Laboratory in 1937, and in 1942 he returned to RCA Laboratories with whom he remained until 1949.

He collaborated with other engineers in several publications, including "New Technique in Synchronizing Signal Generators;" "A Bar Signal Generator," "Electrostatic Deflection for the 7GP4 Kinescope," and "Improvement in Tuning Characteristics of FM Receivers."

❖

George W. Patterson (A'49) has joined he staff of the Research Division of Burroughs Adding Machine Company in Philadelphia, Pa.

G. W. Patterson

Mr. Patterson received the B.S. degree in electrical engineering from the University of Vermont in 1934, and the M.A. degree from Columbia University. He was a physicist in the National Hydraulic Laboratory, Bureau of Standards.

During the war he was on leave from the Bureau of Standards to the David Taylor Model Basin, U. S. Navy, and during 1943 to 1946 was engaged in work for the National Defense Research Council. He was a research associate and assistant professor at the University of Pennsylvania, Moore School of Electrical Engineering, until receiving his new appointment.

As a result of his special work during the war, Mr. Patterson was awarded the Naval Ordnance Development Award for exceptional civilian service.

Mr. Patterson is a member of the Association for Computing Machinery, Sigma Xi, Phi Beta Kappa. He is a member of the IRE Technical Committee on Electronic Computers.

Donald L. Stevens (M'49) has joined the staff of Burroughs Adding Machine Company's Research Division in Philadelphia, Pa.

Donald L. Stevens

He attended Northeastern University, in Boston, Mass., and received the B.S. degree in electrical engineering from Massachusetts Institute of Technology in 1945.

Mr. Stevens was formerly associated with Bendix Aviation Corporation, engaged in the testing and calibration of radio compass and other aircraft instruments.

During the war he was assistant to the electronic supply officer at U. S. Navy Electronics School at Corpus Christi, Texas.

In 1946 Mr. Stevens joined Sylvania Electric Products Inc., as head of the computer group. He was active in the engineering and development of electronic computing equipment, and supervised and co-ordinated work of ten engineers as well as technicians. He also helped develop apparatus for microflash ballistic photography.

He is a member of the technical Committee on Electronic Computers of the IRE, and also of the Association for Computing Machinery.

❖

Carl E. Scholz (M'26–SM'43), vice-president and chief engineer of the Mackay Radio and Telegraph Co., has been named to the same capacity for two other operating subsidiaries of the American Cable and Radio Corporation, the Commercial Cable Co., and All America Cables and Radio, Inc.

He has been associated with the International Telephone and Telegraph Corporation since 1917, when he joined the Federal Telegraph Co. as an engineer. He remained with Federal for 11 years, advancing to the position of chief engineer.

Except for a few years in South America, Mr. Scholz has spent most of his career at IT&T headquarters in New York, N. Y., working on engineering and design problems for the various subsidiary companies.

He has been a vice-president of Mackay since 1945, and a director since 1948. In February of this year he was appointed vice-president and chief engineer of Mackay, in charge of the company's engineering and plant department. He is also a member of the AIEE.

❖

Lorimer P. Brooks (A'50), formerly a senior patent attorney with the International Telephone and Telegraph Corp., New York, N. Y., has become associated with the law firm of Ward, Crosby and Neal, 225 Broadway, New York, N. Y.

Prior to his association with IT&T, Mr. Brooks was a radio engineer with the U. S. Signal Corps, stationed first at the Signal Corps Radar Laboratories and subsequently at the Naval Research Laboratory in Washington, D. C.

He was graduated with honor from Northeastern University, receiving the B.S. degree in electrical engineering, and he is a member of Tau Beta Pi. Also, he received the L.L.B. degree from Fordham University and is both a member of the New York Bar and registered as a patent attorney with the United States Patent Office.

❖

William D. Loughlin (A'19–M'29–F'25), chairman of the board of the Boonton Radio Corporation, manufacturers of radio-frequency precision measuring instruments, died on November 12 at the Community Medical Group Hospital.

W. D. Loughlin

Mr. Loughlin, who was one of the early experimenters in radio, designed, built and operated a 5-kilowatt wireless station while still in college. He was licensed as a first-class wireless operator.

During the first World War he did research and experimental work at the United States Naval Laboratory, Philadelphia, Pa., and later succeeded S. Ballantine as director of the research group, remaining until 1923, when he resigned to join the newly formed radio-frequency laboratories in Boonton.

He began his radio career with the company as a radio engineer, became vice-president and director in 1928, and president in 1931. The company's patents were sold in 1934 to the Radio Corporation of America and Mr. Loughlin formed the Boonton Radio Corporation. During World War II the company received the Army-Navy "E" award with four stars.

He was also a member of the Radio Club of America and the Franklin Institute.

❖

Harold G. Williams (S'50) has been appointed manager of the Beryllium Copper Division of the Lee Mechanical Laboratories. He was formerly chief metallurgist of Instrument Specialties Co., and has been closely associated with the development of beryllium copper as a spring material. The Lee Mechanical Laboratories is a newly formed division of Lee Spring Company.

❖

Walter Charles Gee (SM'47) has been appointed the Divisional Engineer in charge of all Telecommunications in New Guinea and Papua under the Ministry of External Territories.

Industrial Engineering Notes[1]

Television News

There were 3,107,000 **television sets shipped to dealers** throughout the country during the first seven months of 1950, according to RTMA estimates for the industry based on reports by member-companies. . . . The FCC has begun what appears to be another series of lengthy hearings in its inquiry into the feasibility of **utilizing the ultra-high frequencies for television broadcasting.** The Commission has been hearing testimony on the utilization of the uhf band 470 to 890 Mc for television broadcasting. Issues being considered by the Commission include a proposed FCC uhf allocation plan, a proposal of the Bell Telephone Laboratories to assign the band 470 to 500 Mc to broad-band mobile communications, Stratovision, Polycasting, reservation of channels for noncommercial educational TV stations, and metered television. The following are participating in the hearing: The FCC's Ad Hoc Committee under the chairmanship of Edward W. Allen, FCC engineer, and consisting of industry engineering experts; the Joint Technical Advisory Committee, appointed jointly by RTMA and IRE to advise government agencies; RTMA witnesses and representatives of the Allen B. DuMont Laboratories, Inc., Philco Corp., the Radio Corporation of America, Communication Measurements Laboratory, Inc., Westinghouse Electric Corporation; Collins Radio Co., Air King Products Co., Inc., and Federal Telephone and Radio Corporation. . . . With the opening of the hearing on the utilization of the ultra high frequencies for television broadcasting the FCC immediately found itself with another dispute on its hands as **members of the industry-government Ad Hoc Committee failed to agree** on the conclusions reported by the group. This action preceded the opening of the uhf question as the Ad Hoc Committee confined its findings to propagation factors concerning TV and FM broadcasting in the vhf band. Mr. deMars, Dr. Thomas J. Carrol, Jr., microwave research consultant to the NBS, and Chester H. Page, consulting engineer, were the three who refused to sign the report. It was endorsed "with reservations" by Stuart L. Bailey, consulting engineer; C. B. Jolliffe, executive vice-president in charge of the RCA Laboratories Division; Albert F. Murray, consulting engineer; Dr. Norton; Raymond M. Wilmotte and Frank G. Kear, consulting engineers, and by Jay M. Wright, CBS engineer. Those endorsing the report in full, in addition to Mr. Allen and other FCC engineers, included Robert P. Wakeman, head of the propagation department of Allen B. DuMont Laboratories and Ralph N. Harmon of Westinghouse Electric Corporation. . . . The role of the television receiver manufacturer took on added importance during sessions of the FCC's hearing on the utilization of the uhf for television broadcasting as lines were drawn between those favoring the **mixing of vhf and uhf assignments** and those opposing this mixture in TV allocations. The issue began to raise its head as the Commission questioned T. T. Goldsmith, Jr., of the Allen B. DuMont Laboratories, Inc. Dr. Goldsmith emphasized his company's desire to minimize the intermixture of uhf and vhf television channel assignments. Controversy over the proposal continued as other witnesses professed favor for "admixture" in order to create a "stimulus" for TV manufacturers to produce "all channel" or "universal" TV receivers. Commissioner Jones expressed the fear that manufacturers would give uhf the "FM treatment" if it were allocated chiefly to the rural areas. . . . An RCA-NBC witness gave a detailed technical description of uhf transmission and reception in the Bridgeport-Stratford, Conn., area. Raymond F. Guy, manager of radio and allocations engineering of the National Broadcasting Company, described RCA's uhf experiments.

[1] The data on which these NOTES are based were selected, by permission, from *Industry Reports*, issues of October 6, October 13, October 20, October 27, November 3, and November 10, published by the Radio-Television Manufacturers Association, whose helpful attitude is gladly acknowledged.

FCC Actions

The FCC by memorandum opinion and order has established a policy affecting 60-**kc adjacent channel frequency assignment** for miscellaneous common carriers licensed **in the Domestic Public Land Mobile Radio Services.** The change from 120-kc separation had been proposed by the FCC in a notice of proposed rule making issued May 12, 1950. However, the FCC has been allowing a 120-kc separation in the same geographical area since equipments capable of satisfactory adjacent channel operation were not available to these carriers. Improved equipments "are now available," the Commission stated. After the effective date of November 30, 1950, authorizations for the installation of new or additional equipments will not be authorized if the equipments are not capable of satisfactory service operation on an adjacent channel basis, except where it may be shown that the service requirements of a particular area may not reasonably be expected to require the use of adjacent channels in the foreseeable future. Where outstanding authorizations cover equipments not yet purchased or installed, permittees or licensees would be well advised to order and install appropriate new equipments, if that is possible. . . . The FCC has issued a Notice of Proposed Rule Making with the view of relaxing its Rules Governing the **Citizens Radio Service.** The revised rules would allow the utilization of part of the 2.28 Mc from which class-B stations are presently excluded, so as to provide for a tolerance of 0.5 per cent in lieu of the existing 0.4 per cent.

Radio and Television News Abroad

Operation of **Cuba's first television station** commenced on October 5 under the auspices of the Union Radio, Havana, Cuba. . . . **The English channel** has been **crossed by television.** The program originated in Calais, France, and was broadcast in England on BBC Television.

GE Proposes New Method For Improved TV Pictures

The General Electric Co. has submitted a plan for a new "high definition" system of monochrome TV to the FCC. The new method, according to the announcement of W. R. G. Baker, is based on different principles than horizontal interlace but which provides "appreciable improvement" in picture definition. The new system, GE told the FCC, was designed by Robert B. Dome, engineering consultant at Syracuse, N. Y. Mr. Dome originated GE's new color television which was submitted to the FCC last summer and which "will undergo extensive laboratory tests in the near future." Receivers designed to utilize the new method, according to the GE statement, will provide 50 per cent increase in horizontal detail. The new method is "compatible" in that present sets are capable, without alteration. of receiving signals of present black-and-white stations.

The GE proposal was submitted in connection with Section 7 (horizontal interlace possibilities, *RTMA Industry Report*, vol. 6, no. 40) of the FCC's notice of hearing relating to utilization of the uhf. At that time the FCC said it was of the opinion that "horizontal interlace gives promise of being an important development in television broadcasting." Interested persons were urged to conduct tests to determine whether horizontal color television which was submitted to the FCC last summer and which "will undergo extensive laboratory tests in the near future."

Receivers designed to utilize the new method, according to the GE statement, will provide 50 per cent increase in horizontal detail. The new method is compatible in that present sets are capable, without alteration, of receiving signals of present black-and-white stations.

The GE proposal was submitted in connection with Section 7 (horizontal interlace possibilities, *RTMA Industry Report*, vol. 6, no. 40) of the FCC's notice of hearing relating to utilization of the uhf. At that time the FCC said it was of the opinion that "horizontal interlace gives promise of being an important development in television broadcasting." Interested persons were urged to conduct tests to determine whether horizontal interlace can be utilized in the present monochrome system.

Sampling would not be utilized by the new system which "provides for the sending and receiving of fine detail and super-fine detail alternately," GE stated.

Advantages of the proposed system were listed as follows:

(1) All precision equipment is localized at the transmitter so that the receiver can be relatively low in cost, reliable in operation, easy to adjust and maintain, and simple in construction. It is estimated that only four more tubes are required than in present-day monochrome receivers.

(2) The receiver is compatible with present monchrome standards, using the same field, frame, and line rates.

(3) The picture has excellent texture in the sense that no dot structure should be visible since no sampling is employed. The texture thus is quite similar to present-day monochrome pictures.

Sections*

Chairman	Section	Secretary
. S. Hill Box 298 Hudson, Ohio	AKRON (4)	R. L. Burtner Goodyear Aircraft Corp. 1210 Massillon Rd. Akron Ohio
V. J. Miller 074 Springdale Rd. Atlanta 6, Ga.	ATLANTA (6) Nov. 17–Dec. 15	D. B. Buchanan 5130 Powers Ferry Rd. Atlanta Ga.
. W. Hammond Alabama Ct. Towson 4, Md.	BALTIMORE (3)	C. E. McClellan Westinghouse Elec. Corp. 2519 Wilkins Ave. Baltimore 3, Md.
Eric D. Coburn P. O. Box 1527 Beaumont, Texas	BEAUMONT-PORT ARTHUR (6)	C. B. Trevey 2555 Pierce St. Beaumont, Texas
W. H. Hauser WBZ, WBZ-FM, WBZ-TV 1170 Soldiers Field Rd. Boston Mass.	BOSTON (1)	Frank D. Lewis General Radio Co. 275 Mass. Ave. Cambridge, Mass.
M. J. Kobilsky FabricaArgentinaDeProd. Vedia Y AR. Forest Buenos Aires, Argentina, S.A.	BUENOS AIRES (1)	H. T. Magno Mirasol 451 Buenos Aires, Argentina S.A.
K. R. Wendt Colonial Radio Corp. 1280 Main St. Buffalo 9, N. Y.	BUFFALO-NIAGARA (4) Nov. 15–Dec. 20	H. G. Kuhn 211 Frankhauser Rd. R.F.D. #3 Williamsville, N. Y.
D. B. Harris 2003 Glenway Dr. Cedar Rapids, Iowa	CEDAR RAPIDS (5)	J. L. Hollis 2900 E. Ave. N.E. Cedar Rapids, Iowa
Kipling Adams General Radio Co., 920 S. Michigan Ave. Chicago 5, Ill.	CHICAGO (5)	R. M. Krueger Am. Phenolic Corp. 1830 S. 54 Ave. Chicago 50, Ill.
A. B. Bereskin 452 Riddle Rd. Cincinnati 20, Ohio	CINCINNATI (5)	W. B. Shirk 6342 Hamilton Ave. Cincinnati 24, Ohio
J. B. Epperson Scripps-Howard Radio Box 228 Berea, Ohio	CLEVELAND (4) Nov. 23–Dec. 28	J. L. Hunter John Carroll Univ. Cleveland Ohio
G. E. Mueller Elec. Eng. Dept. Ohio State Un. Columbus, Ohio	COLUMBUS (4) Nov. 10–Dec. 8	W. F. Rite 95 E. Ninth Ave. Columbus, Ohio
Herbert L. Krauss Dunham Laboratory Yale University New Haven, Conn.	CONNECTICUT VALLEY (1)	John Merrill 16 Granada Terr. New London, Conn.
J. H. Homsy 6022 Vanderbilt Dallas, Texas	DALLAS-FORT WORTH (6)	R. A. Arnett Box 7166 Dallas, Texas
C. J. Marshall 2412 Rosewood Ave. Dayton 9, Ohio	DAYTON (5)	R. W. Ittelson 551 Daytona Pkwy. Dayton, Ohio
W. E. Clyne 4255 Marshall St. Wheatridge, Colo.	DENVER (5)	R. E. Swanson 1777 Kipling St. Denver 15 Colo.
G. A. Richardson Dept. of Elec. Eng. Iowa State College Ames, Iowa	DES MOINES-AMES (5)	A. D. Parrott Northwestern Bell Tel. Co. 604—Ninth St. Des Moines 9, Iowa
N. C. Fisk 3005 W. Chicago Ave. Detroit 6, Mich.	DETROIT (4) Nov. 17–Dec. 15	P. L. Gundy 55 W. Canfield Ave. Detroit, Mich.
W. J. Peterson 322 E. Allegheny Ave. Emporium Pa.	EMPORIUM (4)	W. R. Rolf 364 E. Fifth St. Emporium, Pa.
R. E. Moe Eng. Dept. General Elec. Co. Owensboro, Ky.	EVANSVILLE-OWENSBORO	H. W. A. Chalberg 316 E. Ninth St. Owensboro Ky.
B. H. Ratts 2506 Terrace Rd. Ft. Wayne 3, Ind.	FORT WAYNE (5)	R. B. Jones 4322 Arlington Ave. Ft. Wayne, Ind.
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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February, 1949, through January, 1950, may be obtained for 2s. 8d postage included from the *Wireless Engineer*, Dorset House, Stamford St., London S. E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

ACOUSTICS AND AUDIO FREQUENCIES

534.152 2955
The Visualization of Vibratory Phenomena within a Resonant Tube—C. Chartier, J. Bourot, and J. Noël. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 2269–2270; June 26, 1950. The motions of powder particles within a resonating tube are recorded photographically; observation of changes produced by variation of experimental conditions is facilitated. Formulas are derived giving the deviation of the phase and longitudinal displacement of the powder particles from the values for the gas within the tube. These deviations vary in the same sense as the frequency used and as the size of the particles; faithful results are obtainable with ordinary Al powder at 1,000 cps. Photographs are reproduced for a tube giving a 32 cps note, taken with (*a*) continuous and (*b*) interrupted illumination.

534.232:534.321.9 2956
Experiments with the Hartmann Acoustic Generator—L. E. Savory. (*Engineering* (London), vol. 170, pp. 99–100 and 136–138; August 4 and 11, 1950.) The air-jet generator has been known for many years but has not been used extensively in industry for generating ultrasonic vibrations on account of its hypersensitivity in relation to changes of air pressure and of certain interelement spacings. Modifications intended to reduce this instability are described; these include side pads or a short cylinder partially or wholly surrounding the jet somewhere between nozzle and resonator, and a rod inserted along the jet axis. Results are tabulated and discussed.

534.321.9:537.228.2 2957
The Generation of Ultrasonic Vibrations by Electrostriction—H. H. Rust. (*Z. Angew. Phys.*, vol. 2, pp. 293–294; July, 1950.) Falkenhagen's method (1943 of 1949) is discussed and compared with much earlier work by Langevin.

534.6:621.395.623.7 2958
Practical Applications of Acoustic Measurements on Receivers—A. Moles. (*TSF pour Tous*, vol. 26, pp. 224–229; June, 1950.) Results of tests in an anechoic chamber are shown by records of the low-frequency, over-all and loudspeaker response curves of a receiver; the effect of varying the receiver selectivity and of using different baffles for the loudspeaker are also shown. The back of a receiver with built-in loudspeaker should be perforated so as to introduce a certain amount of acoustic impedance. Placing a receiver near a partition, which acts as a baffle, results in an appreciable improvement of the bass response; otherwise the position of the receiver has little effect.

534.75:534.321.1 2959
Upper Limit of Frequency for Human Hearing—R. J. Pumphrey. (*Nature* (London), vol. 166, p. 571; September 30, 1950.) Experimental evidence that the sensory elements at the basal end of the cochlea are competent to respond to sounds of frequency up to 100 kc, and that the failure of the normal ear to respond to air-borne sounds of frequency above 20 kc is due wholly to the failure of the middle ear to transmit such frequencies. The apparent pitch associated with high-frequency sounds was that of the highest tone audible by air conduction, no increase of pitch being observed above about 15 kc.

534.78:534.843.5 2960
General Theory of Intelligibility within Rooms—T. Korn. (*Ann. Télécommun.*, vol. 5, pp. 316–320; August, September, 1950.) Reverberation time is not the only important factor governing intelligibility; the effect of the direct sound is another factor of major importance. A method is given for calculating the articulation index anywhere in a room from the values of reverberation time, total absorption and distance from the sound source. The calculated results agree with classical experimental data. The direct sound can be reinforced by architectural or electronic devices.

534.84:061.3 2961
Symposium on Architectural Acoustics, Marseilles, 11th–17th April 1950—P. Chavasse. (*Ann. Télécommun.*, vol. 5, pp. 255–258; July, 1950) A summarized report of the proceedings, with a list of the papers presented.

534.844.1 2962
Measurement of Reverberation Times of Studies of the Radiodiffusion Française—J. Pujolle and J. Boisard. (*Ann. Télécommun.*, vol. 5, pp. 307–315; August, September, 1950.) Apparatus and method are described and degree of accuracy assessed.

534.844.3 2963
A Contribution to the Study of Reverberation—A. C. Raes. (*Ann. Télécommun.*, vol. 5, pp. 259–265; July, 1950.) Paper presented at the International Symposium on Architectural Acoustics, Marseilles, 1950. The relation between the reverberation time of an auditorium and the quality of the sound heard is discussed in the light of R. B. Watson's conception of reverberation modulation. Oscillographically recorded results of experiments made with pure-tone sound sources indicate that the frequency of such modulation may attain the order of 250 cps with associated amplitude reductions of 10–18 db.

621.3.018.78†:621.395.613.3 2964
The Recording of Nonlinear Distortion—A. Bressi and G. G. Sacerdote. (*Alta Frequenza*, vol. 19, pp. 86–92; April, 1950. In Italian, with English, French and German summaries.) The determination of nonlinear distortion is of particular importance in relation to carbon microphones. A method is described for automatically recording the magnitude of this distortion as a function of the applied acoustic pressure at any frequency.

621.395.61/.62 2965
The Constants of Magnetostrictive, Electrostrictive and Piezoelectric Electroacoustic Transducers—U. John. (*Arch. elekt. Übertragung*, vol. 4, pp. 139–145; April, 1950.) From a mathematical treatment of the dynamics of the magnetostriction process an expression is derived for the conversion factor; this is independent of the measurement system used and takes account of the permeability and elastic modulus, and also the magnetostriction factor. A table shows the four different forms of the fundamental dynamic equation, the expressions for their constants and the electromagnetic equivalents. Each group of associated constants can be interpreted as a matrix of a linear quadripole and can be referred to the corresponding groups for the electrostrictive and piezoelectric processes.

621.395.61/.62 2966
The Electrical and Acoustical Impedances and Equivalent Circuits of Electroacoustic Transducers—F. A. Fischer. (*Arch. elekt. Übertragung*, vol. 4, pp. 189–195; May, 1950.) Three basic types of transducer are considered: (*a*) electrically excited, (*b*) magnetically excited, (*c*) electrodynamic. The piezoelectric type is to be discussed later. An expression for complex impedance is derived in each case and the corresponding electrical and mechanical equivalent circuits are deduced.

621.395.61 2967
Directional Microphones with Phase-Shifting Elements—H. Grosskopf. (*Fernmeldetech. Z.*, vol. 3, pp. 248–253; July, 1950.)

The problem of designing a microphone with a directivity independent of frequency is considered generally. Artificial phase-shifting elements are introduced to produce the required directional characteristics. These elements should produce phase shift proportional to frequency; an acoustic by-pass is simplest in theory but is too cumbrous in practice. Elements comprising mechanical impedances are therefore preferred; these are studied by means of mechanical networks. Ribbon and capacitor microphones can be adapted in accordance with the principles given; continuously variable directivity can be obtained by combining in appropriate phase the outputs from two microphones having cardioid diagrams.

621.395.61:621.395.828 2968
Methods of Preventing Acoustical Feed-back—W. Bürck. (*Frequenz*, vol. 4, pp. 161–166; July, 1950.) The positioning of microphone and loudspeaker and the use of compensating circuits are reviewed.

621.395.614 2969
Electroacoustic Apparatus with Rochelle-Salt Crystals—M. Gosewinkel. (*Frequenz*, vol. 4, pp. 142–145; June, 1950.) The effect of temperature change on the crystal characteristics is discussed. Loading methods to compensate for this are not generally satisfactory. In crystal generators the total adverse effect is small. In crystal receivers the change in sensitivity with temperature above 23°C is appreciable and renders them unsuitable where constancy of operation is required.

621.395.625.2 2970
Recent Trends in Disk Recording—P. Gilotaux. (*Onde Elec.*, vol. 30, pp. 301–308; July, 1950.) A discussion of factors limiting fidelity of reproduction and of the special problems of fine-groove recording.

ANTENNAS AND TRANSMISSION LINES

621.3.09:621.315.2+621.392.26†] +621.317.3.029.6 2971
Lines and Circuit Elements for Microwave Measurements—Schäfer and Honerjäger. (*See* 3088.)

621.392.018.8 2972
The Attenuation of Transient or Periodic Electromagnetic Waves Due to Effects in the Dielectric Medium Surrounding Transmission Lines—R. Pélissier. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 2272–2274; June 26, 1950.) An analysis is given of the manner in which the wavefront corresponding to an applied step voltage is deformed by leakage and corona as it progresses along a line. Corona is doubly effective, firstly by increasing the leakage current, and secondly by producing a space charge leading to an apparent increase in the dielectric constant. Its deforming action is greater than that due to skin effect but less than that due to ground return.

621.392.26†:621.3.09 2973
Principles and Applications of Waveguide Transmission—G. C. Southworth. (*Bell Sys. Tech. Jour.*, vol. 29, pp. 295–342; July, 1950.) Excerpts from a book of the same title, in preparation. Chapter 6, which gives a descriptive account of transmission in waveguides, is quoted in full.

621.396.67+621.396.11 2974
Symposium on Antennas and Propagation, San Diego, California, April 3–4, 1950—(Proc. I.R.E., vol. 38, pp. 958–962; August, 1950.) Summaries are given of 38 papers presented at the conference.

621.396.67 2975
A 600-Ohm Multiple-Wire Delta Antenna for Ionosphere Studies—H. N. Cones, H. V. Cottony, and J. M. Watts. (*Bur. Stand. Jour. Res.*, vol. 44, pp. 475–488; May, 1950.) An account of the design and performance of a multiple-wire delta antenna for vertical-incidence ionospheric recorders. The variation of the terminal impedance over the working range (1–25 Mc) is displayed graphically. Radiation pattern measurements using scale-model technique are given and comparative records using the delta antenna and an earlier double-w type are reproduced.

621.396.67 2976
Current Distributions on Transmitting and Receiving Antennas—T. Morita. (Proc. I.R.E., vol. 38, pp. 898–904; August, 1950.) A method is described for the direct measurement of the amplitude and phase of current and charge distribution along cylindrical antennas. For the transmitting antenna, the distribution measurement is continued into the coaxial feeder, so that a picture of the complete system is obtained. For the receiving antenna, the distribution is found to vary as a function of the terminating load, as predicted by theory.

621.396.671 2977
A Rigorous Calculation of Tubular Dipole Aerials—H. Zuhrt. (*Frequenz*, vol. 4, pp. 135–141 and 178–189; June and July, 1950.) First published in 1944. Cylindrical dipoles of diameter small, but not negligible, compared with λ are considered. A rigorous theory is developed from a solution of the wave equation satisfying rigorous boundary conditions, an image method being used. The theory is applied to the calculation of the input impedance of a relatively thick symmetrical transmitting antenna; results are compared with measurements. The theory is then extended to certain other nonsymmetrical receiving and transmitting dipole arrangements and their impedances are calculated.

621.396.671 2978
On the Theory of the Straight Aerial—J. Grosskopf. (*Arch. elekt. Übertragung*, vol. 4, pp. 175–180; May, 1950.) An approximate formula for the radiation impedance is obtained by applying the fundamental principles of the Hallén theory (2763 of 1939) based on Maxwell's equations, but avoiding the complicated iterative process for determining the current distribution in the antenna. Calculations from the approximate formula give results in good agreement with experimental values and of accuracy as good as is given by two terms of Schelkunoff's development.

621.396.677 2979
Electromagnetic Field of the Conical Horn—M. G. Schorr and F. J. Beck, Jr. (*Jour. Appl. Phys.*, vol. 21, pp. 795–801; August, 1950.) The radiation from a conical horn is calculated by the assumed-field method. The calculated results agree well with values found experimentally for cones of half-angle up to about 25°. For greater angles the approximations used are inadequate and result in poor agreement with experiment. The theoretical anomalies are discussed.

621.396.677:621.392.26† 2980
The Radiation from a Transverse Rectangular Slot in a Circular Cylinder—S. Silver and W. K. Saunders. (*Jour. Appl. Phys.*, vol. 21, pp. 745–749; August, 1950.) Results obtained previously for slots of arbitrary shape (1588 of August) are applied to the special case of a transverse rectangular slot. The principal transverse-plane pattern for such a slot in which the excitation has only a circumferential tangential electric-field component is the same as the pattern generated by an infinite axial slot with the same circumferential excitation. Theoretical and experimental curves are given for the narrow slot of length λ/2.

621.396.677.012 2981
Survey of Directional Characteristic Curves and Surfaces—F. Bergtold. (*Frequenz*, vol. 4, pp. 114–117; May, 1950.) The general principles of directional radiation as applied to both em and sound waves are outlined and surfaces are discussed which are such that the radius vector for each point of the surface represents the intensity of the radiation in that particular direction from a source at the origin. The intersection of such a surface with any plane passing through the origin gives the radiation polar diagram for that plane. The advantages of a logarithmic scale for the radiation intensity are pointed out. Two abacs based on rectangular-co-ordinate diagrams enable the directional characteristics of most types of array to be obtained.

CIRCUITS AND CIRCUIT ELEMENTS

549.514.51.001.8 2982
Applications of Quartz Crystals—Villem. (*See* 3110.)

621.3.012.8 2983
Construction and Critical Review of Circuit Equivalents for Linear A.C. Elements—E. Samal. (*Frequenz*, vol. 4, pp. 81–88, 117–125, 149–151; April–June 1950.) The ideal circuit element is one that can be considered to possess only one electrical property (e.g. resistance), the value of which should be independent of frequency, time, temperature, etc. The accurate representation of actual components by means of such ideal elements is considered in detail with numerous examples, using circle diagrams, oscillatory circuits and transducing networks. The effects of eddy currents, ferromagnetic and es phenomena are discussed and also the modifications necessary to the equivalent circuit when such effects are present. When the particular element in question can no longer be represented accurately by simple diagrammatic methods, recourse must be had to the solution of the appropriate Maxwell equations.

621.3.013.78† 2984
Screening in Electrical Communication Apparatus—K. F. Weinert. (*Funk u. Ton*, vol. 4, pp. 358–368; July, 1950.) Study of the effect of screening in and between coupled quadripoles when the coupling is (*a*) galvanic, the circuits having a common resistor, (*b*) capacitive, (*c*) inductive, or (*d*) mixed. Expressions giving the damping due to different types of screen are tabulated. Incorrect earthing arrangements can lead to considerable reduction in the screening effect.

621.3.015.33
621.3.015.7† 2985
Pulses—Please note that, in this and subsequent issues, the UDC number used will be 621.3.015.7†, instead of 621.3.015.33 used hitherto.

621.3.016.352 2986
Stability Criteria—E. Görk. (*Arch. elekt. Übertragung*, vol. 4, pp. 89–96; March, 1950.) Most methods of determining stability conditions are based on the principle of the number of zeros and poles of the function $Z(p)$. In the method described this principle is applied directly. The expression, based on Cauchy's theorem, for the number of zeros when the function is wholly rational or transcendental is integrated in the appropriate half plane to separate the roots; two examples are calculated. Methods of establishing limits for the roots in determining the degree of stability are shown for three values of damping. The Sturm division method and the use of Hurwitz determinants are discussed.

621.3.016.352:621.3.012 2987
Stability Testing by means of Closed and Open Circle-Diagrams—F. Strecker. (*Arch. elekt. Übertragung*, vol. 4, pp. 199–206; June, 1950.) Detailed discussion of the physical

eaning and practical application of circle-iagram stability criteria.

21.314.22.015.7† 2988

Pulse Transformers—H. G. Bruijning. (*Commun. News*, vol. 10, pp. 126–131; April, 950.) The requirements which pulse transormers must satisfy are discussed, a design nethod is described and the advantages reulting from the use of ferroxcube as a core maerial are considered.

21.314.23 2989

The Use of U/I Cores in Transformers—A. Nonnenmacher. (*Funk u. Ton*, vol. 4, pp. 85–395; August, 1950.) Stampings for standrd U/I cores can be obtained by halving E/I aminations, and hence can be produced withut waste. Leakage coefficient is independent of ore size and is only half that with E laminaions. The sensitivity to external magnetic elds is investigated, with laminations stacked alternately the U core is better in this respect than the shell. This advantage is lost when an airgap is used. The apparent width of airgap with the yoke removed is plotted as a function of the measured airgap for U cores and for the 59 core.

21.314.3† 536.58 2990

A Direct-Voltage Amplifier as Power Amplifier and Temperature Regulator—W. Jellinghaus (*Z. angew. Phys.*, vol. 2, pp. 254–261; une, 1950.) Description of a fast-acting reguator comprising a magnetic amplifier controlled by a thermocouple and followed by a ircuit in which the amplified voltage is rectiied and used to bias negatively a tube whose mode circuits includes premagnetization windngs for a second magnetic amplifier supplying he heating element.

621.384.611.2:621.316.726 2991

Frequency Control for the Bevatron Radio-Frequency Voltage—Riedel. (*See* 3116.)

621.316.89:621.385.831 2992

The Negative-Feedback Valve as a Controllable Effective Resistance—A. Klemt. (*Funk u. Ton*, vol. 4, pp. 293–297; June, 1950.) When the negative feedback is such that the alternating voltage at the grid is equal to that at the anode, the ac resistance of an amplifier ube is equal to the reciprocal of its slope. Applications of the principle include conductance measurements in parallel resonant circuits, use as a resistor in a phase-shift oscillator, and noise limitation in FM receivers. Basic circuits are shown.

621.318.4:621–05 2993

Universal Coil Winding—E. Watkinson. (Proc. I.R.E. (Australia), vol. 11, pp. 179–186; July, 1950.) A typical coil-winding machine and the winding patterns derived from it are described. The factors which influence the choice of the coil-winding gear ratio are explained and analysed. The results are presented in a form suitable for slide-rule computation.

621.319.55:517.93 2994

A Note on a Generalized van der Pol Equation—W. Nijenhuis. (*Philips Res. Rep.*, vol. 4, pp. 401–406; December, 1949.) The asymmetrical van der Pol equation, $\ddot{v} - \epsilon(1 - 2\beta v - v^2)\dot{v} + v = 0$, is shown to be equivalent to the classical van der Pol equation ($\beta = 0$) with a constant on the right-hand side. For large values of ϵ and β a simple approximation for the limit cycle is found from which it follows that the ratio of the maximum positive and negative values of $v(t)$ cannot exceed 3:1.

621.319.55:517.93 2995

On the Theory of Relaxation Oscillations—W. Taeger. (*Funk u. Ton*, vol. 4, pp. 298–307 and 341–346; June and July, 1950.) The second-order equation applicable to an oscillatory circuit with periodically varying resistance is reduced to Mathieu's differential form and solutions are obtained for values of damping coefficient (*a*) <1 and (*b*) >1. The theory is then applied to the case of a neon lamp relaxation circuit and formulas for the frequency and amplitude of oscillations are derived.

621.385.2/.4:621.315.59 2996

Transistor and Transistron—Adam. (*See* 3210.)

621.392.52 2997

Tchebycheff Functions—W. Klein. (*Arch. Elektrotech.*, vol. 39, pp. 647–657; 1950.) The Tchebycheff functions of real arguments used in filter theory are presented in five different ways. An extension to complex arguments is made to take into account losses in the circuit elements. A capacitively coupled band-pass filter is discussed as an example of the application of functions with complex arguments.

621.392.52 2998

Use of a Template Method for the Design of Filters—A. Fromageot and M. A. Lalande. (*Ann. Télécommun.*, vol. 5, pp. 277–290; August, September, 1950.) The methods described are based on that of Rumpelt (729 of 1943) and apply to lattice filters with opposite arms consisting of identical two-poles. The critical frequencies of the two-poles in the pass band depend on the image-transfer coefficient, in the stop band they depend on the image impedance. The image-transfer coefficient is defined by the required attenuation the image impedance by its mean value and the permissible variation within the pass band. Some examples are given using piezoelectric crystals as circuit elements. See also 842 of May.

621.392.52:621.3.012.8 2999

The Relation between Recurrent Networks and Radio Filter Circuits—W. Klein. (*Funk u. Ton*, vol. 4, pp. 273–281; June, 1950.) The theory of the recurrent network is discussed by analysing a basic filter consisting of a branched circuit in which series resistance and shunt conductance are equivalent, the former comprising ohmic resistance and pure reactance in series, the latter a parallel combination of ohmic conductance and pure susceptance. On the basis of equivalent circuits having the same transmission factor, a short-circuit tuned, reactively coupled radio filter is developed from the basic network. Bandwidth here is necessarily restricted since all series impedances in the filter are considered as frequency-independent reactances. For transmission over a relatively wide band the basic type of network is required.

621.392.52:621.396.662 3000

A New High-Frequency Filter with Band-Pass-Filter Properties—F. Benz. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, vol. 4, pp. 70–75; May, June 1950.) The filter described comprises two parallel-resonant circuits detuned from each other by a small amount. The input voltage is shared between them so that a differential output is obtained across the two circuits in series. Circle diagrams, frequency-response curves and phase characteristics are plotted, and performance compared with that of the common type of filter circuit. Applications are discussed; in general the arrangement can be substituted for ordinary resonant-circuit couplings. Symmetry of the frequency-response curve is inherent in the design.

621.396.611.1 3001

A Resonant Circuit with a Time-Variant Resistive Element—P. Bura and D. M. Tombs. (*Nature* (London), vol. 166, pp. 483–484; Septemper 16, 1950.) Two phenomena have been observed: (*a*) Multiple resonance; in a circuit tuned to f_0 (6 kc) the resistive element was varied at f_R (50 cps) and the voltage across the main capacitance showed peaks at $f_0 \pm nf_R$, where $n = 1, 2, 3 \cdots$; (*b*) Parametrically excited oscillations; when f_R was about $2f_0$ and the effective resistance became negative during part of the cycle, oscillations at $f_R/2$ occurred over a range of frequencies near f_0. This range is limited by sharp cutoff of the self-oscillation, which suggests the use of the device as a filter, the bandwidth being readily controlled. Calculations, based on a theory briefly indicated, are in good agreement with experimental results.

621.396.615 3002

Electron Transit Time in Negative-Grid Oscillator—S. K. Chatterjee and B. V. Sreekantan (*Indian Jour. Phys.*, vol. 23, pp. 119–130, March, 1949.) Expressions are derived for cathode/grid and grid/anode electron transit times for a triode operating as a class-C oscillator, taking account of the ac voltages on anode and grid and also of the grid bias. Numerical results are given for various tubes

621.396.615 3003

The Limiting Frequency of an Oscillator Triode—K. Rodenhuis. (*Philips Res. Rep.*, vol. 5, pp. 46–77; February, 1950.) The triode is considered as a quadripole. General conditions, which the quadripole coefficients must satisfy in order that the quadripole may be able to deliver power to an external circuit, are derived; these are identified with the oscillation conditions for a triode. The relations between the quadripole coefficients of a triode on the one hand, and the properties of the electron current, the series resistance in the electrode leads, the resistance of the emissive coating, and the dielectric losses, on the other hand, are established and discussed. When the theory is applied to the usw oscillator triode Type EC81, the calculated limiting frequency is 15 per cent above the observed value. This discrepancy is attributed to approximations in the theory. The experimental determination of the limiting frequency is briefly described

621.396.615.14+621.396.645]:
621.385.029.63/.64 3004

Amplification and Generation of Microwave Oscillations with Travelling-Wave Valves—H. Schnitger (*Funk u. Ton*, vol. 4, pp. 347–354; July, 1950.)

621.396.615.141.2 3005

Development of the Turbator for Radio-Relay Equipment—F. Lüdi. (*Brown Boveri Rev.*, vol. 36, pp. 405–409; December, 1949.)

621.396.615.17 3006

Discussion of an Asymmetrical Multivibrator Circuit—J. Blok and C. C. Jonker (*Physica*, 's *Grav.*, vol. 16, pp. 381–390; April, 1950. In English.) The properties of asymmetrical multivibrator circuits are discussed by the method used by Andronow (1648 of August) for the symmetrical multivibrator. The method is extended to take account of grid currents. The operation of such circuits as discriminators is examined and the resolving time is estimated.

621.396.619:538.632 3007

Possible Modulators Based on the Hall Effect—Y. Rocard. (*Rev. Sci.* (Paris), vol. 87, p. 212; October, December, 1949.) The circuit described is intended for amplifying weak dc. The Hall effect is produced in a thin Ge plate by means of an auxiliary alternating current and an alternating magnetic field whose strength is controlled by the dc to be amplified. Voltage amplification of some thousandfold is obtainable.

621.396.645 3008

Improving Differential-Amplifier Rejection Ratios—R. McFee. (*Rev. Sci. Instr.*, vol. 21, pp. 770–771; August, 1950.) Modifications are suggested to two standard differential-amplifier circuits which effect fifty-fold improvements in rejection ratios. Where the input potentials are applied to the grids of a pair of similar triodes having a common cathode resistor, R_K, the modification consists in connecting a resistor equal in value to R_K across the output triode, from cathode to anode. In the symmetrical amplifier in which the output is taken from both anodes of the pair of triodes, the modification consists in connecting equal resistors across each of the two tubes, from anode to the high-potential end of the cathode resistor, the output being taken from suitable symmetrical tapping points.

621.396.645 3009

Stagger-Tuned Intermediate-Frequency Amplifier Design—A. B. Thomas. (*Jour. I.R.E.* (Australia), vol. 22, pp. 141–148; June, 1950). Charts are given for facilitating selection of the most suitable type of interstage coupling for any particular purpose, and for deciding the frequency and damping of each stage in a stagger-tuned unit. A method is given for extracting the response curve of any stage and a simple lining-up procedure is described which can be applied to staggered triples. Reference is made to a radar unit recently constructed, with a mid-band gain of 100 db and a bandwidth of 12.5 Mc centered at 70 Mc. This receiver proved to be stable and easy to handle and duplicate.

621.396.645:621.316.078 3010

Automatic Stabilization of Amplifier Gain—A. W. Keen. (*Jour. Brit. I.R.E.*, vol. 10, pp. 198–207; June, 1950.) "The conventional method of measuring the voltage gain of an amplifier by equating it with the loss introduced by a calibrated attenuator is adapted to provide a continuous and automatic method of accurately defining and maintaining constant amplifier gain. The resultant arrangements are all forms of negative feedback systems and are characterized by high loop gain and the use of a precision attenuator in the β network. Two types are distinguished according to whether the error signal representing the discrepancy between the input and attenuated output signals is made self-cancelling by being fed back into the amplifier as in conventional feedback amplifiers, or used to control the gain of one or more of the amplifying tubes, as in agc systems. Appropriate circuit technique is indicated and suitable applications of the method are briefly described."

621.396.645.37 3011

Selective Amplification by Means of RC Feedback Networks—M. Picchi. (*Alta Frequenza*, vol. 19, pp. 59–85; April, 1950. In Italian, with English, French and German summaries.) General expressions are derived for the resonance frequency and resonance coefficient of a selective amplifier. Three particular amplifier circuits are discussed in which the Scott parallel-T quadripole, the Wien-Robinson bridge, and certain phase-shift quadripoles are used as feedback networks.

621.396.645.371 3012

A Note on the Maximum Feedback Obtainable in an Amplifier of the Cathode-Feedback Type—J. te Winkel. (*Philips Res. Rep.*, vol. 5, pp. 1–5; February, 1950.) The total feedback on the last tube in a feedback amplifier is calculated for the case where the feedback voltage is derived from the cathode load impedance of the last tube. The total feedback is invariably less than the product of individual tube feedback and feedback around the main loop, and has an upper limit dependent only on frequency and on the ratio of transconductance to input capacitance of the last tube.

621.396.645.371:523.165 3013

Linear Preamplifier for Driving a Long Coaxial Cable—L. G. Lewis, J. G. Robinson, and J. Toll. (*Rev. Sci. Instr.*, vol. 21, pp. 593–596; July, 1950.) Linearity and long-term stability under severe weather conditions have been achieved by using negative feedback. The particular application is to an exposed particle-detector system linked by a 100-ft length of cable to the main amplifier.

621.396.665:621.396.645.371 3014

Correction and Clarification of "Automatic Volume Control as a Feedback Problem"—B. M. Oliver. (PROC. I.R.E., vol. 38, p. 904; August, 1950.) Correction to paper abstracted in 2501 of 1948.

621.397.645 3015

Television Intermediate-Frequency Amplifiers—Coenraets. (*See* 3191.)

621.392.52 3016

Siebschaltungen mit Schwingkristallen (Crystal Filter Circuits) [Book Review]—W. Herzog. Publishers: Dieterichsche Verlagsbuchhandlung, Wiesbaden, 1949, 361 pp., 45 DM. (*Fernmeldetech. Z.*, vol. 3, pp. 256–257; July, 1950.) The author has brought together in this comprehensive and clearly presented work the material available on the subject. General laws and principles of filter design are given, and basic properties of crystals are explained by means of equivalent circuits. Examples are calculated, and sections are devoted to adjustment and measurement and temperature dependence of the filters. The text is arranged so that for designing any particular type of filter only the relevant section has to be consulted.

621.396.611.1+621.385].029.6 3017

Technique des ondes très courtes et ultra-courtes: Tome I—Circuits oscillants et tubes à vide [Book Review]—L. Liot. Publishers: Dunod, Paris, 1949, 253 pp., Fr. 780. (*Alta Frequenza*, vol. 19, pp. 110–111; April, 1950.) Essentially a practical book, useful both as an introduction to the subject and for reference.

GENERAL PHYSICS

530.112 3018

Matter, A Mode of Motion—R. V. L. Hartley. (*Bell Sys. Tech. Jour.*, vol. 29, pp. 350–368; July, 1950.) "Both the relativistic and wave mechanical properties of particles appear to be consistent with a picture in which particles are represented by localized oscillatory disturbances in a mechanical ether of the MacCullagh-Kelvin type. Gyrostatic forces impart to such a medium an elasticity to rotation, such that, for very small velocities, its approximate equations are identical with those of Maxwell for free space. The important results, however, follow from the inherent nonlinearity of the complete equations and the time dependence of the elasticity associated with finite displacements. These lead to reflections which permit of a wave of finite energy remaining localized. Because of the nonlinearity, the amplitude and energy of a stable mode, as well as the frequency, are determined by the constants of the medium. Such a stable mode is capable of translational motion and so is suitable to represent a particle. The mass assigned to it is derived from its energy by the relativity relation. While this mass is dimensionally the same as that of the medium, it is differently related to the energy and so need not conform to the classical laws which the latter is assumed to obey. Exchanges of energy between particles and between a particle and radiation involve frequency changes as in the quantum theory. The experimental detection of a uniform velocity relative to the medium is not to be expected. Besides providing a new approach to the problems of particle mechanics, the theory offers the prospect of incorporating the present pictures into a more comprehensive one, with a material reduction in the number and complexity of the independent assumptions." See also 3019 below.

530.112:535.312 3019

The Reflection of Diverging Waves by a Gyrostatic Medium—R. V. L. Hartley. (*Bell Sys. Tech. Jour.*, vol. 29, pp. 369–389; July, 1950.) "This paper furnishes the basis for a companion one [3018 above], which discusses the possibility of describing material particles as localized oscillatory disturbances in a mechanical medium. If a medium is to support such disturbances it must reflect a part of the energy of a diverging spherical wave. It is here shown that this property is possessed by a medium such as that proposed by Kelvin, in which the elastic forces are of gyrostatic origin. This is due to the fact that, for a small constant angular displacement of an element of this medium, the restoring torque, instead of being constant, decreases progressively with time."

535.8 3020

On the Design of Wide-Angle Schmidt Optical Systems—E. M. Wormser. (*Jour. Opt. Soc. Amer.*, vol. 40, pp. 412–415; July, 1950.) "A method of balancing the optical aberrations over a wide angular field of a Schmidt optical system is described."

537.122:538.691 3021

Electron Motion in Alternating Magnetic Fields—A. Kneschke. (*Arch. elekt. Übertragung*, vol. 4, pp. 165–172; May, 1950.) The differential equation expressing the motion is integrated completely. The geometric and kinematic characteristics of the electron paths are investigated and related to curves whose curvature is a periodic function of arc length.

Certain special cases of the general trajectory are analysed in detail and shown graphically.

537.226 3022

Equivalence of Temperature and Frequency in Dielectric Measurements—K. H. Stark. (*Nature* (London), vol. 166, p. 436; September, 1950.) The variation of the complex dielectric constant as a function of frequency is represented by a circular arc. Experiments with two widely different dielectrics yielded circular arcs also for variation of complex dielectric constant as a function of temperature. Equivalence of temperature and frequency for dielectrics with a wide distribution of relaxation times is hence established, but no theoretical explanation is known.

537.311.31 3023

The Bloch Integral Equation and Electrical Conductivity—P. Rhodes. (*Proc. Roy. Soc. A*, vol. 202, pp. 466–484; August 22, 1950.) An extension of the detailed development of the Bloch treatment of the temperature variation of conductivity to cover a wider temperature range. Theoretical and experimental results are in good general agreement.

537.311.33 3024

Thermal Equilibrium in Neutron-Irradiated Semiconductors—J. H. Crawford, Jr., and K. Lark-Horovitz. (*Phys. Rev.*, vol. 79, pp. 889–890; September 1, 1950.) Analysis of conductivity/irradiation curves of Ge semiconductors exposed to neutron flux shows the experimental minimum conductivity to be much lower than anticipated. Redetermination of the product of electron and hole densities from Hall-effect measurements gives a value of 3.6×10^{26} at 300°K, which is about one tenth of the value previously found and which yields a much better estimate of minimum conductivity. Measurements are made with the material in the pile, where photoeffects play an important part.

537.311.33:538.632 3025

Temperature Dependence of the Energy Gap in Germanium from Conductivity and Hall Data—V. A. Johnson and H. Y. Fan. (*Phys. Rev.*, vol. 79, p. 899; September, 1950.)

he data are for the temperature range 500°K 850°K. If the energy gap between the full ıd conduction bands is expressed by the rela-on $E_G = E_{G_0} + (\partial E_G/\partial T)T$, F_{G_0} is determined 0.73 ev and $\partial E_G/\partial T$ as constant at -1.1±0.1)×10^{-4} ev/°K, assuming $m_e m_h$ m_0^2, where m_e, m_h, m_0 are the effective elec-on and hole masses and free-electron mass spectively.

37.311.33:528.632 **3026**

On the Theory of the Hall Effect for the ase of Compound Conductors in an Alternat-g Electric Field—H. Welker. (*Onde Élec.*, ol. 30, pp. 309–316; July, 1950.) In the case onsidered, two types of moving charge carrier e present; large displacements of charge per-endicular to the primary current occur. These splacements have appreciable relaxation mes, whose magnitudes are examined by sub-ituting ac for dc primary current. The value the Hall voltage then depends on the fre-uency of the ac. Results are given for various ompound conductors, including electrolytes, emiconductors and polyvalent metals. Formu-s for determining particle concentrations and obilities are derived and the possibilities of xperimental verification are discussed.

37.525:534.01 **3027**

Effects of Plasma Boundaries in Plasma Oscillations—D. Bohm and E. P. Gross. *Phys. Rev.*, vol. 79, pp. 992–1001; September, 950.) An electron taking part in a travelling lasma oscillation will be reflected at a sheath f infinitesimal thickness with velocity ap-ropriate to the oscillation travelling in the everse direction, so that standing waves are uilt up without loss at the sheath. This ap-roach is extended to sheaths for which a finite me of penetration is necessary before reflec-on occurs, and also to the case of reflection at etallic electrodes. In both cases expressions re derived for the damping. For low-pressure ischarges, the damping resulting from im-erfect reflection from electrode sheaths may e comparable with collision damping, but the amping due to conducting electrodes is unim-ortant.

The excitation of the plasma by sharp eams is considered and expressions are de-ived for the energy transfer of a beam to waves f growing and of stationary amplitude. Beams hould only excite oscillations when a regular eometry exists. With irregular geometry, unching pulses are to be expected. A detailed nalysis of the bunching is given.

37.533.8 **3028**

Measurement of Secondary-Electron Emis-ion from Dielectric Surfaces—H. L. Heydt. *Rev. Sci. Instr.*, vol. 21, pp. 639–642; July, 950.) The method described involves bom-arding the metal-backed dielectric surface vith primary electrons so as to produce a econdary-emission ratio greater than unity nd hence a surface potential of known value.

38.221 **3029**

Departure from the Rayleigh Law of the Magnetization of a Ferromagnetic Material—I. D. Bush. (*Nature* (London), vol. 166, pp. 01–402; September 2, 1950.) An interim re-ort of experimental investigations.

39.233:535.215 **3030**

Photoadsorption of a Thin Layer of Barium—C. Biguenet. (*Le Vide*, vol. 5, pp. 31–836; July, September, 1950.) Experiments re described which confirm that light produces arge variations of the contact difference of otential between an oxide cathode and a ungsten-filament collector electrode; the ex-periments show also that the effect is not due to photoemission. A possible explanation is that he barium adsorbed on the tungsten filament s ionized by the light, causing a change in the electron affinity of this electrode.

548.0:537 **3031**

Scattering of Electrons in Crystals in the Presence of Large Electric Fields—J. Bardeen and W. Shockley. (*Phys. Rev.*, vol. 80, pp. 69–71; October 1, 1950.) "By the calculation of transitions between states appropriate to elec-trons moving in a large uniform electric field superimposed on a periodic crystal field, it is shown that the probabilities of scattering by lattice vibrations or imperfections are inde-pendent of the uniform field and are given by the usual expressions derived for zero field. This justified the procedure of treating acceleration by the field and scattering as independent processes."

548.0:537 **3032**

Deformation Potentials and Mobilities in Non-Polar Crystals—J. Bardeen and W. Shockley. (*Phys. Rev.*, vol. 80, pp. 72–80; October 1, 1950.) "The method of effective mass, extended to apply to gradual shifts in energy bands resulting from deformations of the crystal lattice, is used to estimate the inter-action between electrons of thermal energy and the acoustical modes of vibration. The mobili-ties of electrons and holes are thus related to the shifts of the conduction and valence-bond (filled) bands, respectively, associated with dilations of longitudinal waves. The theory is checked by comparison of the sum of the shifts of the conduction and valence-bond bands, as derived from the mobilities, with the shift of the energy gap with dilation. The latter is obtained independently for silicon, germanium and tellurium from one or more of the following: (1) the change in intrinsic conductivity with pressure, (2) the change in resistance of an *n-p* junction with pressure, and (3) the variation of intrinsic concentration with temperature and the thermal expansion coefficient. Higher mobilities of electrons and holes in germanium as compared with silicon are correlated with a small shift of energy gap with dilation."

621.3.013.783† **3033**

Generalized Impedance Method Applied to Electromagnetic Field Screening—G. Bon-figlioli and G. Montalenti. (*Alta Frequenza*, vol. 19, pp. 93–105; April, 1950. In Italian, with English, French and German summaries.) Cases are first discussed in which problems of em wave propagation in discontinuous media can be treated by Schelkunoff's method for an infinitely long line. Two special cases are in-vestigated: a spherical-wave generator within a spherical screen and a cylindrical-wave generator within a cylindrical screen. For a screen of small thickness, and for a stated relation between screen thickness, conduc-tivity and permeability and generator fre-quency, frequency ranges may occur within which a copper screen is more efficient than an iron one.

621.384.611.1† **3034**

The Theory of Electron Acceleration in the Alternating Magnetic Field—J. Picht. (*Optik*, vol. 6, pp. 40–55, 61–97, and 133–144; January–March, 1950.) The theory of the betatron (or rheotron) is considered. The analysis is based on the Hertzian vector and avoids the distinc-tion drawn in previous work between the "guiding" field and the "accelerating" field, since it is the resultant field that determines the electron motion. The stability of the elec-tron path is investigated and the conditions for stability are given for the case of fields of strengths used in practice. The possibility of focusing electrons emitted in different direc-tions is discussed.

621.396.822:621.315.592† **3035**

On the Noise Spectra of Semiconductor Noise and of Flicker Effect—A. van der Ziel. (*Physica, 's Grav.*, vol. 16, pp. 359–372; April, 1950. In English.) An explanation is given of the fact that in flicker effect and semiconductor noise the noise intensity is inversely propor-tional to the frequency over a wide range of frequency. General theorems relating to Fourier analysis of fluctuating quantities are mentioned and it is shown to be impossible for the noise intensity to be inversely pro-portional to the frequency throughout the interval $0 < f < \infty$. Inverse proportionality over a wide range can be obtained if a distribution function of correlation times is introduced. Such a function should be expected from the various mechanisms involved.

The sources of noise in semiconductors are considered; the shot effect discussed by Bril-louin, Bernamont, and Gisolf (667 of April) is not sufficient to account for all the noise. Other mechanisms, such as flicker effect and resistance fluctuations due to movements of foreign atoms or to lattice distortions, con-tribute to give the requisite distribution of correlation times.

Flicker effect is discussed, with particular reference to Macfarlane's theory (4087 of 1947). It is suggested that flicker effect in oxide-coated cathodes may not be primarily due to the surface of the coating, but much more to what happens inside it. In an appendix, Bernamont's and Gisolf's treatment of shot noise are shown to be equivalent.

539 **3036**

Ions, Electrons and Ionizing Radiations [Book Review]—J. A. Crowther. Publishers: Arnold, London, 8th edn 1949, 322 pp., 21s. (*Nature* (London), vol. 166, p. 328; August 26, 1950.) This new edition of a work which since 1919 has served "as an introduction to the major treatises on atomic physics" retains and brings up to date the account of the main lines of progress.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.165:621.396.645.371 **3037**

Linear Preamplifier for Driving a Long Coaxial Cable.—Lewis, Robinson, and Toll. (*See* 3013.)

523.746:621.396.812.5 **3038**

Results of Statistical Analysis of Fade-Outs—D. Stranz. (*Arch. elekt. Übertragung*, vol. 4, pp. 217–218; June, 1950.) Correlation of sunspot data with 130 cases of total disap-pearance of radio signals observed during 1948.

523.75 **3039**

The Solar Flare of November 19, 1949 and Cosmic Rays—J. Clay and H. F. Jongen. (*Phys. Rev.*, vol. 79, pp. 908–909; September 1, 1950.) Cosmic-ray intensity was recorded using both shielded and unshielded ion chambers. With a normal flare the intensity falls below the quiescent value some few hours after the flare. This flare was abnormal in that there was no difference in cosmic-ray intensity before and after it.

A fade-out of the Noordwijk radio station was noted and also a magnetic disturbance with a maximum range in H of 215γ.

523.854:621.396.822 **3040**

Present-Day Ideas on the R.F. Radiation from the Galaxy—R. Gallet. (*Rev. Sci.* (Paris), vol. 87, pp. 157–161; July, September, 1949.) Jansky's theory, that galactic radiation origi-nates in the ionized interstellar gas, and the theory of Pawsey and his co-workers, that this radiation is emitted by the stars, are discussed. The apparent and electronic temperatures of the sources deduced from the observations of various workers are very high, and alternative theories such as that of plasma oscillations are examined. Further fundamental quantitative investigation is necessary before any theory can be finally accepted.

551.5:621.396.9 3041
The Temperatures at the Tops of Radar Echoes Associated with Various Cloud Systems—R. F. Jones. (*Quart. Jour. R. Met. Soc.*, vol. 76, pp. 312–330; July, 1950.)

551.5:621.396.9 3042
Radar Weather Echoes—R. F. Jones. (*Met. Mag.*, vol. 79, pp. 109–112, 143–145, 170–172, and 198–200; April–July 1950.) Some typical examples are given of 10-cm radar echoes obtained over a two-year period, using ppi and height-range equipment. The echoes correspond with most types of weather phenomena observed in the temperate zone. The interpretation of the echoes from warm and cold fronts agrees well with generally accepted theories, but the phenomena of occluded fronts may be more complex and not capable of interpretation by a simple model. A few examples of unusual types of echo are given, with their probable explanations.

551.5:621.396.9 3043
Radar Echoes—J. E. N. Hooper and A. A. Kippax; R. F. Jones. (*Quart. Jour. R. Met. Soc.*, vol. 76, pp. 330–336; July, 1950.) Discussion on papers abstracted respectively in 2221 of November and 3042 above.

551.510.535 3044
Large-Scale Sporadic Movements of the E-Layer of the Ionosphere—N. C. Gerson. (*Nature* (London), vol. 166, pp. 316–317; August 19, 1950.) Very marked and rapid sporadic-E movements during the period May 15–16, 1949 are indicated according to preliminary information supplied by a network of American amateur observers operating on frequencies around 50 Mc. Results are charted on a sketch map. The displacements of the reflection points may be due either to movements of the atmosphere or to electron drift.

551.510.535:535.36 3045
Diffusion in the Ionosphere—M. H. Johnson and E. O. Hulburt. (*Phys. Rev.*, vol. 79, pp. 802–807; September 1, 1950.) "Diffusion is treated by showing that the action of a medium on a diffusing gas is that of a dissipative force. When the theory is applied to an electrically neutral ionic gas in a gravitational field it is found that the mixture of positive and negative ions diffuse as a single gas because of the electrical polarization charges within the ionic cloud. In the presence of a magnetic field, the diffusion cannot be expressed in terms of the ionic density until the electrodynamical equations governing the flow of electrical current have been explicitly solved. Solutions are obtained for special cases which show that a strong magnetic field completely inhibits the diffusion due to concentration gradients in the transverse plane and has little effect on the diffusion due to the gravitational force."

551.510.535:551.594.5 3046
Narrowly Limited Ionization Clouds at 125-km Height during an Auroral Disturbance—D. Stranz. (*Arch. elekt. Übertragung*, vol. 4, pp. 213–216; June, 1950.) Analysis of Swedish records of September 15–16, 1948, showing sporadic-E reflections lasting $2\frac{1}{2}$ hours. Corresponding records of variations of the magnetic vector are also given.

551.55:551.510.535:621.396.9 3047
Meteoric Echo Study of Upper Atmosphere Winds—Manning, Villard, and Peterson. (*See* 3052.)

551.594.6 3048
Noise Levels in the American Sub-Arctic—N. C. Gerson. (PROC. I.R.E., vol. 38, pp. 905–916; August, 1950.) The level of atmospherics on a frequency of 150 kc was studied over a period of six months in Northern and Southern Canada. The average, quasi-peak, and maximum peak levels were determined. Diurnal variation is absent or small in winter, but shows a marked increase in summer, with a minimum after sunrise and a maximum after sunset. As summer approaches the noise-level rise occurs earlier and earlier. Static intensity is higher in June than in January and at southern stations noise increased progressively up to June. A rise in noise level at some stations in February is attributed to precipitation static caused by blizzards. The noise level varied as a nonlinear function of colatitude. Recorded data are included.

LOCATION AND AIDS TO NAVIGATION

621.396.9:551.5 3049
The Temperatures at the Tops of Radar Echoes Associated with Various Cloud Systems—R. F. Jones. (*Quart. Jour. R. Met. Soc.*, vol. 76, pp. 312–330; July, 1950.)

621.396.9:551.5 3050
Radar Echoes—J. E. N. Hooper and A. A. Kippax; R. R. Jones. (*Quart. Jour. R. Met. Soc.*, vol. 76, pp. 330–336; July, 1950.) Discussion on papers abstracted respectively in 2221 of November and 3042 above.

621.396.9:551.5 3051
Radar Weather Echoes—Jones. (*See* 3042.)

621.396.9:551.55:551.510.535 3052
Meteoric Echo Study of Upper Atmosphere Winds—L. A. Manning, O. G. Villard, and A. M. Peterson. (PROC. I.R.E., vol. 38, pp. 877–883; August, 1950.) The Doppler frequency shift imparted to continuous waves reflected from a meteoric ionization column in the 80–110-km height region enables the wind drift of the trail to be found. Statistical analysis enables average wind velocities to be measured to within about 20 per cent and direction to within about 20°, in a period of one or two hours. Typical wind velocities were found to be about 125 km per hour during the early morning hours in the summer of 1949; occasional evidence of nonuniform wind structure was found.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.743:621.396.611.21 3053
A Double-Crystal X-Ray Goniometer for Accurate Orientation Determination—W. L. Bond. (PROC. I.R.E., vol. 38, pp. 886–889; August, 1950.) The slit system of an X-ray goniometer is replaced by an AT quartz plate, thus turning it into a double-crystal instrument suitable for accurate measurements on AT quartz plates. Errors due to temperature change and to X-ray refraction are discussed. Measurement accuracy to within about 0.1′ of arc is attained.

533.5 3054
Measurement of the Pumping Rate of Rotary Pumps—R. Henry. (*Le Vide*, vol. 5, pp. 859–865; July–September, 1950.) Fundamental definitions are stated, the method and apparatus used are described, and results obtained with various French and American models are tabulated.

535.215.1:549.211 3055
Photoelectric Properties of Diamond, Measured with a Crystal Counter—G. P. Freeman and H. A. van der Velden. (*Physica, 's Grav.*, vol. 16, pp. 486–492; May, 1950. In English.)

535.37 3056
Dependence of Emission Spectra of Phosphors upon Activator Concentration and Temperature—G. R. Fonda. (*Jour. Opt. Soc. Amer.*, vol. 40, pp. 347–352; June, 1950.)

535.37 3057
The Rise in Brightness of Infrared-Sensitive Phosphors—P. Brauer. (*Jour. Opt. Soc. Amer.*, vol. 40, pp. 353–355; June, 1950.)

535.37:546.472.84 3058
Rise and Decay of Willemite Luminescence—G. Gergely. (*Jour. Opt. Soc. Amer.*, vol. 40, pp. 356–361; June, 1950.) The luminescence/time curve for cr excitation times of 1–280 μs can be resolved into three monomolecular components with time constants of 200, 1,800 and 72,000 per second. The fast component is predominant for short excitation times and the slow component for long times. The duration of excitation does not affect the decay rates, but varies the proportions of the three components.

537.363+621.357.1]:546.841 3059
On the Polarization of the Electrodes in the Electrophoresis and the Electrolysis of Thoria and Thorium Nitrate—G. Mesnard. (*Le Vide*, vol. 5, pp. 866–869; July–September, 1950.)

538.27:621.314.22.042.2.029.6 3060
Theory of Magnetically Inhomogeneous Surface Layers in Transformer Laminations—R. Feldtkeller. (*Frequenz*, vol. 4, pp. 129–134; June, 1950.)

539.11:621.315.612.4 3061
Atomic Positions and Vibrations in the Ferroelectric $BaTiO_3$ Lattice—W. Kaenzig. (*Phys. Rev.*, vol. 80, pp. 94–95; October 1, 1950.)

539.2:537.228.1 3062
Domain Structure of Rochelle Salt—J. Furuichi and T. Mitsui. (*Phys. Rev.*, vol. 80, pp. 93–94; October 1, 1950.)

539.234 3063
Production of Thin Layers of Uniform Controlled Thickness by Evaporation: Applications in Electronics—C. Dufour. (*Le Vide*, vol. 5, pp. 837–843; July–September, 1950.) Illustrated description of methods used for producing thin layers such as are required especially in television and radar tubes. Layers uniform to within about 1 per cent have been produced on a plane surface 20 cm in diameter, the thickness being controlled photoelectrically during processing to within about 0.02μ.

546.431.82 3064
Ferroelectricity, Domain Structure, and Phase Transitions of Barium Titanate—A. von Hippel. (*Rev. Mod. Phys.*, vol. 22, pp. 221–237; July, 1950.) Detailed summary of MIT research to date, with discussion of: variations in the ferroelectric parameters of multicrystalline $BaTiO_3$ cooling through the Curie region; piezoelectric response with changing frequency and temperature; hysteresis phenomena; phase transitions of the BaO-TiO_2 system; the domain structure of $BaTiO_3$ single crystals and the effect of an applied electric field; theory of the ferroelectric state in $BaTiO_3$. Extensive references and many diagrams are included.

549.514.5:621.3.011.5 3065
Dielectric Constant of Silica—S. K. K. Jatkar and B. R. Y. Iyengar. (*Indian Jour. Phys.*, vol. 23, pp. 145–152; April, 1949.) The permittivities of crystal-quartz and fused-silica plates, as determined by the plate-capacitor method, are respectively 4.54 and 3.2. Measurements by the liquid-mixture method give for powdered quartz the value 4.55. The results are discussed with reference to the theory of molecular structure.

620.19 3066
Investigations into Tropic Proofing of Electrical Materials, 1943–46: Parts 1–5.—(*Aust. Jour. Appl. Sci.*, vol. 1, pp. 80–132; March, 1950.)

Part 1. The protection of electronic equipment for use under humid tropical conditions—

. G. Dobbie. An outline of investigations in ustralia. Moisture was found the main cause · electrical deterioration, particularly under nsuitable storage conditions. Suitable materials, protective coatings and layout are uggested and laboratory tests outlined. Complete hermetic sealing of components is the nost satisfactory method of protection. Two ppendices give general data on evaporation nd water-vapor diffusion.

Part 2. The influence of moisture on insulating materials.—J. S. Dryden and P. T. Vilson. Limitations of methods commonly sed for measuring surface and volume resistivity are discussed. Different types of naterial are compared; surface-treated glass nd ceramics were found the best for maintaining high values of insulation resistance nder humid conditions.

Part 3. Some experiments on the application of organosilicon compounds to glass and ceramic—R. J. Meakins, J. W. Mulley, and . R. Churchward. The insulation resistance f glass and steatite at high relative humidities s increased by treatment with a solution of nethyl or ethyl chlorosilane in benzene. Similar esults are obtained with methyl- and ethylilicon amine solutions, but prior cleaning vith chromic acid is essential.

Part 4. The treatment of glass and steatite ceramic with quaternary ammonium compounds—R. J. Meakins. Insulation resistance f glass and steatite is increased nearly 1,000 imes after immersion for one second in an queous solution of certain quaternary ammonium compounds, but as a practical means f surface treatment these compounds are less ffective than the alkylchlorosilanes.

Part 5. The corrosion of copper wires at dc otential in contact with electrical insulating naterials—R. J. Meakins. The anodic corrosion of copper wire in contact with a wide ange of insulating materials under very moist onditions was investigated. Results indicate hat the degree of corrosion depends on the ffinity for water of the insulator. With polar naterials, such as cellulose products, severe corrosion occurred, but little corrosion was ound with nonpolar materials such as paraffin nd ceresine waxes, bitumens and polyethylene.

621.3.011.5:676.19 3067

Investigations of the Dielectric Properties of Paper Fibre—P. Henninger. (*Frequenz*, vol. 4, pp. 167–177; July, 1950.) The experimentally determined dielectric properties of different types of paper are discussed in the light of its cellulose structure. Temperature coefficients for the different structures and the effect of absorption of water are studied.

621.3.032.2:621.315.61:68 3068

Microspacer Electrode Technique—O. M. Stuetzer. (PROC. I.R.E., vol. 38, pp. 871–876; August, 1950.) Closely spaced multiple-electrode systems with glass, quartz or other thermoplastic insulation can be produced by drawing down from a convenient larger size, the geometry of the system undergoing little change in the process. Electrode materials can in some cases be inserted prior to the drawing; in others, wires of suitable size are inserted after drawing and sealed if necessary. Applications of the technique in the construction of subminiature tubes, crystal amplifiers, and flat parallel-wire screens, are described.

621.315.592† 3069

On the Nature of a Soldered Contact on a Semiconductor—J. I. Pantchechnikoff. (*Phys. Rev.*, vol. 79, pp. 1027–1028; September 15, 1950.) An account of experiments made to verify the assumption that the soldered contact is effectively a gradual transition from a metal to a semiconductor. Metal atoms diffuse into the semiconductor during the soldering process, their concentration being greatest near the surface through which they have diffused. A theoretical explanation of effects observed with soldered contacts is given.

621.315.592†:[621.316.86+621.314.632 3070

Semiconductors and Their Applications—R. W. Douglas. (*GEC Jour.*, vol. 17, pp. 107–124; July, 1950.) Conduction in metals is considered and a typical crystal structure for a metal is contrasted with that of a semiconductor. Intrinsic, impurity and nonstoichiometric types of semiconductors are considered and their properties discussed from the chemical-bond viewpoint and more fully in terms of quantum mechanics. Applications discussed in detail include thermistors, point-contact rectifiers, high-back-voltage Ge diodes, layer-type rectifiers and Ge triodes.

621.315.61 3071

Fluid Dielectrics for the Decimetre and Centimetre Wave Band—W. Endres and H. Köhler. (*Frequenz*, vol. 4, p. 145; June, 1950.) Correction to paper abstracted in 1708 of August.

621.318.23 3072

The Design of [Magnetic] Circuits for Permanent Magnets—A. Hug. (*Bull. Schweiz. Electrotech. Ver.*, vol. 41, pp. 661–669; September 2, 1950. In German.) A semi-empirical method of calculation is described which is simple yet sufficiently accurate for practical purposes. Electrical analogies are used in explaining the theory. Magnetizing processes and measurements on permanent magnets are discussed briefly.

621.318.4.042.15:538.22 3073

The Magnetic Characteristics of Coils with Pot-Type Cores of Powdered Iron—M. Kornetzki. (*Frequenz*, vol. 4, pp. 105–113; May, 1950.) Inductance and iron losses of such coils with a high-μ pot core completely enclosing the coil are calculated theoretically and referred to constants for the material measured on ring cores. The effective permeability of the pot core is proportional to that of the core material, but the proportionality factor depends on the type of winding. Losses due to eddy currents and magnetic after-effect are practically equal to those of the ring core. Hysteresis losses are determined from a calculation of effective length of magnetic path. Measurements on pot-type coils confirm the calculations. The effect of the stray field of these coils is discussed.

621.646.958:537.534 3074

Positive-Ion Emission, a Neglected Phenomenon—W. C. White. (PROC. I.R.E., vol. 38, pp. 852–857; August, 1950.) General discussion, with description of a leak detector based on positive-ion effects.

666.1:621.317.374 3075

Some Experiments and Theories on the Power Factor of Glasses as a Function of Their Composition: Part 1—J. M. Stevels. (*Philips Res. Rep.*, vol. 5, pp. 23–36; February, 1950.) Power factors and permittivities of various series of glasses whose composition varied systematically were measured at a temperature of 20°C and a frequency of 1.5 Mc. Results are given in tables and graphs.

669.018.58 3076

Recent Developments in Magnetic Materials—H. Fahlenbrach. (*Z. Ver. Dtsch. Ing.*, vol. 92, pp. 565–570; July 21, 1950.) A survey with 38 references. The modifying effect of the new materials on the design of apparatus is indicated, both where permanent magnets and where electromagnets are used.

679.5:[53+54 3077

Fundamentals of Synthetic-Polymer Technology in Its Chemical and Physical Aspects [Book Review]—R. Houwink. (Elsevier's Polymer Series, No. 1.) Publishers: Elsevier Publishing Co., New York and Amsterdam; Cleaver-Hume Press, London, 1948–1950, 258 pp., 28s. (*Nature* (London), vol. 166, pp. 575–577; October 7, 1950.) "The subject matter includes the molecular and colloid chemistry of polymers, their more important mechanical and physical properties and methods of testing these, and their general processing, followed by an account of the manufacture and outstanding properties of industrial products of this type, whether obtained by chemical treatment of natural colloidal products or by synthesis from non-colloid raw materials."

679.5:[53+54 3078

Elastomers and Plastomers: Their Chemistry, Physics and Technology [Book Review]—R. Houwink (Ed.). (Elsevier's Polymer Series, No. 3.) Publishers: Elsevier Publishing Co., New York and Amsterdam; Cleaver-Hume Press, London, 1948–1950. Vol. 1: General Theory. 495 pp., 52s. 6d. Vol. 2: Manufacture, Properties and Applications. 515 pp., 50s. 6d. Vol. 3: Testing and Analysis: Tabulation of Properties. Contributed by B. B. S. T. Boonstra, A. G. Epprecht, R. Houwink, J. H. Teeple, and J. W. F. van't Wout. 174 pp., 27s. 6d. (*Nature* (London), vol. 166, pp. 575–577; October 7, 1950.) The ground covered is similar to that for No. 1 (3077 above) but the treatment is much fuller and is extended to include also products such as natural resins and asphalts.

MATHEMATICS

519.21+519.271.3]:517.9 3079

A Sampling Method for Determining the Lowest Eigenvalue and the Principal Eigenfunction of Schrodinger's Equation—M. D. Donsker and M. Kac. (*Bur. Stand. Jour. Res.*, vol. 44, pp. 551–557; May, 1950.)

519.3 3080

A Variational Method Suitable for studying the Wave Equation—T. Kahan. (*Rev. Sci.* (Paris), vol. 87, pp. 205–211; October–December, 1949.)

681.142 3081

Arithmetic Operations in a Binary Computer—R. F. Shaw. (*Rev. Sci. Instr.*, vol. 21, pp. 687–693; August, 1950.)

681.142 3082

Electronic Pile Simulator—P. R. Bell and H. A. Straus. (*Rev. Sci. Instr.*, vol. 21, pp. 760–763; August, 1950.) A calculator is described which can deal with the set of equations applicable to a chain-reaction pile. The voltage output varies as a function of time just as the neutron flux would vary in a pile. A potentiometer varies the effective multiplication factor and simulates the action of a control rod. Five delayed neutron periods are simulated. An electronic integrator is used and the accuracy is well within 1 per cent.

681.142 3083

Programme Organization and Initial Orders for the EDSAC—D. J. Wheeler. (*Proc. Roy. Soc. A*, vol. 220, pp. 573–589; August 22, 1950.)

681.142 3084

Solution of Simultaneous Equations through Use of the A.C. Network Calculator—L. M. Haupt. (*Rev. Sci. Instr.*, vol. 21, pp. 683–686; August, 1950.)

51 3085

Progressive Mathematics [Book Review]—P. Clyne. Publishers: Chapman and Hall, London, 1950, 270 pp., 15s. (*Beama Jour.*, vol. 57, p. 239; August, 1950.) Directed mainly to students of engineering and physics up to the Higher National Certificate standard. Some unorthodox methods are used. "The book is well printed and produced, and its modest price should be the means of making it available to most students."

MEASUREMENTS AND TEST GEAR

621.317.1:621.397.61:621.396.619.24 3086

Television Transmitter Lower-Sideband Measurements—G. E. Hamilton and R. G. Artman. (*TV Eng.*, vol. 1, pp. 12–15, 25, and 14–15, 30; April and May, 1950.) Theoretical and practical considerations are discussed which are involved in the measurement of the spectral energy distribution of television signals. Account is taken of transmitter adjustment and receiver sensitivity. Measurement methods described include (*a*) radio-frequency excitation of the input circuit of the modulation amplifier, (*b*) sine-wave modulation of the transmitter, (*c*) composite-video-signal modulation of the transmitter. Method (*b*) is preferred.

621.317.3/.4:621.315.592 3087

Measurement of Electric and Magnetic Constants of Semiconductors at Ultra-High Frequency using Concentric Line—G. Untermann. (*Z. Angew. Phys.*, vol. 2, pp. 233–241; June, 1950.) The formulas necessary for evaluating measurements on concentric lines are derived. Calculations are facilitated by the graphical presentation of certain functions, notably $z \cot z$ and $z \tan z$ for complex arguments. By means of these it is possible to determine rapidly the values of dielectric constant, conductivity, permeability and magnetic loss angle at ultra-high frequency.

621.317.3.029.6+[621.3.09:621.315.2+621.392.26† 3088

Lines and Circuit Elements for Microwave Measurements—O. Schäfer and R. Honerjäger. (*Arch. tech. Messen*, pp. T87–T90; August, 1950.) Discusses propagation of em waves in coaxial lines and waveguides, the design of terminations, attenuators, directional couplers, impedance bridges, resonator wavemeters etc.

621.317.321.027.21 3089

The Low-Loss Measurement of Direct Voltage by Conversion into Alternating Voltage—H. H. Rust. (*Z. Angew. Phys.*, vol. 2, pp. 290–293; July, 1950.) The method previously described (1451 of July) using a carbon microphone as transducer is only suitable if the resistance of the dc source is low; otherwise a capacitor microphone is more appropriate, on account of its high input resistance. The unknown direct voltage is used as a polarizing voltage, and a proportional ac output is obtained when the microphone is exposed to a tone of constant amplitude and frequency (here 2 kc). With an aperiodic amplifier the lowest voltage measurable is 10^{-3} v, but sensitivity can be increased by using a selective amplifier.

621.317.324(083.74)† 3090

Development of Very-High-Frequency Field-Intensity Standards—F. M. Greene and M. Solow. (*Bur. Stand. Jour. Res.*, vol. 44, pp. 527–547; May, 1950.) A description is given of the development of two field-intensity standards which are being used at the National Bureau of Standards for the calibration of commercial field-intensity sets in the range 30–300 Mc. These standards are employed to establish known values of field intensity by either of two methods: (*a*) the standard-antenna method in which the open-circuit voltage at the center of a receiving dipole is measured directly; (*b*) the standard-field method in which the current at the center of a transmitting dipole is accurately known. The techniques used for determining the antenna current and voltage are described. The current distribution on the antenna is determined theoretically, using Schelkunoff's method, which gives the effective length. These values are compared with those obtained by measurement. Results of field tests at 100 Mc are presented in which the above two methods were directly compared, using horizontal polarization. Their accuracy and limitations are discussed.

621.317.333.4.015.7:621.315.2 3091

Location of Line Faults by Oscillographic Observation of a Pulse—C. Béguin and G. Maugard. (*Bull. Soc. Franç. Elec.*, vol. 10, pp. 313–328; July, 1950.) A theoretical study is made of the conditions of propagation of a pulse along a line and of the distortion of pulses reflected at line faults. The pulse method of investigation is particularly suitable for coaxial telephone lines and high-voltage cables, but not for lines with high attenuation. Equipment described uses a circular timebase, with arrangements for locating the steep front of the initial pulse at the zero of the scale. Pulse duration is 1–4 μs, depending on the circuit range, and peak power is 250 w. Tests on power lines are described in detail; the method is quicker than the bridge method.

621.317.335.3.029.62/.63† 3092

Complex-Dielectric-Constant Measurements in the 100-1000-Mc/s Range—A. G. Holtum, Jr. (Proc. I.R.E., vol. 38, pp. 883–885; August, 1950.) A sample of a medium-loss or high-loss material is used as the dielectric of a capacitor terminating a slotted measuring line. The complex impedance of this termination, and hence the dielectric constant and conductivity of the lossy material, can be determined by the conventional method from the voltage SWR and the position of the minimum.

621.317.336.029.64 3093

The Accuracy of Impedance Measurement with Microwaves—F. Tischer. (*Kungl. tekn. Högsk. Handl.*, (Stockholm), 31 pp.; 1950. In German, with English summary.) The theory of impedance matching is discussed and standing-wave ratio and matching error are defined; approximate formulas are derived for the case of small errors. Known types of apparatus, such as reflectometers, directional couplers, impedance bridges and comparators, are discussed. The method using a standing-wave meter is dealt with at length, this giving the best accuracy. Errors originate from two sources, the line discontinuities of the standing-wave meter and the probe used as voltage pickup. The errors can be largely eliminated by proper design. The results of the investigation have been applied in the development of two instruments, one a precision meter giving the SWR to within 0.5 per cent and the other a rotary automatically recording meter for the 10-cm band, accurate to within 3 per cent.

621.317.352:621.392.5 3094

Determination of Attenuation from Impedance Measurements—R. W. Beatty. (Proc. I.R.E., vol. 38, pp. 895–897; August, 1950.) The dissipative and reflective components A_d and A_r of a linear, passive, quadripole attenuator are measured separately. A_d, a function of the attenuator efficiency, is determined from reflection-coefficient measurements of the short-circuited attenuator and A_r from a single voltage-SWR measurement with the attenuator terminated by a matched load. Experimental results are in close agreement with measurements by another method.

621.317.432 3095

Eddy-Current Losses in Electrical Equipment—A. Roth. (*Rev. Gén. Elec.*, vol. 59, pp. 268–278; June, 1950.) Formulas are derived and graphs presented for rapidly finding the approximate value of eddy current losses in various cases. Numerical examples are given. Wattmeter methods used in industry for measuring these losses are reviewed, and precautions necessary when using them are indicated. A simple voltmeter method is described.

621.317.7:621.314.632.1† 3096

Design of Square-Law Rectifier Circuits for Measuring Instruments—D. C. Walker, D. L. Richards, and G. P. Horton. (*P.O. Elect. Eng. Jour.*, vol. 43, Part 2, pp. 74–77; July, 1950.) For accurate rms indications of speech currents, a scale having a range of at least 16 db is desirable. Networks consisting of rectifiers in combination with series and shunt resistors are described with which measurement accuracies to within about ±0.3 db have been achieved.

621.317.723 3097

New Types of Electrometer—P. Böning. (*Arch. tech. Messen*, p. T70; June, 1950.) The device described consists basically of a metal cylinder, with axis horizontal, containing a fixed metal frame electrically connected to a similar frame rotatable about the axis and carrying a pointer. A potential to be measured is applied between cylinder and frames. The theory of operation is given.

621.317.726 3098

The New Modulation Meter—M. Bidlingmaier. (*Frequenz*, vol. 4, pp. 146–149; June, 1950.) Description of a peak voltmeter with an approximately logarithmic scale and an optical indicator system. The scale covers an amplitude range of about 1 to 300 and peaks of duration as short as 10 ms can be measured within 1 db.

621.317.755 3099

Oscillograms Obtained without Timebase: "Integral-Curve" Method—F. Perrier. (*Rev. Gén. Élec.*, vol. 59, pp. 345–351; August, 1950.) A cro method is described in which a function is compared with its derivative. Examples of application include both slowly varying and rapid phenomena. Oscillograms reproduced correspond to the superposition of $>10^5$ curves and demonstrate the immobility of the trace on the screen. Thus long photographic exposures can be used to give the mean value, and the fluctuations about the mean value, of phenomena repeating at irregular intervals.

621.317.755:621.3.012 3100

High-Frequency Curve Tracer for 460-480 kc/s and 10.0-11.5 Mc/s—A. Klemt. (*Funk u. Ton*, vol. 4, pp. 396–402; August, 1950.) A cro is described for checking intermediate-frequency band-pass filters and aligning intermediate-frequency amplifiers of broadcast receivers. The upper frequency range deals with FM receivers, where particular care is required.

621.317.772 3101

High-Frequency Phase Measurement with Direct Indication: Part 2—With Pointer Indicators—A. Ruhrmann. (*Arch. tech. Messen*, pp. T64–65; June, 1950.) The pointer instrument has the advantage over the cro as regards simplicity of operation and long life. Arrangements using crossed-coil and four-quadrant indicator instruments are discussed. Part 1: 2574 of December.

621.317.79:621.396.822:621.396.61 3102

An Instrument for the Determination of Frequency-Modulated Hum in a Radio-Frequency Signal—W. W. Boelens. (*Commun. News*, vol. 10, pp. 101–109; April, 1950.) The instrument uses a very sensitive PM detector which can be calibrated without any additional apparatus. The theory of the method of hum measurement [2837 of 1949 (Boelens and Stumpers)] is outlined and several special circuits incorporated in the apparatus are described. These include a high-pass filter without any inductance, a double-T *RC* filter and an attenuator.

621.385.001.4 (083.74) 3103

Standards on Electron Tubes: Methods of Testing, 1950—(Proc. I.R.E., vol. 38, pp.

9-948, and 1079–1093; August and September, 1950.) Reprints of this standard, 50 IRE7.S2 (Parts I & II) may be purchased while available from The Institute of Radio Engineers, 1 East 79 Street, New York 21, at $1.25 per copy, with 20 per cent discount for 100 or more copies.

6.396.621.001.4 3104
Critical Study of Measurements on Receivers—L. Chrétien. (*TSF pour Tous*, vol. 26, pp. 212–217; June, 1950.) Methods of making electrical measurements to determine receiver performance are discussed. Standards are fixed for AM test signal, dummy antenna for medium and short waves, output signal, sensitivity necessary to give this output in the presence of noise. The best simple methods of checking the efficiency of the automatic control circuits and measuring the selectivity and distortion are indicated. One criticism of normal tests is that a receiver which is judged "perfect" according to such tests may not entirely satisfy the ear of a true artist.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

5.53:621.316.8:621.396.822 3105
The Use of Spontaneous Voltage Fluctuations for the Measurement of Low Temperatures—P. M. Endt. (*Physica, 's Grav.*, vol. 16, pp. 481–485; May, 1950. In English.) Discussion of Lawson and Long's method (483 of 47). The statistical fluctuations in the measured noise are calculated, taking account of the bandwidth of the amplifier. The calculation shows that it is not necessary to use reduced anode, screen-grid, and heater voltages in the input tube, and that a quartz crystal is essentially worse than a resistor or an *LC* circuit as a noise generator for temperature measurement.

16.58:621.314.3† 3106
A Direct-Voltage Amplifier as Power Amplifier and Temperature Regulator—Jellinghaus. (*See* 2990.)

37.533.73:621.315.611:539.213.26 3107
Electron Diffraction Apparatus for Investigating Insulator Surfaces—J. J. Trillat and Oketani. (*Le Vide* (Paris), vol. 5, pp. 827–30; July, September, 1950.) The apparatus, which is described in detail, includes a new type of neutralizing gun for preventing accumulation on the specimen of negative charges carried by the exploring beam. This gun uses a platinum filament coated with barium and strontium oxides and operated at a temperature below that at which vaporization occurs. Thus the investigation can extend over a long period without the specimen becoming coated. Photographs are shown of diagrams obtained with the apparatus.

38.569.2.047:621.315.61.011.5 3108
Dielectric Properties of the Human Body for Wave-Lengths in the 1–10 cm Range—T. S. England. (*Nature* (London), vol. 166, pp. 480–481; September 16, 1950.) Previously reported measurements on a wavelength of 3.18 cm (2282 of 1949) have been repeated on wavelengths of 1.27 cm and 10.0 cm. Results are tabulated and discussed, using the values obtained for water as a comparison standard.

539.16:614.8 3109
Measurement Apparatus for the Protection of Personnel against Radioactive Radiations—Weill. (*Onde Elec.*, vol. 30, pp. 328–334; July, 1950.) The various radiations to be guarded against and the measurement of their intensity by means of photographic plates, Geiger counters and ionization chambers are briefly described.

549.514.51.001.8 3110
Applications of Quartz Crystals—R. Villem. (*Rev. Gén. Élec.*, vol. 59, pp. 247–268; June, 1950.) The equivalent circuit of the quartz crystal and the order of magnitude of the impedances involved are discussed. The mechanisms of oscillation stabilization and of frequency selection are studied, and the advantages of using crystals for these purposes are demonstrated. Applications described in detail include a highly stable generator for common-frequency broadcasting, the Lepaute oscillometer for observing the operation of watches, underwater sounding equipment, ultrasonic generators, and devices for measuring pressure, especially in machine tools, vehicles and power-line pylons. Filters making use of quartz crystals receive special attention. An appendix gives a résumé of the properties of T, π, ladder and bridge filters.

621.317.755 3111
A Cathode-Ray Oscillograph for Investigating Functions of Two Variables—S. Genç. (*Tech. Mitt. Schweiz. Telegr.-Teleph. Verw.*, vol. 28, pp. 342–348; September, 1950. In German.) A description is given of a cro in which the functions are displayed as surfaces in three dimensions. Circuit relations are established by means of which the usual deflector system comprising two mutually perpendicular em fields is able to produce spot displacements corresponding to a three-field deflector system. Several photographs of screen displays are shown.

621.365.54† 3112
A General Theory and Methods of Design for Medium-Frequency [Electric] Furnaces—M. Van Lancker. (*Bull. Soc. Franç. Élec.*, vol. 10, pp. 439–455; Discussion pp. 455–456; September, 1950.) Simple and generalized penetration are discussed. Parameters affecting the resistance and inductance for a furnace of infinite length are calculated, and coefficients of resistance and generalized inductance and correction factors for furnaces of finite length are derived. Form factor, field factors, and total input power are investigated. The theory is applied to the study of the design of an industrial furnace.

621.365.54† 3113
Rapid Tempering by Induction Heating—J. F. Libsch and A. E. Powers. (*Metal Progress*, vol. 58, pp. 176–180; August, 1950.) Short-term tempering may be performed successfully on carbon and low-alloy steels by the induction-heating method, provided the temperature is increased to compensate for the reduced time.

621.383:621.385.15:535.245.1 3114
Detection of Short Pulses of Light Using an Electron-Multiplier Circuit—F. Valentin. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 2271–2272; June 26, 1950.) Light pulses from a Hg arc, with a repetition rate of 50 per second, and an instantaneous brilliance inversely proportional to their duration, are used in experiments on the molecular diffusion of light, the re-emitted flashes being received by a multiplier connected to a suitable amplifier. The signal is modulated at high-frequency to enable a mains-operated ac amplifier to be used. Signal-to-noise ratio of the circuit is compared with that of a classic dc amplifier; for flashes of duration 80 μs the two are about equal, but in the new circuit the threshold of detection can be considerably reduced by shortening the flashes.

621.384.611.1† 3115
The Theory of Electron Acceleration in the Alternating Magnetic Field—Picht. (*See* 3034.)

621.384.611.2†:621.316.726 3116
Frequency Control for the Bevatron Radio-Frequency Voltage—J. Riedel. (*Elec. Eng.*, vol. 69, pp. 721–722; August, 1950.) Summary of AIEE Summer General Meeting paper. The variable element in the control oscillator is a saturable inductance consisting of toroidal coils wound on ferroxcube III.

621.385.833 3117
Contrast Improvement in Electron Microscopy—J. Hillier and E. G. Ramberg. (*Z. Angew. Phys.*, vol. 2, pp. 273–278; July, 1950.) Methods of preparing specimens so as to obtain optimum contrast are discussed.

621.385.833 3118
Electron-Optical Properties of Space-Charge Clouds—L. Marton and D. L. Reverdin. (*Jour. Appl. Phys.*, vol. 21, p. 842; August, 1950.) Tests have been made which tend to confirm the suggestion that a space-charge lens might be used as a corrective element for the reduction of the aperture defect of electron lenses. However, the experiments also support the idea that the random noise in such a space-charge cloud might reduce image definition.

621.385.833 3119
New Simplified Electron Microscopes—V. E. Cosslett. (*Nature* (London), vol. 166, pp. 305–306; August 19, 1950.) A British and an American model are briefly described, both operating at 50 kv and giving a resolving power approaching 100 Å; they cost much less than standard models hitherto available.

621.387.4† 3120
A New Method of Operating Counter Tubes with High Resolving Power—A. Trost. (*Z. Angew. Phys.*, vol. 2, pp. 286–289; July, 1950.) A method using a thyratron circuit provides direct quantitative indication of intensity even at the very high pulse rates encountered with powerful X-rays or γ-rays. An experimental investigation of the life of quenched counter tubes is also described.

621.387.4† 3121
The Behaviour of Counters with Pure Vapour Filling in the Proportional and Geiger Regions—E. Fünfer and H. Neuert. (*Z. Angew. Phys.*, vol. 2, pp. 241–249; June, 1950.) The variation of pulse height with tube voltage was measured for various vapors, and for α- and β-radiation. Counters with pure vapor filling have a much more extensive proportional region than mixture-filled counters. For β-radiation the limit of the true proportional region corresponds to amplification factors of 10^6–10^7. With heavy vapors so-called 'oversize' pulses may occur.

621.387.4† 3122
Three Examples of Electronic Apparatus Used for [Particle] Detection—G. Valladas. (*Onde Elec.*, vol. 30, pp. 317–320; July, 1950.) The auxiliary apparatus required with particle detectors, viz., amplifier, selector and recorder, is discussed.

621.387.4†:549.211 3123
Mobility of Electrons and Holes in Diamond—E. A. Pearlstein and R. B. Sutton. (*Phys. Rev.*, vol. 79, p. 907; September 1, 1950.) From cro measurements of pulse rise times with voltages of 300–5,000 v applied across a crystal of thickness 2 mm, the mobilities deduced for electrons and holes were respectively 3,900 and 4,800 cm² per second v, with possible errors of 15 per cent and 20 per cent.

621.387.4†:549.211:535.215 3124
The Influence of Red and Infra-red Light on a Crystal Counter—H. A. van der Velden and G. P. Freeman. (*Physica, 's Grav.*, vol. 16, pp. 493–500; May, 1950. In English.) The influence of red light in reducing the space charge built up in a diamond crystal counter is investigated. The counting property of the crystal can be maintained for an unlimited time if the intensity of the incident light is sufficient.

621.387.42† 3125
High-Pressure Ionization Chamber Counters and Their Use—R. Wilson, L. Beghian, C. H. Collie, H. Halban, and G. R. Bishop. (*Rev. Sci. Instr.*, vol. 21, pp. 699–705; August, 1950.) New techniques and counters filled with H_2, D_2 or CH_4 at pressures up to 35 atm., and also the characteristics and applications of these counters, are described.

621.387.42:621.3.015.7 3126
Pulses in Argon Counters—L. Colli, U. Facchini, and E. Gatti. (*Phys. Rev.*, vol. 80, pp. 92–93; October 1, 1950.)

621.387.422† 3127
Boron Trifluoride Proportional Counters—I. L. Fowler and P. R. Tunnicliffe. (*Rev. Sci. Instr.*, vol. 21, pp. 734–740; August, 1950.) Counters with filling of BF_3 and of calculable efficiency for neutron-beam measurements are described.

621.398 3128
Remote Control of [Atomic] Pile Reaction—J. Pottier and V. Raievski. (*Onde Elec.*, vol. 30, pp. 323–327; July, 1950.) A servo arrangement is described which has been in operation without interruption for a year and which controls the position of a neutron-absorbing plate within a cavity in the pile.

621.385.833 3129
Electron Microscopy: Technique and Applications [Book Review]—R. W. G. Wyckoff. Publishers: Interscience, New York and London, 1949, 248 pp., 40s. (*Nature* (London), vol 166, pp. 285–286; August 19, 1950.) Includes much to instruct and interest specialists, but appeals also to a far wider audience by providing in effect 'the story of the new field of vision.'

PROPAGATION OF WAVES

531.74:621.396.11.029.62 3130
Measurement of Small Angles of Elevation of Incoming Metre Waves. Part 1: Theory—H. Stenzel. (*Arch. elekt. Übertragung*, vol. 4, pp. 125–132; April, 1950.) Analysis shows that with antenna systems in the horizontal plane with horizontal polarization and adjustable directivity in the vertical plane, the measurement of small angles of elevation for meter waves is only possible with antennas of impracticable dimensions. For vertical antenna systems with horizontal polarization, however, small angles of elevation can be measured if they are at least 3/4 of the half-value width of the directional characteristic, but in order to measure an angle of 4°, the size of the antenna system must be about 10λ.

A new method is indicated which is based on the comparison of two fixed vertical systems with different directional characteristics, the midpoints being at equal heights above the ground. This method is particularly suitable for the measurement of angles down to 1/4 degree, since the dimensional requirements can be carried out in practice and the effect of the plane and horizontal ground is completely eliminated. This method will be considered more fully in part 2.

538.566 3131
Propagation of Electromagnetic Waves through a Stratified Medium: Part 1—B. Salzberg. (*Jour. Opt. Soc. Amer.*, vol. 40, pp. 465–470; July, 1950.) A general analysis is developed which is applicable to propagation through a medium divided into *m* plane parallel layers, the first and last being semi-infinite. The em properties of the layers are unrestricted and dissimilar, while the thicknesses of the (*m*-2) finite layers are in general different. Formulas for the resulting reflection and transmission coefficients are derived. In part 2 the analysis will be applied to the detailed solution of specific problems.

621.396.11+621.396.67 3132
Symposium on Antennas and Propagation, San Diego, California, April 3-4, 1950—(Proc. I.R.E., vol. 38, pp. 958–962; August, 1950.) Summaries are given of 38 papers presented at the conference.

621.396.11 3133
Elements of Anomalous Radio Propagation—E. Knighting. (*Met. Mag.*, vol. 79, pp. 74–81; March, 1950.) Discussion of the dependence of the refractive index of the atmosphere on the pressure, temperature and amount of water vapor. The changes with altitude of these parameters can explain radio reception at distances beyond the normal optical range, by refractive bending of the ray paths and by the formation of radio ducts.

621.396.11 3134
Passage of Electromagnetic Waves [of Frequency], below the Critical Frequency, through Plasma Layers of Finite Thickness—W. O. Schumann. (*Arch. elekt. Übertragung*, vol. 4, pp. 173–174; May, 1950.) It is usually assumed that for frequencies below a certain critical frequency on em wave incident perpendicularly on a plasma layer of infinite extent will be totally reflected, an exponentially damped stationary wave being produced within the plasma. In the case of a plasma layer of finite thickness it is found that energy can be transported through the layer and appropriate formulas are derived.

621.396.11 3135
Reflection of Electromagnetic Waves at a Discontinuity of the Permittivity Gradient—G. Eckart. (*Funk u. Ton*, vol. 4, pp. 354–357; July, 1950.) A particular case, such as might occur in the troposphere, is considered and an expression is derived for the reflection coefficient (r). Unlike the Fresnel coefficient, r in this case is dependent on frequency, being smaller the shorter the wavelength. A table shows the angle of incidence, for values of λ from 1 cm to 1,000 m, at which $r=1/100$, for two values of gradient jump. See also 977 of May.

621.396.11:535.42 3136
On the Diffraction of a Radar Wave by a Conducting Wedge—R. B. Watson and C. W. Horton. (*Jour Appl. Phys.*, vol. 21, pp. 802–804; August, 1950.) "Diffraction patterns of radar waves have been measured about the edge of a perfectly conducting wedge. Theoretical patterns have been calculated using an asymptotic solution suggested by Pauli (906 of 1939). Good agreement is observed between experimental and calculated patterns. The thin wedge tested showed much similarity in diffracting properties to a suitable semi-infinite conducting screen."

621.396.11.029.51 3137
Polarization of Low-Frequency Radio Waves Reflected from the Ionosphere—A. H. Benner, C. H. Grace, and J. M. Kelso. (Proc. I.R.E., vol. 38, pp. 951–952; August, 1950.) Vertical-incidence measurements of the limiting polarization of 150 kc em waves, together with theoretical calculations, indicate that (*a*) observed split echoes are not due to magneto-ionic splitting; (*b*) the polarization is left-handed; (*c*) the wave leaves the layer at a level where the collision frequency is roughly 0.8×10^6 per second.

621.396.11.029.55 3138
Investigations on the Propagation of Short Waves over Great Distances—W. Messerschmidt. (*Arch. elekt. Übertragung*, vol. 4, pp. 181–188; May, 1950.) Report of experiments conducted in 1938 on the possibility of zigzag reflections in the transmission of sw signals over great distances. The transit time of a pulse transmitted and retransmitted over known distances was measured. Two fixed stations or one fixed station and one airborne retransmitting station were used; distances involved were chiefly about 8,000 km. The 'detour factor' (ratio of actual path length to great-circle distance) is correlated with the apparent height of the relevant ionospheric layer simultaneously measured at different places along the wave path. The transit time was often constant for long periods. At other times a jump from one value to another occurred; this is attributed to a variation in the number of zig-zag reflections.

621.396.11.029.55 3139
Studies of Multiple Circuits of the Earth by Short-Wave Signals—H. A. Hess. (*Fernmeldetech. Z.*, vol. 3, pp. 243–248; July, 1950.) Account of an investigation made on 19th November 1944, on a frequency of 19.947 Mc. Morse signals of duration 12 ms were sent out at 0.5-sec intervals from DLO (Berlin) and the echoes were observed at Randers (Denmark) and Gatow (Berlin). Results are shown graphically. The times for the second circuit lay between 0.1376 and 0.1384 per second. Periodic amplitude fluctuations of the echo were correlated with multiple paths and the movements of ionosphere layers. The high intensity maintained after repeated circuits is evidence of propagation in a narrow great-circle beam.

621.396.11.029.6 3140
The Propagation of Very Short Electromagnetic Waves—A. Grün. (*Z. Angew. Phys.*, vol. 2, pp. 294–301; July, 1950.) Approximate formulas derived by geometrical ray treatment are compared with more accurate formulas taking account of diffraction. In the decimeter range certain simple approximations are sufficient, since reflection and divergence factors do not differ greatly from unity, and it is safe to neglect both irregularities of the earth's surface and the fine variation of refractive index with height. Fluctuations of ray curvature due to weather conditions are important in this frequency range, producing a variable interference field which is the main cause of fading; this can be overcome by suitable choice of wavelength, antenna height, and range. The effects of disturbing reflections from isolated obstacles and of unusual refraction phenomena are discussed briefly.

621.396.11.029.64:535.42 3141
Diffraction Pattern in a Circular Aperture Measured in the Microwave Region—C. L. Andrews. (*Jour. Appl. Phys.*, vol. 21, pp. 761–767; August, 1950.) Measurements were made of the diffraction patterns of circular apertures (of diameter from λ to 8λ) in and near the aperture planes, for incidence of a plane-polarized em wave of 8 cm wavelength. From Young's theory the points of maximum intensity near the aperture were calculated and the results checked with experiment.

621.396.81:551.510.535 3142
Geometrical Optics of Ionospheric Propagation—K. Rawer. (*Nature*, (London), vol. 166, p. 316; August 19, 1950.) Three focusing effects have been observed, due respectively to refraction, curvature of the reflecting layer, and the spherical form of the earth. Neither the first nor the third of these effects can be applied regularly for ionospheric prediction. The method of constructing a prediction curve (field intensity/distance) to take account of the second effect is indicated. See also 514 of 1948, 717 of April and 1229 of June (Lejay and Lepechinski).

621.396.812.3 3143
Periodic Fading in Short Wave Propagation—H. A. Hess. (*Funk u. Ton*, vol. 4, pp.

3-340; July, 1950.) Photographic records of wave-trains of signals received from long-distance oversea transmissions in 1944 are reduced and analysed. Multiple-hop and round-the-world signals are considered in comparing ionospheric path lengths to determine rapidity of movement of the reflecting layer. The beat frequencies observed in certain received signals, in one instance a 34-cps frequency modulation of a 15-Mc signal, cannot be explained by vertical movement of the layer. In the region of the critical frequency, multiple reflections will not normally occur . fading is minimized. Similarly, periodic fading can be avoided by using strongly-modulated telegraphy signals.

.396.812.5 3144

Drift Phenomena for Rapid Fluctuations of Field Strength of Ionospheric Echoes—J. Hautkrämer. (*Arch. elekt. Übertragung*, vol. 4, pp. 133–138; April, 1950.) Observations obtained from three recorders in line indicate that drift of the reflection region may occur in the ionosphere. The magnitude and direction of the movement are determined from the difference in the time of observation of rapid fluctuations at three recording stations at the corners of a right-angled triangle. Velocities up to several hundred ms have been observed. In the *E* layer a definite directional trend is noted, but in the F layer, where higher velocities are more frequent, no such trend is apparent. The observations are explained by an interference effect of waves reflected from different irregularities in the layer. Preferred direction in the *E* layer is westward.

621.396.812.5 3145

Origin of the Mögel-Dellinger Effect—Siedentopf. (*Arch. elekt. Übertragung*, vol. 4, pp. 97–98; March, 1950.) Review of observations reported by various authors, and of possible causes of the effect.

621.396.812.5:523.746 3146

Results of Statistical Analysis of Fadeouts—D. Stranz. (*Arch. elekt. Übertragung*, vol. 4, pp. 217–218; June, 1950.) Correlation of sunspot data with 130 cases of total disappearance of radio signals observed during 1948.

RECEPTION

621.396.621 3147

The Broadcasting Receiver and the Pickup Connection—M. Lechenne. (*TSF pour Tous*, vol. 26, pp. 235–237; June, 1950.) Suggested additions to the receiver circuit to improve the quality of reproduction are: (*a*) at least 6 db nonselective negative feedback; (*b*) a special adjustable tone-control circuit with a linear potentiometer for the pickup position; (*c*) a 5- or 9-kc filter in the low-frequency amplifier anode circuit.

621.396.621:621.396.822 3148

Receiver Sensitivity and Its Limitation by Background Noise—J. Rousseau. (*TSF pour Tous*, vol. 26, pp. 218–221; June, 1950.)

621.396.621.54:621.396.662.4 3149

Calculation of the Three-Point-Alignment Circuit in Superheterodyne Receivers—O. Meisinger. (*Arch. elekt. Übertragung*, vol. 4, pp. 99–104; March, 1950.) The whole of the auxiliary circuit, comprising fixed capacitors, which in superheterodyne receivers makes three-point alignment possible, is treated as a quadripole. Calculation shows that the short-circuit and open-circuit impedances, seen from the variable capacitor, are given by polynomials involving three parameters. Practical formulas and tables, of general application, are given which enable all the necessary calculations to be carried out with sufficient accuracy on a slide rule for inductance and capacitance values, bandwidth, and the like.

621.396.621.57 3150

Receivers with Crystal Detectors ('Crystal-Video')—S. Marmor. (*Ann. Télécommun.*, vol. 5, pp. 266–275; July, 1950.) Results are presented of measurements made at the CNET on a silicon crystal of average characteristics at a wavelength of 10 cm. An explanation is given of the poor sensitivity of receivers using crystal detectors as compared with superheterodyne receivers. Difficulties encountered in constructing high-gain crystal-detector sets for pulse reception are described, and steps proposed by American designers (see 206 and 1638 of 1949) to prevent overshoots causing prolonged desensitization are reviewed. Details are given of a set designed at CNET, and photographs are shown of the output pulses obtained.

621.396.8.015.7†:621.396.822 3151

The Effect of a Video Filter on the Detection of Pulsed Signals in Noise—D. Middleton. (*Jour. Appl. Phys.*, vol. 21, pp. 734–740; August, 1950.) Theoretical examination of the effect of inserting a video filter of finite bandwidth on the observability of pulsed signals (of Gaussian or rectangular shape) in random noise. When the pulse and intermediate-frequency filter are matched, i.e., when they are each other's conjugate Fourier transforms, a video filter of infinite bandwidth gives the best results with weak signals. This is also the case when the spectral width of the pulses is greater than that of the intermediate-frequency filter, but a video filter of suitable bandwidth gives an improvement when the intermediate-frequency filter is wider than necessary for the length of pulse used. Curves are given showing the effect of the video filter on the signal-to-noise ratios in the different cases.

621.396.823+621.397.823 3152

Telephonic and Radio Interference from High-Voltage Systems—C. W. Marshall. (*Engineering* (London), vol. 170, p. 103; August 4, 1950.) Abridged version of report of the Internation Study Committee to the CIGRE, 1950. Cases of interference recorded by the BEA during 1949 are tabulated. 53 relate to radio and 28 to television interference

621.396.828 3153

Research Laboratories of the British Electricity Authority—J. S. Forrest. (*Nature* (London), vol. 166, pp. 334–335; August 26, 1950.) Investigations undertaken include research on high-voltage transmission using a line 800 yards long operated at 250–300 kv. The intensity of the radio interfering field from the line is being measured and comparisons are being made of the signal-to-noise ratios for AM and FM transmissions. Good television reception can be obtained with the antenna only 10 yards from the line conductors.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 3154

Recent Developments in Communication Theory—C. E. Shannon. (*Tech. Mitt. Schweiz. Telegr.-Teleph. Verw.*, vol. 28, pp. 337–342; September, 1950. In German.) Translation of article noted in 1764 of August.

621.39.001.11 3155

Introduction to Signal and Information Theories—J. Loeb. (*Ann. Télécommun.*, vol. 5, pp. 246–254; July, 1950.) A tentative synthesis of published studies in this field. Emphasis is laid on fundamental problems rather than on the theories developed. The discussion covers: (*a*) the analytical signal [1293 of 1948 (Ville)]; (*b*) definition of quantity of information for the case of discrete signals; (*c*) the case of continuous signals; (*d*) the relation between information theory and statistical physics.

621.39.001.11 3156

Multiplex and Theory of Communications—J. Icole. (*Ann. Télécommun.*, vol. 5, pp. 291–297; August, September, 1950.) The coefficient of utilization of a radiotelephony channel is defined and its value is calculated approximately for various cases of FM and PM. Its variation with different parameters is determined and shown graphically. The advantage of using several interdependent channels on a single carrier is demonstrated.

621.396.41 3157

Comparison between Multichannel Systems with Common and Multiple Carriers—P. Güttinger and G. Valko. (*Brown Boveri Rev.*, vol. 36, pp. 396–401; December, 1949.) From a theoretical standpoint, the difference between common-carrier and multiple-carrier multichannel systems lies mainly in the causes of crosstalk. With common-carrier systems, both linear distortion in the filters and nonlinearity of the modulation and demodulation processes are responsible, whereas with multicarrier systems, nonlinearity of mixer and output stages is the main cause. Experimental multichannel multicarrier equipment is described, and the most essential technical and test data are given.

621.396.619.11/.13:621.397.61 3158

On the Simultaneous Amplitude and Frequency Modulation of a Carrier Wave—F. Kirschstein. (*Funk u. Ton*, vol. 4, pp. 282–292; June, 1950.) Experiments made in 1939 consisted in modulating the amplitude of a low-power, 200-kc carrier with speech currents and at the same time transmitting a coded signal by keying the carrier to produce a tone of a few hundred cps. No interference was observed. Using a similar circuit the AM carrier was then modulated in frequency by four telegraphy tones of frequencies 120, 360, 600 and 840 cps. With AM above 80 per cent, slight distortion of the FM tones occurred. Using two speech channels the FM channel was affected by crosstalk. Application of the method for double-side-band television transmission is discussed, the frequency of the vision-signal carrier being modulated to provide the sound channel.

621.396.619.14 3159

Product Phase Modulation and Demodulation—D. B. Harris. (Proc. I.R.E., vol. 38, pp. 890–895; August, 1950.) A modulating voltage l_m impressed in parallel on sine and cosine phase converters produces at their outputs voltages proportional to the functions $\sin kl_m$ and $\cos kl_m$. With two balanced modulators, these modulate respectively the carrier wave, $\cos \omega t$, and the carrier displaced in phase by 90°, $\sin \omega t$, the outputs being added to produce a phase-modulated wave according to the identity $\cos kl_m \cos \omega t + \sin kl_m \sin \omega t = \sin (\omega t - kl_m)$. The phase converters proposed are oscilloscope tubes with the fluorescent screen replaced by an anode for collecting electrons. A mask is interposed in the beam to produce a voltage proportional to the sine or cosine of the linear beam deflection, which in turn is proportional to the modulating voltage applied to the deflection plates. Phase deviations up to $\pm 25\pi$ radians should be obtainable.

621.396.65 3160

Planning Radio Relays—W. Steinmann. (*Brown Boveri Rev.* vol. 36, pp. 410–414; December, 1949.) Discussion of the various factors which must be taken into account.

621.396.65 3161

Fundamentals for the Planning of Beam Radio Links—J. Kornfeld. (*Öst. Z. Telegr. Teleph. Funk Fernschtech.*, vol. 4, pp. 85–91; July, August, 1950.)

621.396.65 3162
A Multichannel Metre-Wave Radio Link with Terminal Stations Outside the Optical Range—A. Heilmann. (*Fernmeldetech. Z.*, vol. 3, pp. 221–233; July, 1950.) Experimental radiotelephone links installed between Berlin and the Western Zone of Germany in and since 1948 are described. Selection of terminal sites in relation to path configuration is discussed; useful results can be expected even when neither terminus is at high elevation. FM is used. A link operating on 64 Mc in one direction and 68 Mc in the other, with a frequency deviation of ±75 kc, gave improved quality and less lost time as compared with earlier links on lower frequencies and with smaller frequency deviations. Receiver input voltages have been recorded continuously for some months, and charts are shown and discussed. An appendix develops formulas for the nonlinear distortion.

621.396.65 3163
Wave Propagation Studies in the Alps—W. Klein. (*Brown Boveri Rev.*, vol. 36, pp. 387–395; December, 1949.) See 2326 of 1949.

621.396.65:621.396.5 3164
Multichannel Common-Carrier Radiocommunication System—W. Zimmermann. (*Brown Boveri Rev.*, vol. 36, pp. 373–378; December, 1949.) A description of the equipment used in the Zürich/Geneva radiotelephone relay. The system uses common-carrier multichannel transmission, a single carrier wave being modulated in frequency by an intermediate-frequency signal derived from the frequency-shifted channel signals. Independent 'go' and 'return' channels are used. The frequency range is 150–220 Mc and the output of the terminal and repeater transmitters is 50 w. Robust corner reflectors with folded dipoles are used for transmission and for reception. Stand-by equipment is provided.

621.396.65:621.396.5 3165
Multiplex Radiotelephone Relays with Pulse-Modulated Microwaves—H. J. v. Baeyer. (*Brown Boveri Rev.*, vol. 36, pp. 379–386; December, 1949.) Description of a 2-way radiotelephone relay operating on frequencies between 1.9 and 2.1 kmc and providing 23 telephony channels and one service channel. Pulse-time modulation is used for the speech channels while width modulation is applied to the service-channel pulses. The radio-frequency pulse power is 30–50 w for a duty cycle of 1:5 and the parabolic reflectors commonly used have a gain of about 25 db. With the attenuation figures recommended by the CCIF, the alternating-frequency band extends from 300 to 3,400 cps.

621.396.65:621.396.619.16 3166
Methods of Modulation with Frequency Sharing and with Time Sharing (Pulse Modulation) for Multichannel Directional Radio Links—H. Holzwarth. (*Frequenz.*, vol. 4, pp. 33–40; 64–71, and 97–101; February-April 1950.) The principles of various types of modulation are described and the improvement in signal-to-noise ratio obtainable by the different systems is investigated. Using double FM or PPM, crosstalk between channels may be kept slightly lower than with FM, which is the best method for single-channel working. Amplitude spectra are given for most of the known modulation systems, PAM and PWM being particularly considered, since these occur in the end stages of most PM systems. The distortion is calculated for each method of modulation; with more than 6 channels distortion is normally extremely small.

621.396.65:621.396.822 3167
The Signal/Noise Ratio in a Decimetre-Wave Relay Chain—P. Barkow. (*Arch. elekt. Übertragung*, vol. 4, pp. 155–158; April, 1950.) Measured values are compared with values calculated from three different formulas.

621.396.65:621.397.5 3168
A Portable Microwave Television Radio Link—(*Engineer* (London), vol. 190, p. 128; August 4, 1950.) Developed by Standard Telephones and Cables for use in a motor van, this FM equipment operates on a frequency of about 4,000 Mc with a radiated power of 300 mw. The transmitting and receiving circuits are mounted in weather-proof canisters on their associated paraboloid antennas and are respectively connected by coaxial cable to a control and monitor unit, and to a seven-stage intermediate-frequency amplifier and a discriminator unit within the van.

621.396.65:621.397.5 3169
The London/Castleton Experimental Radio-Relay System—A. H. Mumford, C. F. Booth, and R. W. White. (*P. O. Elec. Eng. Jour.*, vol. 43, Part 2, pp. 93–99; July, 1950.) A wide-band FM system for the transmission of television signals. The mean carrier frequency is 195 Mc and four intermediate relay stations are used. To minimize feedback at the relay stations, the receiving and transmitting equipments are located on opposite sides of suitable hills and connected by land line. Rhombic antennas with a forward gain of 15 db with respect to a λ/2 dipole and with a front-to-back ratio of 25 db also help to obtain an attenuation between the transmitting and receiving aerials at a repeater station exceeding 120 db. The terminal- and intermediate-station circuits are described and an indication is given of the quality of the transmission of pulse and television signals. See also 2026 of September (Mumford and Booth).

621.396.65.029.63 3170
A Wide-Band V.H.F. Radio-Relay System—W. S. McGuire. (*AWA Tech. Rev.*, vol. 8, pp. 295–307; June, 1950.) See 745 of April.

621.396.712:534.86 3171
Audio-Frequency Equipment for Broadcasting Services—J. E. Telfer. (*AWA Tech. Rev.*, vol. 8, pp. 309–340; June, 1950.) See 2027 of September.

621.396.931:621.396.619.13 3172
Mobile Radio Equipment, Type SRR 192—D. J. Braak. (*Commun. News*, vol. 10, pp. 120–125; April, 1950.) Two pre-set frequencies are available in the bands 70–78 Mcs and 80–86 Mc. The frequencies are crystal controlled and FM is used.

621.396.932 3173
The Port of Liverpool V.H.F. Radio System—(*Electronic Eng.*, vol. 22, pp. 328–330, 337; August, 1950.) See also 2633 of December.

621.396.932.004.15 3174
Radio for Merchant Ships. Performance Specification for Lifeboat Portable Radio Equipment [Book Notice]—Publishers: H. M. Stationery Office, London, 4d. (*Govt. Publ.* (London), p. 20; July, 1950.)

SUBSIDIARY APPARATUS

621.316.7.078 3175
Contribution to the Study of Automatic Regulation—M. Cuénod. (*Bull. Schweiz. Electrotech. Ver.*, vol. 41, pp. 673–678; September 2, 1950. In French, with German summary.) The methods of operational calculus are used to study the variations of the regulated quantity following on a disturbance. These variations depend not only on the dynamic characteristics of the regulator, but also on other elements, such as the exciter and generator in the case of voltage regulation. The use of Nyquist curves (response functions) facilitates theoretical investigation of conditions for stability and of variations of the regulated quantity, thus permitting comparison with results obtained in practice.

621.316.722.1 3176
Voltage Stabilizer for 200-kV Acceleration—J. T. Dewan. (*Rev. Sci. Instr.*, vol. 21, pp. 771–772; August, 1950.) A stabilization ratio of 40 is achieved by the use of a 2-stage direct-coupled feedback amplifier, controlling a series regulating pentode. The error signal applied to the amplifier is the difference between a stable balancing voltage and a fraction of the main output, tapped off a 250-MΩ wire-wound bleeder resistor. Standard regulating devices are used for the supply voltages and pentode heater current to minimize amplifier drift.

621.355 3177
Electrical Accumulators: Recent Patents—L. Jumau. (*Rev. Gén. Élec.*, vol. 59, pp. 323–343; August, 1950.) A review of developments in acid and alkaline types of cell. Thermoplastics are being used more and more for cases, separators, frames, etc.

771.36 3178
A 100000000-Frame-per-Second Camera—M. Sultanoff. (*Rev. Sci. Instr.*, vol. 21, pp. 653–656; July, 1950.) High photographing rates are obtained with a multislit focal-plane shutter which is transported optically across the film plane by means of a rotating mirror. Application is to self-luminous phenomena such as are associated with explosions.

TELEVISION AND PHOTOTELEGRAPHY

621.396.823+621.397.823 3179
Telephonic and Radio Interference from High-Voltage Systems—Marshall. (*See* 3152.)

621.397.335 3180
The Effect of Interference on the Synchronization of [television] Timebases, and the Remedies to be Applied—H. Somers. (*Rev. belge Élec.*, vol. 2, pp. 3–8; May, June, 1950.) Synchronization is liable to be destroyed by interference, particularly when negative modulation is used. This disadvantage can be avoided by providing a local control voltage to maintain correct phasing of the scanning. Particular systems discussed are: direct control of the sawtooth oscillations; automatic phase control of a sine-wave oscillator by means of a reactance tube; flywheel circuits. See also 2336 of 1949 (Clark).

621.397.5 3181
TV in England—J. Moir. (*FM-TV*, vol. 10, pp. 12–13, 38; August, 1950.) A report on progress under BBC-controlled development.

621.397.5:535.241.44 3182
Brightness and Contrast in Television—P. C. Goldmark. (*Jour. Brit. I.R.E.*, vol. 10, pp. 219–225; June, 1950.)

621.397.5:621.396.65 3183
A Portable Microwave Television Radio Link—(*See* 3168.)

621.397.5:621.396.65 3184
The London/Castleton Experimental Radio-Relay System—Mumford, Booth, and White. (*See* 3169.)

621.397.5(083.74) 3185
Standards for Television Systems—(*Nature* (London), vol. 166, pp. 472–473; September 16, 1950.) After examining the state of television in the United States and in various European countries in March and April, 1950, a study group of the CCIR recommended the following points to the main committee: (*a*) systems should be independent of power supply frequency; (*b*) picture aspect ratio should be 4:3; (*c*) line interlacing should be used in the

tio 2:1; (*d*) asymmetric sideband transmis- ›n should be adopted for the vision signal; · standardization of the polarization of the dio transmission is not necessary. Present line ındards were retained by France, Great ·itain and the United States. See also 2642 of ›cember.

1.397.6 3186
Three-Colour Television System with Faith- l Reproduction—F. Lachner. (*Öst. Z. Telegr. ›leph. Funk Fernsehtech.*, vol. 4, pp. 63–70 and –103; May, June, and July, August, 1950.) ›lor theory is dealt with at some length and e method of representing a color stimulus by ichromatic coordinates is explained. When ›plied to color television the analysis leads to e requirement at the transmitter of color ters whose frequency response curves have a ›gative portion between two positive ones. In ·actice this condition can be simulated by bstituting for the negative-response filter one ith a positive response of equal absolute agnitude and reversing the current from the sociated pickup tube. To make the proper gnal additions possible, three pickup tubes, ›ch with associated partial filter, are required r each primary color. Cross-controlled sys- ms, i.e., where fractions of the signal corre- ›onding to one color contribute to the control the other two colors at the receiver, are also scussed.

!1.397.6:535.88 3187
Recent Progress in Large-Screen Projec- ›n of Television Pictures—A Cazalas. (*Ann élécommun.*, vol. 5, pp. 298–306; August, ›ptember, 1950.) Developments in cinema ›levision are surveyed, both intermediate-film ıd projection methods being considered. A rojection tube is described in which the uorescent screen is excited continuously by n auxiliary cathode a little distance from and f about the same area as the screen. Passage f the scanning beam across this cathode pro- uces point-to-point variations of its emission nd hence of screen brightness.

21.397.61:621.396.619.11/.13 3188
On the Simultaneous Amplitude and Fre- uency Modulation of a Carrier Wave— Kirschstein. (*See* 3158.)

21.397.61:621.396.619.24:621.317.1 3189
Television Transmitter Lower-Sideband Measurements—Hamilton and Artman. (*See* 086.)

21.397.62 3190
The TV5 De Luxe—A. V. J. Martin. *Television*, pp. 133–140; July, August, 1950.) Detailed design for a reliable, high-quality tele- ision receiver, presented as suitable for manu- acturers who do not run a television research epartment of their own.

21.397.645 3191
Television Intermediate-Frequency Ampli- ers—A. Coenraets. (*Rev. Belge Élec.*, vol. 2, ›p. 9–14; May, June, 1950.) General considera- ions of bandwidth, gain and frequency re- ponse are discussed, typical circuits are de- cribed and practical hints on screening are iven.

21.397.8 3192
Flicker Color Fringing Phenomena in Color Television—P. C. Goldmark. (*Jour Brit. I.R.E.*, vol. 10, pp. 208–217; June, 1950.) The etermination of the threshold of flicker and he maximum permissible highlight illumina- ion in the sequential color system are dis- ussed. Flicker would be inappreciable at a olor-frame frequency of 60 per second, but a requency of 48 per second has been adopted lue to bandwidth considerations, giving a hreshold of flicker at 23 foot-lamberts for an observation distance seven times the picture height. In a European 50 per second system this value would increase to 40 foot-lamberts. Tests with the field-sequential system have proved that color fringing is of rare occurrence.

621.397.8 3193
95-Mile TV Reception—R. F. Allison. (*FM-TV*, vol. 10, pp. 14–15; August, 1950.) A report on results obtained under extremely adverse conditions.

621.397.62 3194
The Principles of Television Reception [Book Review]—A. W. Keen. Publishers: Pitman and Sons, London, 1950, 319 pp., 30s. (*Jour. Brit. I.R.E.*, vol. 10, p. vii; August, September, 1950.) The book is addressed to the technician rather than the engineer and is nonmathematical. Numerous references facilitate further study, and both British and American practice are described.

TRANSMISSION

621.396.619.029.64:621.396.611.4 3195
Microwave Modulation by Variable Circuit Elements Comprising Waveguides or Cavity Resonators—A. N. Bhattacharyya. (*Indian Jour. Phys.*, vol. 23, pp. 175–183; April, 1949.) A driving mechanism similar to that of an electrodynamic loudspeaker is used to vary the dimensions of a waveguide or cavity resonator according to the speech signals, so that microwaves passing through the waveguide may be modulated in either amplitude or phase or the output of the cavity resonator modulated in amplitude.

TUBES AND THERMIONICS

537.533:538.3.029.6 3196
Quantum Effects in the Interaction of Electrons with High Frequency Fields—J. C. Ward. (*Phys. Rev.*, vol. 80, p. 119; October 1, 1950.) Treating the electron as quantized and the field as classical, the phase-integral approximation is used to predict the probability of energy transfer between the microwave field inside a resonator and electrons projected through the field. When the number of quanta transferred is large, the result tends to that obtained by classical theory.

539.233:535.215 3197
Photoadsorption of a Thin Layer of Barium—Biguenet. (*See* 3030.)

621.315.59:621.385 3198
A Crystal Amplifier with High Input Impedance—O. M. Stuetzer. (PROC. I.R.E., vol. 38, pp. 868–871; August, 1950.) An amplifier is described consisting of a point- or line-type metal/semiconductor contact, in the neighborhood of which a high electric field can be applied between an additional electrode and the crystal surface. This can be used to control the current flowing through the rectifying contact. The device has been termed a 'fieldistor.' ac response, amplification, noise and dc measurements show that the device has a high input impedance and appears useful as an impedance transformer with an appreciable current and power gain at lower frequencies.

621.383.42 3199
Constitution and Mechanism of the Delenium Rectifier Photocell—J. S. Preston. (*Proc. Roy. Soc. A*, vol. 202, pp. 449–466; August 22, 1950.) Experiments are described which indicate the structure and mode of operation of the thin surface film in the Se photocell. Preliminary results on films of Au, Al and Cd, prepared by evaporation in a high vacuum, established the importance of CdO as a film material. A technique is described for sputtering thin films of this compound having high light-sensitivity and conductivity, and open-circuit voltages up to 0.5 v under high illumination. Double films of ZnO and Au were also investigated; these gave a lower sensitivity. It is suggested that in the practical photocell, the essential feature is a contact between two suitable semiconductors of opposite types.

621.385.001.4(083.74) 3200
Standards on Electron Tubes: Methods of Testing, 1950—(*See* 3103.)

621.385.029.62/.63 3201
A Survey of Modern Radio Valves: Part 5—Valves for Use in the Range 30–3,000 Mc/s—R. W. White. (*P.O. Elec. Eng. Jour.*, vol. 43, Part 2, pp. 78–84; July, 1950.) The poor performance of normal tubes at the higher frequencies is discussed and the more important contributory factors are indicated. Desirable design features are tabulated and examples of tubes specially designed for use in the above range are briefly described. Part 4: 2058 of September.

621.385.029.63/.64+621.396.615.14 3202
Transmitting Valves for Decimetre and Centimetre Waves—F. M. Penning. (*Commun. News*, vol. 10, pp. 109–118; April, 1950.) A review of developments resulting from the application of transit-time effects, the use of cavity resonators, the transfer of energy to the high-frequency field by induction, and the use of pulse methods.

621.385.029.63/.64 3203
Traveling-Wave Tubes: Part 3—J. R. Pierce. (*Bell. Syst. Tech. Jour.*, vol. 29, pp. 390–460; July, 1950.) Chapter VII develops equations which sum up, in terms of the more important parameters, the linear operation of travelling-wave tubes for small values of the gain parameter. The effect of the various parameters on the rate of increase and velocity of propagation of the three forward waves is dealt with in the next chapter. Chapter IX discusses the effect of introducing losses into the circuit and of severing the helix. In chapter X the factors affecting the noise figure are discussed and the conditions for low noise are given. The possibility of the amplification of a nearly synchronous backward-travelling wave is investigated briefly in chapter XI. Appendices deal with the evaluation of the space-charge parameter Q, diode equations, the evaluation of impedance and Q, for thin and solid beams, and the calculation of gain. Part 1: 1810 of August. Part 2: 2377 of November.

621.385.029.63/.64 3204
The Travelling-Wave Valve—L. Brück. (*Arch. Elektrotech.*, vol. 39, pp. 633–647; 1950.) Principle, construction, and properties are described. Details are given for correct design of the electron gun. Alternative delay-line systems are discussed and the influence of voltage, current and helix dimensions on amplification and available high frequency output power are demonstrated by reference to measured characteristics and simplified equations. The possibility of self-excitation and its suppression are considered.

621.385.029.63/.64 3205
The Efficiency of the Travelling-Wave Valve—O. Döhler and W. Kleen. (*Arch. elek. Übertragung* vol. 4, pp. 207–212; June, 1950.) Continuation of previous work (2669 of 1949). Expressions are derived for electronic efficiency, circuit efficiency and total efficiency. Circuit efficiency can be calculated exactly, but the formula for electronic efficiency is based on approximations. These are justified by the similarity of results obtained by different methods of investigation.

621.385.029.63/.64 3206
On the Operation of the Travelling Wave Tube at Low Level—R. Kompfner. (*Jour. Brit.*

I.R.E., vol. 10, pp. 283–289; August, September, 1950.) The small-signal theory of the travelling-wave tube is developed on the assumption of finite distributed attenuation in the helix and beam voltages other than the 'synchronous' voltage. On restricting the gain parameter z to values <2, a simplified analysis gives results not more than about 6 per cent in error. The voltage gain is then derived for the first 'zero-gain' point, when the tube produces infinite attenuation of an input signal. Under these conditions a measurement of beam current enables an estimate to be made of the effective impedance of the helix, a parameter of considerable importance in design. Theoretical and experimental results are presented.

621.385.032.213 3207
Boride Cathodes—J. M. Lafferty. (*Phys. Rev.*, vol. 79, p. 1012; September 15, 1950.) The thermionic emission properties of the borides of the alkaline-earth and rare-earth metals and thorium have been investigated. These compounds all have the same formula MB_6 and the same crystal structure consisting of a 3-dimensional boron framework in whose interlattice spaces the metal atoms are embedded. The valence electrons of the metal atoms are not accepted by the B_6 complex, thus releasing free electrons which impart a metallic character to these compounds. LaB_6, for example, has a specific resistance of 27 μΩ cm at room temperature and a positive resistance temperature coefficient of 0.071 μΩ/°C. The compounds are very refractory, with melting points above 2100°C, and they are very stable chemically; moisture, oxygen and even HCl do not react with them. At a sufficiently high temperature, surface evaporation of the metal atoms occurs, but is compensated by diffusion from underlying cells so that an active surface is constantly maintained. This feature, together with the high electrical conductivity and high thermal and chemical stability, makes such materials ideal for cathodes. The rare-earth borides have better emission properties than the others, LaB_6 giving the highest emission, nearly twice that of BaB_6 and over 10 times that of CaB_6. Boride cathodes require no special activation; when heated for a few minutes at 1,400–1,600°C for outgassing, they are found to be completely active. Pulsed emission is the same as the dc emission. Owing to diffusion of the boron atoms into the lattices of refractory metals normally used as cathode bases, and consequent collapse of the boron framework, tantalum carbide or carbon should be used as the support for boride cathodes.

621.385.032.216 3208
The Conduction Mechanism in Oxide-Coated Cathodes—R. Loosjes and H. J. Vink. (*Philips Res. Rep.*, vol. 4, pp. 449–475; December, 1949.) A full account of the work noted in 491 of March 1. If σ is the conductivity of the oxide coating, *T* the absolute temperature, *I* the current (mA/mm²) and *V* the voltage applied across a 100-μ coating, the curves for log σ against 1/*T* and for *I* against *V* can be divided into three parts: (*a*) below 800°K, (*b*) 800–1,000°K, (*c*) above 1,000°K. The log σ curves have low slope in ranges (*a*) and (*c*) and are steeper in range (*b*), while the *I*/*V* curves are linear in ranges (*a*) and (*c*) and curved in range (*b*). These results are explained by postulating two conduction mechanisms operating in parallel, one being the electronic conduction through the particles of the oxide coating, the other the conduction through the electron gas in the pores of the coating. A calculation based on the known emission of a well-activated cathode gave a value for the conductivity of the same order of magnitude as that determined experimentally.

621.385.2/.3 3209
Vacuum Tubes with Mutually Bombarding Oxide Cathodes—E. G. Hopkins. (*Jour. Appl. Phys.*, vol. 21, pp. 841–842; August, 1950.) A diode with closely spaced oxide electrodes can be made to pass ac after the removal of the initial heating source, while the same structure with an interposed grid can similarly be used as a triode oscillator. The power usually wasted in anode dissipation can thus be used to heat a cathode surface, and practical power tubes are envisaged working on this basis.

621.385.2/.4:621.315.59 3210
Transistor and Transistron—M. Adam. (*Tech. Mod.* (Paris), vol. 42, pp. 220–224; July 1–15, 1950.) Design and application developments of the American and French versions of Ge-crystal tubes are briefly reviewed. The importance of avoiding confusion with the transitron tetrode oscillator is emphasized. See also 2978 of 1949.

621.385.3.029.64 3211
416A Tube for Microwave Relays—K. P. Dowell. (*FM-TV*, vol. 10, pp. 20–22; August, 1950.) A planar triode with high gain-bandwidth product at 4,000 Mc is described. New manufacturing and assembly techniques are necessary on account of the close interelectrode spacings.

621.385.38 3212
Ignition and Discharge Processes in Gas Triodes—E. Knoop and W. Kroebel. (*Z. Angew. Phys.*, vol. 2, pp. 281–285; July, 1950.) Discharge processes in various gases were investigated by means of an oscillograph. The discharge was recorded photographically with time scales of 1.7×10^{-8} to 3×10^{-9} second per mm of oscillograph trace. From these records the exponential indices of the discharge rise and fall curves were found. With grid ignition the cathode/anode discharge is delayed by a fixed interval after the grid/anode discharge. For Type-EC50 tubes (helium-filled) the delay time is 10^{-8} second; for Type-4690 tubes (argon-filled) about 5×10^{-8} second. The difference is attributed to the effect on transit time of the greater mass of the argon ions. No fluctuations were observed in the time of occurrence of ignition or discharge.

621.385.832:621.318.572 3213
Valves with a Ribbon-Shaped Electron Beam: Contact Valve; Switch Valve; Selector Valve; Counting Valve—J. L. H. Jonker. (*Philips Res. Rep.*, vol. 5, pp. 6–22; February, 1950.) A discussion of results obtained in the development of special cathode-ray tubes. The size of such tubes can be reduced so much that customary tube techniques can be applied in their construction. The new possibilities thus created are illustrated by (*a*) an electronic contact tube, which may serve as a telephony switch, (*b*) a tube of similar design to replace magnetic relays, (*c*) a multicontact switch tube, and (*d*) a high-speed counting tube. See also 1801 of August.

621.396.615.141.2 3214
A Simple Approximate Formula for the Resonance Frequency of a Cavity Magnetron—F. Borgnis. (*Z. Angew. Phys.*, vol. 2, pp. 278–280; July, 1950.) The magnetron multicavity system is usually considered as a number of similar coupled oscillatory systems. It is here treated as a cylindrical waveguide with plane metal ends, the actual boundary of the cavity coinciding approximately with the hypothetical surface containing the magnetic field lines of the corresponding circular-section waveguide. A very simple formula is thus obtained for the desired (π-mode) resonance frequency. A numerical example is worked for λ=1 cm.

621.396.615.142 3215
Theory of Velocity-Modulation Transit-Time Valves: Parts 3 & 4—H. Döring. (*Arch. elekt. Übertragung*, vol. 4, pp. 147–153, and 223–231; April and June, 1950.) The commencement of oscillations is considered for the theoretical case of a tube with infinitely short alternating fields (electrical double layer). The lower part of the general efficiency curve is represented by a parabolic approximation, which is applied to determine the conditions necessary for oscillations to begin. The practical case of tubes with alternating fields of finite length is then treated and conditions for oscillation are determined for single-field tubes, drift tubes with in-phase or out-of-phase alternating fields, and reflex tubes. Part 1: 2175 of 1944. Part 2: 1034 of May.

621.385+621.396.611.1.]029.6 3216
Technique des ondes tres courtes et ultra-courtes: Tome I—Circuits oscillants et tubes a vide [Book Review]—Liot. (*See* 3017.)

MISCELLANEOUS

53 3217
Physics Research at T.R.E—R. A. Smith. (*Science*, vol. 112, pp. 71–73; July 21, 1950.) Work on electronics, millimeter waves, semiconductors, infrared radiation, radio meteorology, low temperature and theoretical physics is briefly described.

621.396.69:061.4 3218
Radio at the 39th Paris Fair—R. Ar. (*Radio Prof.* (Paris), vol. 19, pp. 15–26; June, 1950.) General survey of exhibits, with notes on individual receivers (radio and television), antennas, loudspeakers, victrolas, intercommunications equipment and electronic apparatus.

621.3.089.6+53.001.4 3219
Testing by the National Bureau of Standards NBS Circular 483 [Book Review]—National Bureau of Standards. Publishers: U. S. Government Printing Office, Washington, 1949, 25 cents. (*Engineering* (London), vol. 170, p. 102; August 4, 1950.) This document states the Bureau's policy regarding testing, outlines the procedure for applying to have tests carried out, and sets out in detail the types of test which can be made and the schedule of fees.

621.39.029.64 3220
Les Hyperfréquences, Circuits et Propagation des Ondes (en vue de l'Application au Radar et aux Telecommunications) [Book Review]—R. Rigal. Publishers: Ed. Eyrolles, Paris, 1950, 219 pp., 1470 fr. (*Ann. Télécommun.*, vol. 5, p. 276; July, 1950.) The author is Directeur des Études à l'Ecole Nationale Supérieure des Télécommunications, and the book covers his course for engineering students.

621.396 3221
N.A.B. Engineering Handbook, Fourth Edition [Book Review]—Publishers: National Association of Broadcasters, Washington, D. C. 1949, 650 pp., $17.50. (Proc. I.R.E., vol. 38, p. 962; August, 1950.) "The material in part consists of reprints of articles appearing in various technical magazines, and in part of articles directly written for this handbook. The style and content is therefore variable. . . . It is recommended to the attention of all engineers who are engaged in broadcast and television activities."

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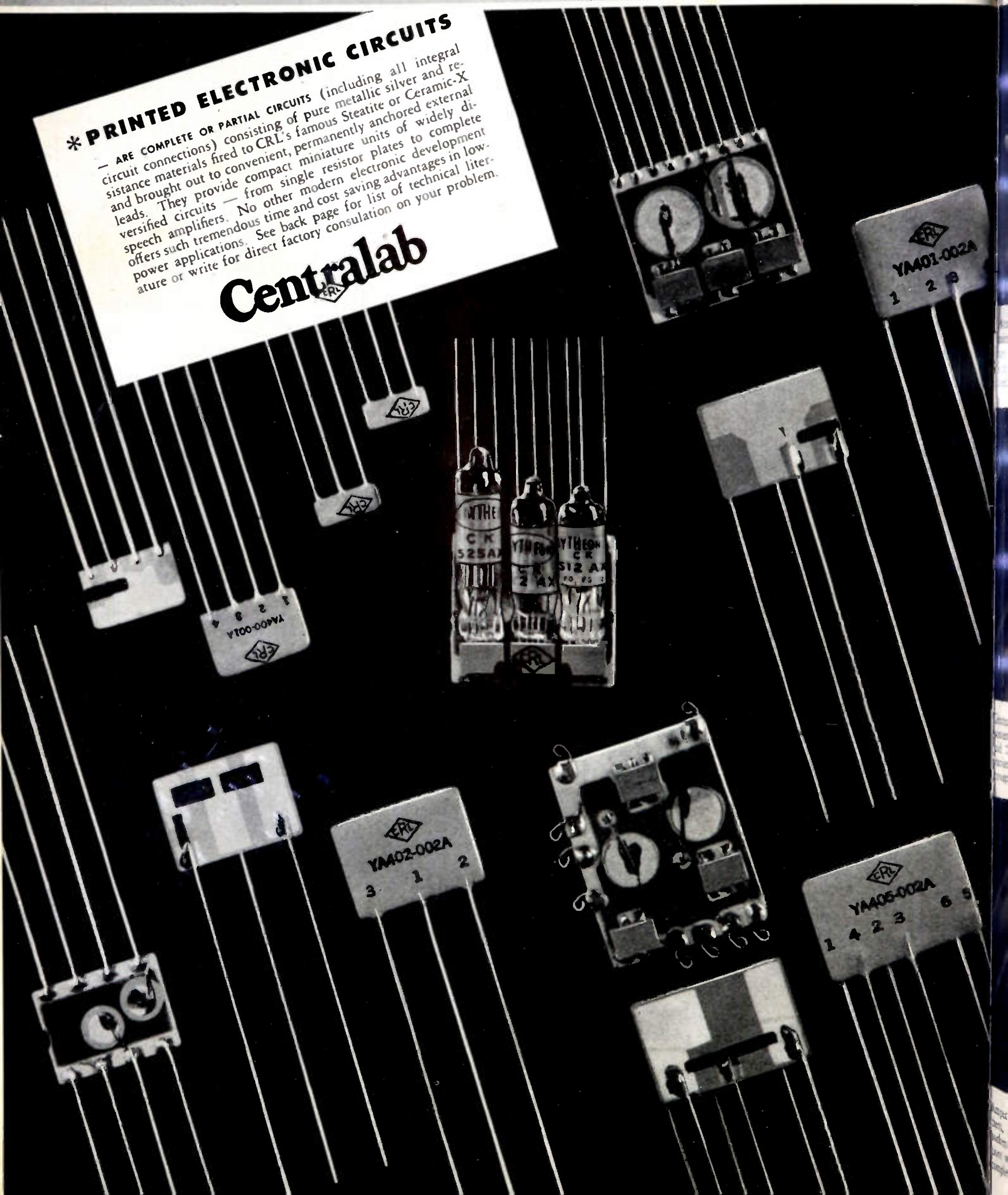

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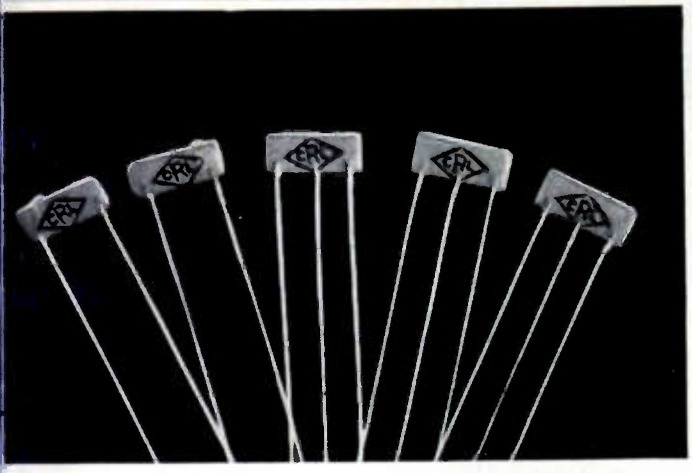

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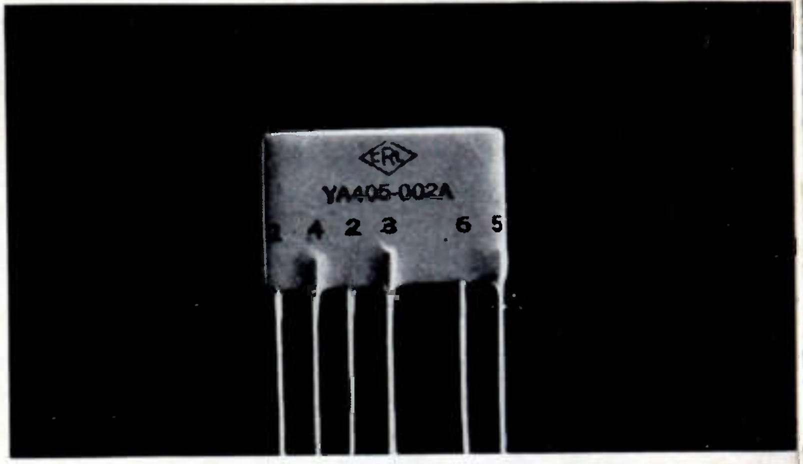

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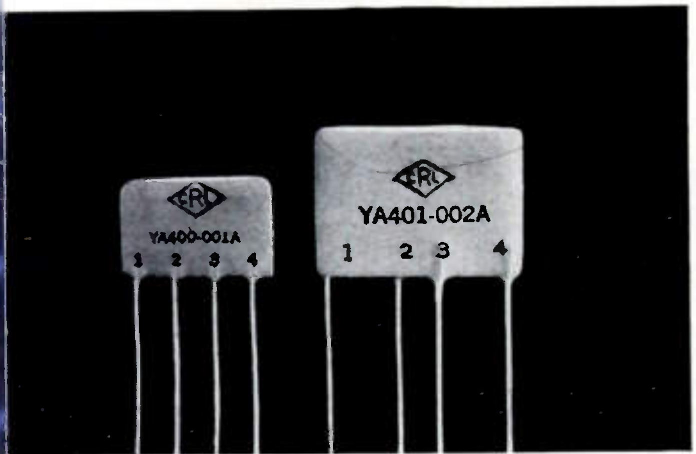

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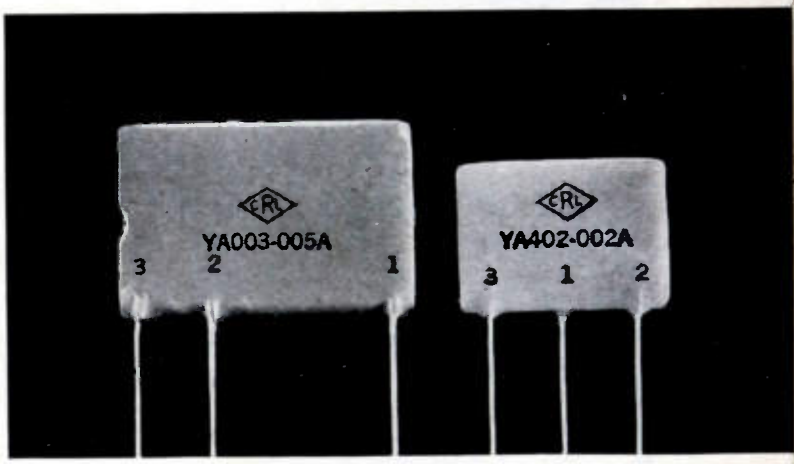

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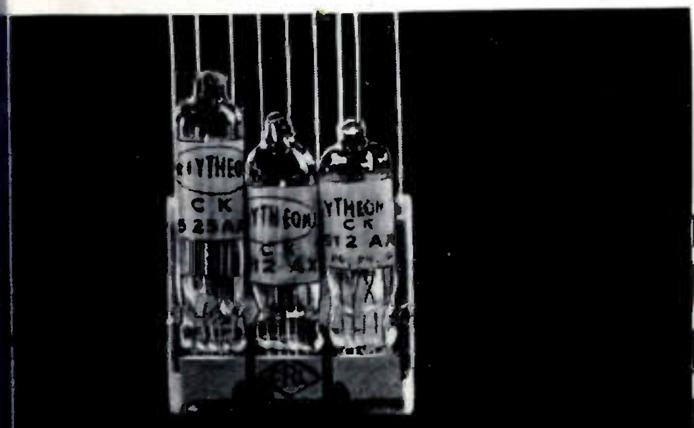

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SECTION MEETINGS

Atlanta

"Measurement Units Used in Communication Engineering," by N. B. Fowler, American Telephone and Telegraph Company; October 27, 1950.

Baltimore

"A Precise Continuously Variable Pulse Delay Unit," by J. F. Gordon; November 8, 1950.

Boston

"Engineering Aspects of International Short-Wave Broadcasting Conference," by C. B. Plummer, Federal Communications Commission; September 14, 1950.

"Research With High-Energy Electrons," by J. G. Trump, Faculty, Massachusetts Institute of Technology; October 19, 1950.

"Radar in Meteorology," by A. C. Bemis, Faculty, Massachusetts Institute of Technology; November 16, 1950.

Buffalo-Niagara

"The Effect of Recent Color Television Developments," by Nathan Marchand, Sylvania Electric Products Inc.; October 18, 1950.

Cedar Rapids

"Home Problems that Challenge Us Today," by Lillian Gilbreth; October 17, 1950.

"Improvements in Transformer Performance," by Ruben Lee, Westinghouse Electric Corporation; October 24, 1950.

Chicago

"Two-Way Radio Communication in Large Industrial Plants," by G. B. Soviers, Westinghouse Electric Corporation; "Quality Control in the Electronics Industry," by Harold May, Motorola, Fred Thoorobridge, Sentinel, and Dan Harris, Sentinel; September 15, 1950.

"Transcontinental Television," by B. D. Wickline, Illinois Bell Telephone Company; "UHF Receiver Design and Field Experience," by J. F. Bell, Zenith Radio Corporation; October 20, 1950.

Cincinnati

"Electronic Umpire," by R. F. Shea, General Electric Company; October 17, 1950.

Columbus

"New Test Equipment for Pulse, UHF, and Audio Measurements," by F. E. Waterfall, Alfred Crosley and Associates; October 19, 1950.

"Industry's Future," by Tom Campbell, Editor, *Iron Age*; November 2, 1950.

Connecticut Valley

"Electronics in Navigation Control," by R. L. Wanamaker, General Electric Company; Film: "V-2 Rocket Project"; October 19, 1950.

Dallas-Fort Worth

"Theory and Application of the Vibrotron," by John Ohman, Southwest Research Institute; October 25, 1950.

Denver

"Nuclear Resonance by Super-Regenerative Technique," by J. R. Zimmerman, Faculty, University of Colorado; "Methods of Investigating Beta-Ray Spectra with Magnetic Spectrometers," by A. A. Bartlett, Faculty, University of Colorado; November 10, 1950.

Des Moines-Ames

"Bell System Intercity Television Networks," by M. E. Strieby, American Telephone and Telegraph Company; September 26, 1950.

Evansville-Owensboro

"The FM Ring Antenna," by Harvey Kees, Electronics Research, Inc.; November 8, 1950.

(Continued on page 38A)

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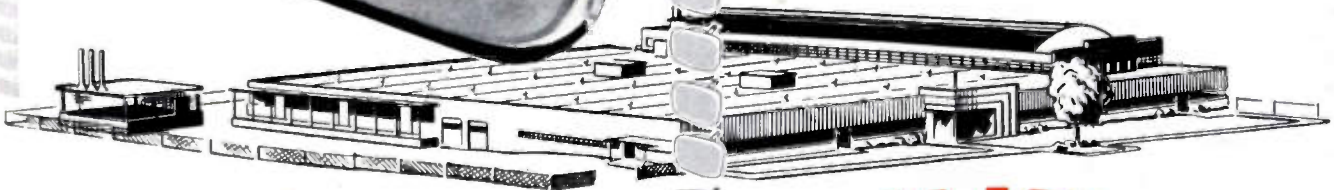

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SECTION MEETINGS

(Continued from page 36A)

Fort Wayne

"Fourth Color in TV," by G. H. Timmings; October 23, 1950.

Kansas City

"Fundamentals of Atomic Theory," by Max Dresden, Faculty, University of Kansas; October 17, 1950.

London

"Process of Conduction and Rectification in Metal Rectifiers," by H. Tull, Faculty, University of Western Ontario; October 2, 1950.

"Electronic Research at the University of Toronto Department of Physics," by J. M. Anderson, Faculty, University of Toronto; October 19, 1950.

"Development of Equipment for Remote Metering of Broadcast Transmitters," by J. M. Toye, Canadian General Electric Company, Ltd.; Inspection of Radio Station CFPL; November 20, 1950.

Los Angeles

"Nonreciprocal Structures for Radar Antennas," by G. Fonda-Bonardi, and "A New Recording Head for Magnetic Drum Memories," by C. B. Forrest; November 7, 1950.

Louisville

"Electronics in Industry," by J. R. Roe, Allis-Chalmers Company; October 27, 1950.

"Tape Recorders," by J. S. Boyer, Magnecord Corporation; "High-Fidelity Amplifiers Using Receiving Tubes with High-Output and Low Distortion," by Frank McIntosh, McIntosh Laboratories; November 10, 1950.

Milwaukee

"Exploration in High-Fidelity Sound," by Marvin Camras, Illinois Institute of Technology; October 24, 1950.

Montreal

"Kinescope Recording," by J. E. Hayes, Canadian Broadcasting Company; November 16, 1950.

New York

"Reliability of Electronic Equipment for Military Service," by F. R. Lack, Western Electric Company; "Reliability in Miniature and Subminiature Tubes," by P. T. Weeks, Raytheon Manufacturing Company; "Tube and Circuit Considerations for Individual Applications," by N. F. Mackenzie, RCA Victor Division; October 4, 1950.

North Carolina-Virginia

"What Makes the IRE Click," by G. W. Bailey, The Institute of Radio Engineers; October 20, 1950.

"Information Systems Research," by Ferdinand Hamburger; November 17, 1950.

Ottawa

"Wave Propagation Beyond Normal Horizons," by J. A. Saxton, N.P.L., England; October 26, 1950.

Philadelphia

"Transient Testing of Loudspeakers," by M. S. Corrington, RCA Victor Division; November 2, 1950.

"Application of Radio Activity to Industry," by George Pieper, Tracerlabs, Inc.; November 8, 1950.

Pittsburgh

"Recent Progress in Television," by D. G. Fink, Electronics; October 16, 1950.

"Magnetic Amplifier Circuits and Applications," by W. J. Dornhoefer, Vickers, Inc.; November 13, 1950.

(Continued on page 39A)

SECTION MEETINGS

(Continued from page 38A)

Princeton

"Control Problems in Nuclear Power Plants," by M. A. Schultz, Westinghouse Electric Corporation; October 12, 1950.

"Television in Europe—Present and Future," by D. G. Fink, Electronics; November 9, 1950.

Rochester

"The Radio Engineer Looks at Industrial Electronics," by E. D. Cook, General Electric Company; October 26, 1950.

San Antonio

"Radio Progress in Europe," by A. W. Straiton, Faculty, University of Texas; October 26, 1950.

"The Vibrotron," by John Ohman, Southwest Research Institute; November 16, 1950.

San Diego

"The Present Status of Color Television," by L. N. Papernow, Television Broadcasting Company; October 3, 1950.

San Francisco

"Printed Circuits and What They Mean to the Parts User," by W. C. Parsons, Centralab Division of Globe Union, Inc.; September 18, 1950.

"Automatic Anti-Roll Ship Stabilization—A Servo Problem," by J. H. Chadwick, Stanford University, and A. J. Morris, Office of Naval Research; October 4, 1950.

"Electronic Equipment Reliability—The Military Problem," by Jerre Noe, Stanford Research Institute; October 25, 1950.

Saint Louis

"Brains for Guided Missiles," by G. J. Fiedler, Sverdrup and Parcel, Inc.; September 20, 1950.

"Calibration of Laboratory-Type Microphones," by R. W. Benson, Central Institute for the Deaf; October 26, 1950.

Schenectady

"The Engineers Approach to the Human Mechanism," by J. R. Ragazzini, Faculty, Columbia University; November 13, 1950.

Seattle

"The Impact of Television on the Radio Serviceman," by Nick Foster, Faculty, Edison Technical School; October 27, 1950.

Syracuse

"Problems in the Television Industry," by A. B. DuMont, Allen B. DuMont Laboratories; October 30, 1950.

Toledo

"The Principles and Application of the Strobo-conn," by S. L. Krauss, C. G. Conn, Ltd.; October 12, 1950.

"The Oscillator System for the Rochester 130 Inch Syncro-Cyclotron," by W. W. Salisbury, Collins Radio Company; November 14, 1950.

Toronto

"New Types of UHF Tubes," by P. A. Redhead, National Research Council; October 23, 1950.

"Kinescope Recording," by J. E. Hayes, Canadian Broadcasting Corporation; November 13, 1950.

Twin Cities

"TV Network Facilities of the Bell System," by M. E. Strieby, American Telephone and Telegraph Company; September 28, 1950.

Washington

"Radio Radiation from the Sun and its Meaning," by J. P. Hagen, Naval Research Laboratory; November 13, 1950.

(Continued on page 40A)

SECTION MEETINGS

(*Continued from page 39A*)

SUBSECTIONS

Binghamton

"High-Frequency Propagation in the Lower Atmosphere," by H. G. Booker, Faculty, Cornell University; October 26, 1950

Hamilton

"The Hammond Organ," by W. Benger, Northern Electric Company; November 6, 1950.

Long Island

"Color Television System for Industry," by R. F. Cotellessa, A. B. DuMont Laboratories, discussed by A. V. Loughren, Hazeltine; October 11, 1950.

Mid-Hudson

"Traveling-Wave Tubes," by L. Brillouin, International Business Machine Corporation; October 24, 1950.

Monmouth

"Application of Printed Circuits to Subminiature Design," by M. U. Cohen, Balco Research Laboratories; April 19, 1950.

"A Survey of Latest Color Television Developments," by A. G. Jensen, Bell Telephone Laboratories, Inc.; October 18, 1950.

STUDENT BRANCH MEETINGS

University of Alberta, IRE Branch

"Master Control Room Facilities for Broadcast Studios," by Mr. Tanner, Northern Electric Company; October 19, 1950.

University of Arkansas, IRE Branch

"Troubleshooting Television Receivers," by Mr. Smith, Smith Radio Shop; October 25, 1950.

Carnegie Institute of Technology, IRE-AIEE Branch

"Electronic Lighting System Recently Installed for the Control of Stage Lighting," by G. Royer, Faculty Carnegie Institute of Technology, and Professor Nelson, Faculty, Carnegie Institute of Technology; November 21, 1950.

Clarkson College of Technology, IRE Branch

"Voice Highways on the Air," by M. S. Paige, Bell Telephone Company; November 2, 1950.

Business Meeting; Films: "DC Commutation," and "AB Circuit Breakers"; November 16, 1950.

University of Colorado, IRE-AIEE Branch

Business Meeting; Film: "Hoover Dam"; October 18, 1950.

"Magnesium Manufacture," by John Thomas, Student, University of Colorado; Films: "Battle for Oil," and "Our State University"; October 30, 1950.

"Nuclear Resonance by Super-Regenerative Technique," by J. R. Zimmerman; "Methods of Investigating Beta Ray Spectra with Magnetic Spectrometers," by A. A. Bartlett, Faculty, University of Colorado; November 10, 1950.

"Power Transformers," by W. S. Pullen, Jr., R. E. Ingmire, and Mr. Fabion, General Electric Company; November 15, 1950.

(*Continued on page 42A*)

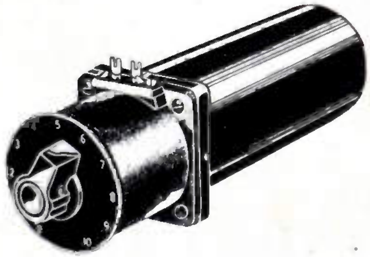
FREED

"PRODUCTS OF EXTENSIVE RESEARCH"

TRANSFORMERS & INSTRUMENTS

DECADE INDUCTORS BY FREED

A complete line of precision high stability decade inductors covers the range from one tenth of millihenry to hundred henries and frequencies from 30 cycles to 300,000 cycles.



The inductors can be used either as secondary laboratory standards or as high Q components in wave filters, equalizers and tuned circuits for audio and radio frequencies. Individual inductors are wound on temperature stabilized molybdenum permalloy toroidal core. Four coils (nominal values 1, 2, 3, 4) are combined in an individual decade. A special low loss ceramic switch combines the coils in such a way as to give the eleven successive values from 0 to 10.

Special silver alloy contacts insures very low contact resistance.

OUTSTANDING FEATURES

Very high Q at frequencies up to 300,000 cycles.
High natural frequency.
Astatic to external magnetic fields.
Very low temperature coefficient.
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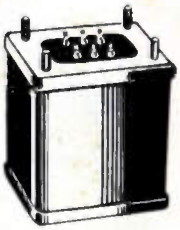
EQUENCY "Q" INDICATOR

NO. 1210 NULL
ETECTOR & VACUUM
TUBE VOLTMETER

FREQUENCY RANGE 30 TO 2000 CYCLES			
Q = 50 @ 200 CYCLES			
*#1164	DECADE INDUCTOR	111	Henry total in steps of .1 Henry
FREQUENCY RANGE 100-2000 CYCLES			
Q = 80 @ 500 CYCLES			
#1341	DECADE INDUCTOR	100	Henry total in steps of 10 Henry
FREQUENCY RANGE 500-20,000 CYCLES			
Q = 60 @ 1000 CYCLES			
#1160	DECADE INDUCTOR	11.1	Henry total in steps of .01 Henry
#1163	DECADE INDUCTOR	1.11	Henry total in steps of .001 Henry
#1260	DECADE INDUCTOR	11.11	Henry total in steps of .001 Henry
FREQUENCY RANGE 500-20,000 CYCLES			
Q = 160 @ 1000 CYCLES			
#1220	DECADE INDUCTOR	.01	Henry total in steps of .001 Henry
#1230	DECADE INDUCTOR	.1	Henry total in steps of .01 Henry
#1240	DECADE INDUCTOR	1	Henry total in steps of .1 Henry
#1260	DECADE INDUCTOR	11.11	Henry total in steps of .001 Henry
#1270	DECADE INDUCTOR	10	Henry total in steps of 1 Henry
#1280	DECADE INDUCTOR	1.11	Henry total in steps of .001 Henry
#1290	DECADE INDUCTOR	11.11	Henry total in steps of .001 Henry
#1310	DECADE INDUCTOR	11.1	Henry total in steps of .01 Henry
FREQUENCY RANGE 2000-50,000 CYCLES			
Q = 200 @ 10,000 CYCLES			
#1161	DECADE INDUCTOR	1.11	Henry total in steps of .001 Henry
FREQUENCY RANGE 10,000-300,000 CYCLES			
Q = 200 @ 100,000 CYCLES			
#1162	DECADE INDUCTOR	.111	Henry total in steps of .1 millihenry

*#1164 DECADE INDUCTOR is wound on a special nickel alloy core.



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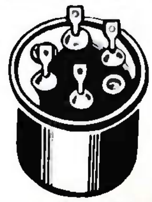

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STUDENT BRANCH MEETINGS

(Continued from page 40A)

Cooper Union, IRE Branch

"Particle Accelerators," by H. C. Wolfe, Faculty, Cooper Union; October 30, 1950.

Tour: Columbia University Cyclotron; November 4, 1950.

Business Meeting, November 14, 1950.

Cornell University, IRE-AIEE Branch

"Galloping Bridges," by D. B. Steinman, October 25, 1950.

University of Denver, IRE-AIEE Branch

Business Meeting; November 1, 1950.

University of Detroit, IRE Branch

"Psychology in Engineering," by A. Schneiders, Faculty, University of Detroit; October 10, 1950.

University of Florida, IRE-AIEE Branch

"The Application Engineer in Industry," by C. W. Drake, Faculty, University of Florida; Film "Television"; October 17, 1950.

Illinois Institute of Technology, IRE Branch

Film: "Sightseeing in the Home"; Election of Officers; October 16, 1950.

State University of Iowa, IRE Branch

Organization of IRE and AIEE Student Groups; September 27, 1950.

Organization of Student Branch and Program Planning; October 4, 1950.

Student Speeches; October 11, 1950.

Construction of Homecoming Monument; October 18, 1950.

Student Speeches; October 25, 1950.

Film: "Radio-Frequency Heating and its Application"; November 1, 1950.

"Cyclotron," by J. J. Livinggood, Collins Radio Company; November 8, 1950.

University of Kentucky, IRE Branch

Film: "Electrified Farming," by General Electric Company; October 12, 1950.

"Human Relations," by Gifford Blyton, Faculty, University of Kentucky; October 26, 1950.

"History of the Piano," by Nathaniel Patch, Faculty, University of Kentucky; November 9, 1950.

Lafayette College, IRE-AIEE Branch

"Quality Sound Systems," by H. H. Elliott, Student, Lafayette College; November 14, 1950.

Lehigh University, IRE Branch

Films: "Radar, Finding the Enemy," and "Radar, Weapon of Attack"; October 26, 1950.

Marquette University, IRE-AIEE Branch

"Color Television," by A. F. Petrie, Student, Marquette University; October 19, 1950.

Smoker, October 25, 1950.

Business Meeting, November 2, 1950.

Business Meeting, November 16, 1950.

University of Maryland, IRE-AIEE Branch

"Radio Telemetering," by R. S. Butts, Melpar Corporation; November 15, 1950.

Massachusetts Institute of Technology, IRE-AIEE Branch

Tour of Television Station WBZ-TV; October 10 and 11, 1950.

(Continued on page 43A)

STUDENT BRANCH MEETINGS

(Continued from page 42A)

"The Application and Manufacture of Circuit Breakers," by E. A. Kilgour, Allis-Chalmers Company; October 23, 1950.

Tour of Allis-Chalmers Boston Works; October 24–26, 1950.

"Feedback Control Mechanisms," by G. S. Brown, Massachusetts Institute of Technology; October 31, 1950.

Michigan College of Mining and Technology, IRE-AIEE Branch

"Summer Employment in Sweden," by Lynn Zellmer, Student, Michigan College of Mining and Technology; October 17, 1950.

Business Meeting, October 31, 1950.

University of Minnesota, IRE-AIEE Branch

Film: "This is Resistance Welding"; October 4, 1950.

"Honeywell Research Projects," by F. J. Larson, Minneapolis-Honeywell Company; October 11, 1950.

Electronics Demonstration by Aifred Crossely and Associates, Fred Nearing, Speaker; November 1, 1950.

"The Future of Electrical Engineering," by H. Hartig, Faculty, University of Minnesota; November 8, 1950.

New Mexico College of Agriculture & Mechanic Arts, IRE-AIEE Branch

"Electron Tubes," by H. H. Brown, Faculty, New Mexico College of Agriculture and Mechanic Arts; October 12, 1950.

"Plant Operations of the Nichols Electrolytic Copper Refinery," by D. Kunklen, Nichols (Phelp Dodge Copper Company); October 26, 1950.

College of the City of New York, IRE Branch

Business Meeting; Films: "Squirrel Cage Rotor Principles," "Repulsion Motor Principles," "Split Phase Motor Principles," and "Naturally, It's FM"; October 26, 1950.

"Test Equipment in the Production and Servicing of TV," by Mr. Petrasek, Radio Corporation of America; November 9, 1950.

"UHF Impedance Measurements," by Mr. Thurston, General Radio Corporation; November 16, 1950.

New York University, IRE Branch (Day Division)

Film: "Just Imagine," by Bell Telephone Company; October 26, 1950.

"Opportunities for Students in the Graduate School of Engineering," by S. G. Lutz, Faculty, New York University; November 2, 1950.

"Errors in Telemetering," by Leon Hillman, Faculty, New York University; Smoker; November 8, 1950.

Business Meeting, November 9, 1950.

New York University, IRE-AIEE Branch (Evening Division)

"Feedback Amplifiers," by Dr. Mulligan, Faculty, New York University; Election of Officers; October 27, 1950.

Northeastern University, IRE-AIEE Branch

"Current Practices in Television," by Hollis Baird, Faculty, Northeastern University; October 3, 1950.

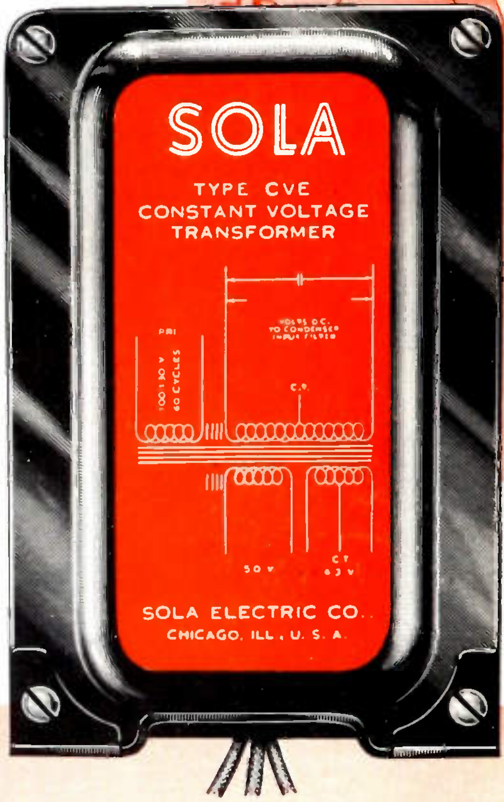
Tour of High Voltage Engineering Corporation; October 19, 1950.

(Continued on page 46A)

Built-in VOLTAGE REGULATION *at Moderate Cost*

Insure constant plate and filament voltage for your electronic products with standard SOLA "CVE" POWER TRANSFORMERS.

Specify the new SOLA "CVE" Constant Voltage Power Transformers in your circuit design to eliminate the variable of fluctuating line voltage at unusually low cost. Regulation of filament and plate supply is ±3% at line voltage variations from 100 volts to 130 volts.



The SOLA "CVE" standard Power Transformers are completely automatic and continuous in regulation . . . have no moving parts or tubes . . . and are self-protecting against short circuit. They are stocked in 42 V.A., 75 V.A. and 210 V.A. capacities to cover most electronic power supply requirements. We invite your inquiries on the application and benefits of the moderately priced "CVE" Constant Voltage Power Transformers to your product. For full electrical and mechanical specifications write for Bulletin K-CVE-138.

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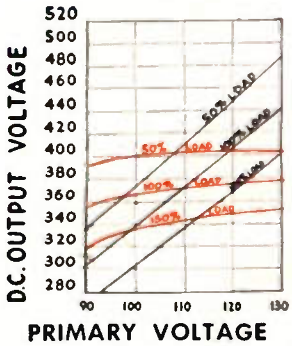

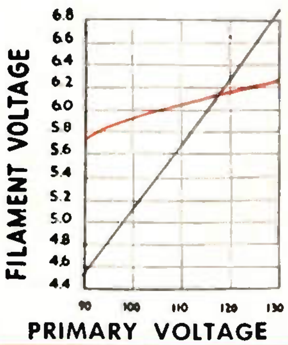

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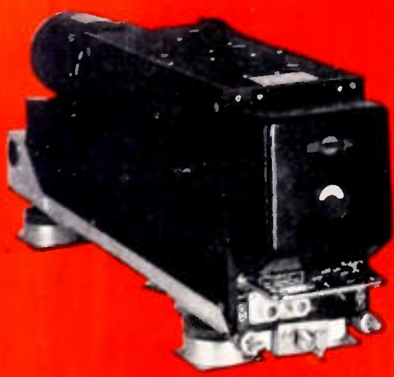
"CV" for high precision voltage regulation. "CVE" for regulated electronic power supplies. "CVH" for constant voltage with less than 3% harmonic distortion. "CVA" for constant voltage on television receivers.

Transformers for: Constant Voltage • Cold Cathode Lighting • Airport Lighting • Series Lighting • Fluorescent Lighting • Luminous Tube Signs
Oil Burner Ignition • X-Ray • Power • Controls • Signal Systems • etc. • **SOLA ELECTRIC COMPANY, 4633 W. 16th Street, Chicago 50, Illinois**

Manufactured under license by: ENDURANCE ELECTRIC CO., Concord West, N. S. W., Australia • ADVANCE COMPONENTS LTD., Walthamstow, E., England
UCOA RADIO S.A., Buenos Aires, Argentina • M. C. B. & VERITABLE ALTER, Courbevoie (Seine), France

TWA buys Collins vhf transmitters

for its entire Martin 4-0-4 fleet



Collins shockmounted 17L-2 transmitter

Collins 314Y-1 remote control unit

By its purchase of 40 Martin 4-0-4's, which will begin to go into domestic service next spring, Trans World Airlines follows its traditionally vigorous course of progress.

Also reflecting TWA's policy of providing the most efficient air transportation that modern facilities permit, is the fact that the radio complement of every one of the new 4-0-4's will include a Collins 17L-2 vhf transmitter.

TWA engineers made a careful study of available vhf transmitting equipment. They knew that vhf communication is line-of-sight communication, noise-free within its range, but not applicable over great distances. They chose the 17L-2 largely because it had the best size-weight to power ratio.

The 17L-2, small enough to be housed in a standard ½ATR size case, weighs only 19 pounds. Yet it has a power output of eight watts or better into a 52-ohm load, thus assuring that transmissions will be received and acknowledged at the busiest air terminals.

In addition, it provides fingertip remote frequency control of all 180 channels reserved for aircraft communication in the vhf band. The 314Y-1 remote control unit can be located wherever it is most convenient.

All airline operations and communications people should be fully informed about this highly developed transmitter. A descriptive bulletin will be sent you on request.

IN RADIO COMMUNICATIONS, IT'S . . . COLLINS

COLLINS RADIO COMPANY, Cedar Rapids, Iowa

11 West 42nd Street, NEW YORK 18 — 27 West Olive Avenue, BURBANK

Student Branch Meetings

(Continued from page 43A)

Tour of Mystic Station of Boston Edison; October 26, 1950.

"Nature, Art and Engineering," by E. W. Boehne, Faculty, Massachusetts Institute of Technology; "Electronics in Overalls," by F. Hansen, General Electric Company; November 2, 1950.

Business Meeting; Films: "Adventures in Research," and "Echoes in War and Peace"; November 7, 1950.

UNIVERSITY OF NOTRE DAME, IRE-AIEE BRANCH

Business Meeting, September 20, 1950.

OHIO STATE UNIVERSITY, IRE-AIEE BRANCH

"Commercial Distribution of Power," by E. E. Kimberly, Faculty, Ohio State University; November 9, 1950.

PENNSYLVANIA STATE COLLEGE, IRE-AIEE BRANCH

Films: "Railroad Signals," and "X-Ray Inspection"; November 2, 1950.

PRATT INSTITUTE, IRE BRANCH

"Electronic Computers," by Simon Gluck, Faculty, University of Pennsylvania; October 26, 1950.

PRINCETON UNIVERSITY, IRE-AIEE BRANCH

Film: "Daredevils of the Alps"; October 12, 1950.

"Is Engineering a Profession?" by J. F. Fairman, Consolidated Edison Company; November 2, 1950.

RENSSELAER POLYTECHNIC INSTITUTE, IRE-AIEE BRANCH

"Operator Toll Dialing," by J. C. Macrow and G. W. Schaible, New York Telephone Company; November 15, 1950.

RUTGERS UNIVERSITY, IRE-AIEE BRANCH

"Trends in Electrical Measuring Instruments," by Mr. Estopey, Weston Instruments Corporation; October 24, 1950.

"Theory and Application of Transistors," by S. M. Christian, RCA Laboratories; Business Meeting, November 14, 1950.

SYRACUSE UNIVERSITY, IRE-AIEE BRANCH

"Engineers in Industry," by C. H. Heiden, General Electric Company; October 25, 1950.

"Korea," by Albert Lee, Student, Syracuse University; November 16, 1950.

UNIVERSITY OF TEXAS, IRE-AIEE BRANCH

"Austin Power System," by W. E. Seaholm; October 2, 1950.

Demonstration of Electrical Engineering Laboratories, October 30, 1950.

Film and Award Presentation, November 6, 1950.

UNIVERSITY OF TOLEDO, IRE-AIEE BRANCH

Business Meeting; Films by Bell System; October 23, 1950.

"Squirrel-Cage Motors," by Lester Fox, Electric Auto-Lite Company; November 14, 1950.

TUFTS COLLEGE, IRE-AIEE BRANCH

"Concepts of Microwave Techniques," by Raymond Crisp, General Communications Company; October 25, 1950.

TULANE UNIVERSITY, IRE-AIEE BRANCH

Business Meeting, September 30, 1950.

"Relationship of City Sections and Student Branches," by J. A. Cronvich, Faculty, Tulane University; "Role of the Student Branch in the College Scheme," by L. H. Johnson, Faculty, Tulane University; October 2, 1950.

VIRGINIA POLYTECHNIC INSTITUTE, IRE BRANCH

"The Application of our Field in the Field of Applied Mechanics," by E. J. Maher, Faculty, Virginia Polytechnic Institute; October 10, 1950.

(Continued on page 47A)

Student Branch Meetings

(Continued from page 46A)

Film: "RF Heating," by Westinghouse Flm Library; October 24, 1950.

UNIVERSITY OF VIRGINIA, IRE BRANCH

Film: "Summer Storms"; October 10, 1950.

"Computers," by Hill Montague, Student, University of Virginia; October 24, 1950.

UNIVERSITY OF WYOMING, IRE-AIEE BRANCH

Business Meeting; Films: "Sound Recording," and "Radio Antennas"; October 18, 1950.

Business Meeting; Films: "National Air Races," and "Principles of Radio Receivers"; November 8, 1950.

MEMBERSHIP

The following transfers and admissions were approved and will be effective as of January 1, 1951:

Transfer to Senior Member

Adler, H. J., Hallicrafters Company, 4401 Fifth Ave., Chicago, Ill.
Antman, M. A., 1244 Kumler Ave., Dayton 7, Ohio
Bernstein, J., 1363 Findlay Ave., New York 56, N. Y.
Chandler, G. C., 846 Howe St., Vancouver, B. C., Canada
Cherry, S. J., 410 Prospect St., East Orange, N. J.
Felton, W. W., 10 Hemlock Rd., Lansdowne, Pa.
Hansen, J. A., 23 Bee St., Valley Stream, N. Y.
Jarmie, T. W., 759 Douglas Ave., Oxnard, Calif.
Keller, A. C., 125 White Plains Rd., Bronxville 8, N. Y.
Ketchum, P. M., 127 Nichols Ave., Syracuse 6, N. Y.
Krevsky, S., 76 Throckmorton Ave., Red Bank, N. J.
Robinson, S. A., 111 Ivy Rock Lane, Westgate Hills, Havertown, Pa.
Saxon, M., Box 575, Lufkin, Tex.
Smith, O. J. M., 228 Amherst Ave., Berkeley 8, Calif.
Thurston, J. N., Electrical Engineering Department, University of Florida, Gainesville, Fla.
VanScoyoc, J. N., 946 S. Maple, Oak Park, Ill.
Wilhelm, J. F., Leaman Acres, R. D. 7, Box 223, Lancaster, Pa.
Williams, J. B., Philco Corporation, C & Tioga Sts., Philadelphia, Pa.
Wright, R. R., Box 568, Blacksburg, Va.
Zink, A. J., Jr., 64 Whittier St., Andover, Mass.

Admission to Senior Member

Butler, E. W., 100 Kingsland Rd., Clifton, N. J.
McDonough, J. A., 12 Andover Pl., Glen Cove, L. I., N. Y.
Williams, W. B., 13103 Corbett Ave., Detroit 13, Mich.

Transfer to Member

Abjanich, G. S., Sylvania Electric Company, 1280 Main St., Buffalo 9, N. Y.
Allen, J. E., Eighth & Oak Park Rd., Lansdale, Pa.
Bennett, W. P., RCA Victor Division, New Holland Pike, Lancaster, Pa.
Butler, H. J., 126 Seminary Ave., Lutherville, Md.
Cole, H. H., 1232 N. Stanford, Albuquerque, N. Mex.
Grimm, E. M., 1 East St., Boonton, N. J.
Johnston, S. L., 2795 Peachtree Rd., Atlanta, Ga.
Judge, H., Pingree, Idaho
Manley, T. M., Box 3381, Wright-Patterson AF Base, Dayton, Ohio
Miller, H. G., Bell Telephone Company of Pennsylvania, 121 N. Broad St., Philadelphia 7, Pa.

(Continued on page 48A)

MEMBERSHIP

(Continued from page 47A)

Newbold, R. F., Instrumentation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.
Opie, A., 15 Tulip Lane, Short Hills, N. J.
Pavela, H. W., 4541 S. Greenwood Ave., Chicago 15, Ill.
Peterson, T. A., Jr., 615 Baker Pl., Elizabeth 3, N. J.
Rogers, A. J., 324 Roxbury Rd., Dayton 7, Ohio
Slusser, E. A., R. D. 4, Beech St., Babylon, N. Y.
Walker, R. C., Box 414, Bucknell University, Lewisburg, Pa.

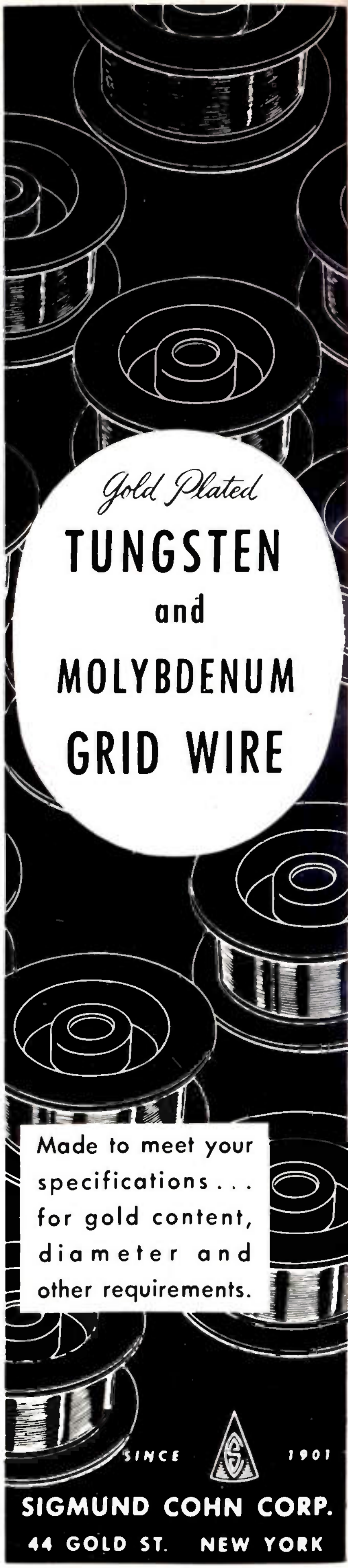
Admission to Member

Beaver, J. A., Jr., 937 Easton Rd., Roslyn, Pa.
Creighton, A. M., Jr., 2221 E. Osborn Rd., Phoenix, Ariz.
Gresko, J. J., 9 Doyer Ave., White Plains, N. Y.
Holley, M. V., 214 Bedford Ave., Buffalo, N. Y.
Makalanda, A., Hydraulic Research & Soil Mechanics Laboratories, Jawatta Rd., Colombo, Ceylon
McLaughlin, C., 567 Superior Ave., Dayton, Ohio
Pederson, C. N., Armour Research Foundation, 35 W. 33 St., Chicago 16, Ill.
Petersen, C. C., 3640 Fullerton Ave., Chicago 47, Ill.
Rubinfien, D., 117 S. Central Ave., Chicago 44, Ill.
Stuarts, A., Jr., 341 Taylor Ct., Troy, N. Y.

The following elections to Associate grade were approved and will be effective as of December 1, 1950:

Ashus, A., 17 Sycamore St., Massapequa, N. Y.
Bahl, D. R., Technical & Development Ctr., Govind Bhawan, Jabalpur, India

(Continued on page 59A)



NEW INDICATOR ION TRAP

a Rauland "Exclusive"

Helps you Cut Production Costs

Rauland's new Indicator Ion Trap can help you in your battle to cut pennies off production costs and thereby to price receivers competitively.

First of all, the Indicator Ion Trap completely eliminates the need for any equipment and any trained judgment in the adjustment of ion trap magnets. Adjustment can be made faster than equipment could be attached. The ion trap magnet is simply moved until the green glow signal is reduced to minimum. It can be done in seconds with absolute accuracy—without even seeing the front of the picture tube.

Second, the Rauland Tilted Offset Gun which incorporates this Indicator Ion Trap requires only one Ion Trap Magnet instead of two, nibbling a little more off production costs. Yet it gives better results—the electron beam is bent only once and is focused to maximum sharpness.

Specify Rauland tubes with these exclusive advantages, and get the benefits that only Rauland offers.

For further information, write to . . .

RAULAND

The first to introduce commercially these popular features:

Tilted Offset Gun

Indicator Ion Trap

Luxide (Black) Screen

Reflection-Proof Screen

Aluminized Tube

THE RAULAND CORPORATION

Perfection Through Research

4245 N. KNOX AVENUE · CHICAGO 41, ILLINOIS

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . .
The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.
I East 79th St., New York 21, N.Y.

ELECTRONIC ENGINEER

Senior graduate engineer with at least 10 years experience in receiver design. Minimum of 2 years experience UHF desirable. Must me capable of assuming project responsibility. Location Connecticut. Excellent opportunity. Salary high. Submit resume. Box 631.

ENGINEER

Small California transmitting tube company requires engineer for medium frequency tube work, also for development and manufacture of klystrons and pulse tubes. Give full details giving age, stating education, experience, salary expected. Box 632.

ENGINEERS AND PHYSICISTS

Project and senior engineers desired for work on several theoretical and experimental programs of diversified nature involving military applications of electronics. Applicants should have 3 or more years of experience in research and development in some branch of electronics and preferably advanced graduate training. Command of physical fundamentals and analytical ability important. Small, expanding company located in college town. Opportunities of graduate study. Reply Personnel Manager, Haller, Raymond and Brown, Inc. State College, Pa., stating education, experience, salary expected.

ENGINEERS

National Broadcasting Company needs experienced engineers with commercial television operating experience or standard broadcasting control room experience. Apply Room 505, 30 Rockefeller Plaza, New York, N.Y.

ELECTRON TUBE ENGINEERS

For development and production. Experience miniature or sub-miniature tubes desirable. Apply by letter only to Personnel Dept., Sonotone Corp., Elmsford, New York.

RADIO AND TELEVISION ENGINEERS

Experienced in design of high frequency circuits such as FM tuners, TV boosters and TV antennae. Salary commensurate with ability. Write giving full details—Mr. Stone, Talk-A-Phone Company, 1512 South Pulaski Rd., Chicago 23, Ill. Lawndale 1-8414.

ENGINEERS

Physicists, chemists or EE's with Ph.D. or equivalent and experience in the field of solid state physics for research work on semi-conductor devices employing germanium and silicon. An excellent opportunity in a research laboratory of a leading manufacturer with laboratories

(Continued on page 52A)

POSITIONS OPEN

(Continued from page 50A)

in New York State. Send complete resume. Our employees have been notified. Box 633.

SENIOR ELECTRONICS ENGINEER

For design and development of circuitry for ultrasonic equipment timing circuits, audio oscillators amplifiers, and audio measuring equipment to meet Navy specifications. Experience required: 3-5 years in development of audio or supersonic equipment for government or industrial usage. Must have B.S. in physics or electrical engineering. Write: Personnel Director, Box 30, State College, Pa.

DEVELOPMENT TECHNICIANS

At least 3 years experience in layout. (Including rough drafting) of electronic chassis. Should also be experienced in electronics testing and trouble shooting. Write: Personnel Director, Box 30, State College, Pa.

ELECTRICAL ENGINEER

Graduate electrical engineer with a minimum of 2 years' experience. For design and development of audio transformers and filters. Permanent position with progressive firm located in Chicago. Give details stating age, education, experience, references, availability for work and salary expected. Box 634.

ELECTRONIC SCIENTIST

For research in upper atmosphere rocket program. Must have an appropriate degree, and at least 3 years experience, with emphasis on electronics as applied to upper atmosphere research or an allied field. Please address replies, containing a brief resume of experience to Employment Officer, Naval Research Laboratory, Washington 25, D.C.

RADIO & RADAR ENGINEERS

Radio and radar engineers for aircraft installation and application design work. Should have 5 or more years' experience with aircraft radio or radar systems and preferably have aircraft installation or antenna design, selection and application experience. Experienced aircraft electrical engineers are also needed. Contact Engineering Personnel Section, Chance Vought Aircraft, P. O. Box 5907, Dallas, Texas.

ENGINEERS

Florida East Coast opportunity for permanent positions in Electronic Instrumentation Research and Development. Telemetering, Pulse Circuits and Computer experience desirable. Junior, Senior and Project Engineers may send resume of education, experience, references, availability and salary expected to Radiation, Inc. Melbourne, Florida.

ELECTRONICS ENGINEER-WRITER

A long-established concern of high repute in the military instruction book field requires a technical writer of project director calibre for permanent employment. A solid electronics background is essential, as is direct experience in the planning and writing of instruction books to Navy, Air Force, and Signal Corps

(Continued on page 54A)

UTC for Military Components

UTC was the largest supplier of transformer components in World War II. Present UTC production is on a similar basis. Illustrated below are a few of the thousand military types in UTC 1950 production.

United Transformer Co.

150 VARICK STREET • NEW YORK 13, N. Y.

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

Positions Open

(Continued from page 52A)

specifications. Security clearance will be required after employment. Concern is located in downtown New York City. Send detailed resume to Box No. 638.

ELECTRONIC ENGINEERS

Senior engineers or Physicists, degree and experience in Radar, Pulse Circuits, Digital or Analogue Computers, or Servomechanisms. Electronic Engineering Company, 180 S. Alvarado St., Los Angeles 4, California.

CIRCUITS AND MICROWAVE ENGINEERS

Rapidly expanding company in the instrumentation field has openings for Senior engineers with several years of experience. Men with Masters or Ph.D. Degrees preferred. Excellent opportunities in field of VHF, UHF and microwave test equipment. Location Brooklyn. Send complete resume to Box 639 I.R.E.

ELECTRONIC ENGINEERS

Naval Ordnance Laboratory, outside Washington, D.C., offers unusual opportunities to men with college degrees, in the fields of communications and instrumentation. Starting salaries may run as high as $6400. Men with specialized experience in the video or the instruments fields are urgently required. Attractive openings exist for men with inclinations in other directions. Address: Personnel Dept., Atten: UT, Naval Ordnance Laboratory, White Oak, Silver Spring 19, Md.

ELECTRONIC ENGINEERS

At least three (3) years post-college experience in development, DC amplifier, digital computers, pulse and servo design. Established Company, New York City, Box 640.

ENGINEERS

Expanding Southwest organization has need for electronic engineers and technicians, mechanical engineers, machinists, draftsmen, mechanical and electronic; to work with cinetheodolites, cameras, radars, computers and telemeters, as used for guided missile test instrumentation. Work consists of operation, maintenance, development and construction. Write for application form. Salaries based on application and personal interview are commensurate with education, training, experience, job responsibility and other factors. Particularly needed: engineers and technicians experienced in small quantity production to JAN specifications. Address reply to: C. E. Riggs, Land-Air, Inc. Box 76 Holloman Air Force Base, New Mexico.

ELECTRONIC ENGINEERS

Sales, design and application engineering. Positions open in Sales and Engineering Depts. of manufacturer of communications equipment. Salary commensurate with experience. Location Chicago. Box 642.

TELEVISION ENGINEER

For design and development of Television and Radio Receivers by one of Chicago's oldest Television and Radio manufacturers. Applicant required to have college degree or equivalent. Experience necessary in design and measuring technique. Give age, experience, reference, etc. Salary open. Apply box 643.

ELECTRONICS ENGINEERS

A large Cleveland Manufacturer of Automotive and Aircraft Engine Parts is entering the Electronics Field. De-

(Continued on page 55A)

Positions Open

(Continued from page 54A)

velopment Engineers and Laboratory Technicians are needed to conduct development, testing of qualification test samples, design and build test setups, writing reports, etc. Work is original in the field of high frequency signal transmissions (frequencies up to 11,000 megacycles per second). Reply should give training, experience in detail, past employers, earnings and other pertinent facts. Thompson Products, Inc., 2196 Clarkwood Road, Cleveland 3, Ohio.

ENGINEERS

Four senior electronics research and development engineers with a minimum of five years' experience in pulse and general radar design. Assignments will include microwave test equipment projects and guided missile development. Degree essential. Excellent opportunity for advancement in rapidly expanding company. Modern plant of 250 workers located in Los Angeles area. Three experienced junior engineers also required. Send resume of education, experience, salary desired, to Box 644.

ELECTRICAL ENGINEER

B.S. or M.S. to work with group on circuits and miniaturization problems relating to electronic equipment and technique. Must be capable of job planning and project control. Minimum 5 years experience in both audio and TV circuits and tube component design or equipment production. Prefer engineering physicist or electrical engineer with some mechanical design experience and major interest in circuits. Position is with physics Lab-

(Continued on page 56A)

Positions Open

(Continued from page 55A)

oratories. Sylvania Electric Products Inc., Bayside, Long Island. Please address replies to Personnel Manager 40-22 Lawrence Street, Flushing.

Positions Wanted By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants, and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The Institute publishes free of charge notices of positions wanted by I.R.E. members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The Institute necessarily reserves the right to decline any announcement without assignment of reason.

TELEVISION ELECTRONICS TECHNICIAN

2½ years B.E.E. New York University. 2 years Naval Electronics Radar Technician. 2 years T.V. experience in production test and field engineering for New York manufacturer. Desires position New York—Long Island area. Resume on request. Box 489 W.

ENGINEER

B.E.E. electronics option, University of Minnesota, June 1949, high scholastic standing. 3 years Army Signal Corps radio repairing and operating. Desires electronic design or development position. Location immaterial. Box 490 W.

ENGINEER

Age 43, married. Extensive practical radio, F.M., T.V. and communications experience. Desires position as Service Manager with a reputable distributor. New York, New Jersey or Conn. area. Resume on request. Box 491 W.

JUNIOR ENGINEER

B.E.E. June 1950, New York University. Desires position in electronics or communications field. Experience remote control repairman on AA guns. 1½ years television and radio repairman. Age 27, single. Willing to relocate anywhere in U.S. Box 493 W.

SALES ENGINEER

B.E.E. June 1947, Polytechnic Institute of Brooklyn, Eta Kappa Nu. 2 years varied radar, radio, electronic technical experience. 1 year own business. Single, age 25. Box 494 W.

ELECTRICAL ENGINEER

Six years experience in electromechanical and industrial electronics and measurements. Professional engineer. New York area preferred. $7500 per year. Box 500 W.

SALES ENGINEER

Graduate "Salesmanship for Engineers" course at C.C.N.Y., B.E.E. communica-

(Continued on page 58A)

General Electric's GL-5855 is designed for control work that requires high current capacity . . . toughness and stamina . . . utmost dependability.

HERE'S a new motor-control tube that handles big currents, and can take the punishment of full-time, heavy-duty operation . . . on through a long life span which squeezes value out of every dollar.

Study the clean, sturdy design. Under the heavy glass envelope is a structure so simple, so inherently braced, that vibration and shocks have little effect. Bolts fasten the tube to the panel, assuring firm support and providing tight, solid electrical contacts.

Cleanness and simplicity go further . . . right through to application needs. No snubber circuit is required. The high commutation factor (200) means that the anode gas absorption from inductive loads is negligible. This makes for long tube life with a straightforward, economical circuit.

To a husky 12.5-amp rating are added high anode voltage, high peak-to-average current ratio, stable control characteristics, and short heating time. This pattern has made the earlier, smaller GL-5544 and GL-5545 popular; assures enthusiastic acceptance for the new thyratron. Also, the GL-5855 has a wide temperature range, making the tube virtually climate-proof.

Wire or write for complete ratings and performance data. Or better, ask for an across-your-desk talk with an experienced G-E tube engineer! *Electronics Department, General Electric Company, Schenectady 5, N. Y.*

GL-5855

12.5-amp control thyratron

Filament voltage	2.5 v
Filament current	34 amp
Peak anode voltage, forward and inverse	1,500 v
Peak cathode current	150 amp
Avg cathode current	12.5 amp
Current averaging time	15 sec
Ambient temp range	−55 to +70 c
Commutation factor	200

* * *

NOTE—Commutation factor is the product of the rate of current decay in amperes-per-microsecond just prior to commutation and the rate of inverse voltage rise in volts-per-microsecond just after commutation.

GENERAL ELECTRIC

180-K1

Positions Wanted

(Continued from page 56A)

tions—N.Y.U. 1950. Single, age 24. Looking for job as junior or trainee that needs a man that has technical ability, knows how to sell, and has sincere desire to sell. Box 501 W.

ENGINEER

M.S. in E.E., University of Wisconsin February 1951. UHF and microwave major; 26, married, 2 children; Desires design and test or field position preferably in west. Resume on request. Box 502 W.

ELECTRONIC ENGINEER

Ph.D. in communications expected June 1951. 3 years full time University teaching. 2 years Navy electronics. Age 25. Married. Desires academic or industrial position in eastern U.S. Available July 1, 1951. Box 503 W.

TELEVISION ENGINEER

B.S.T.E. American Institute of Technology, Sept. 1949. 5 years experience in radio and TV servicing. 1st class radio telephone license. Class B amateur license. Age 23. Will relocate. Résumé upon request. Box 504W.

ELECTRONIC ENGINEER

B.E.E. June 1949. Communications option. Rensselaer Polytechnic Institute. 3 years electronic technician 1/cl. U. S. Navy. 1½ years experience test and development fire and control equipment; 1 year M.B.A. at N.Y.U. major management. Position desired: test and development or project engineering with communication or electronic instruments. New York—Long Island area. Resume on request. Box 487 W.

MEMBERSHIP

(Continued from page 48A)

Baker, G., Jr., 609 Jackson St., Amarillo, Tex.
Ball, R. M., 9407 S. Winchester Ave., Chicago, Ill.
Bell, C. W., 36 Lascelles Blvd., Toronto 12, Ont., Canada
Bicking, H. P., Gough Ave., Ivyland, Pa.
Boorsma, C. J., 224 Eighth St., Ann Arbor, Mich.
Boroff, D., 6051 N. Talman, Chicago 45, Ill.
Brooks, J. H., 17551 Santa Rose Dr., Detroit 21, Mich.
Bruno, G. A., 358 Union Ave., Wood Ridge, N. J.
Chartoff, M., 5 Scheneck Ave., Matawan, N. J.
Cooley, C. C., Jr., Philco Corporation, 6513 N. Broad St., Philadelphia 26, Pa.
Crane, R. L., 2345 Ocean Ave., Brooklyn 29, N. Y.
Dannemiller, J. R., 4334 Groveland Ave., Cleveland, Ohio
Davis, R. G., 1358 N. Springfield, Chicago 51, Ill.
Decho, K. R., 2952 Wellington Ave., Chicago 18, Ill.
Doctor, B. H., 34–58 74 St., Jackson Heights, L. I., N. Y.
Donics, W., 1632 Roberts Lane, Falls Church, Va.
Dresser, S. R., Jr., Derrick City, Pa.
Edwards, J. B., Jr., 2632 Reagan St., Dallas, Tex.
Edwards, N. R., Ala Wai Terr., Apt. 246, 1547 Ala Wai Blvd., Honolulu, Oahu, T. H.
Ertman, R. J., 1700 T St., S. E., Washington 20, D. C.
Ficek, J. A., 340 W. First St., Dayton 2, Ohio
Flour, W., 2453 W. Chicago Ave., Chicago, Ill.
Fong, L. F., 1043 Powell St., San Francisco 8, Calif.
Friedman, E., Trav-Ler Radio Corp., Orleans, Ind.
Gardner, C. L., 3069 N. 32 St., Kansas City 2, Kan.
Gollwitzer, L. H., 3507 Merrick, Houston, Tex.
Gordon, A. D., R.D. 4, Groffs Trailer, Lancaster, Pa.
Gray, J. S., 4069 W. 13 Ave., Vancouver, B. C., Canada
Grzywacz, W., Lacey Park, 14 Downey Dr., Hatboro, Pa.
Guba, R. F., 36 Marianne Rd., Waltham 54, Mass.
Hajny, E. J., 16900 Delaware Ave., Lakewood 7, Ohio
Harvey, J. D., 570 Lexington Ave., New York, N. Y.
Hay, E., 508 Winneconna Pkwy., Chicago, Ill.
Heitzer, J. B., 1002 E. Brady St., Butler, Pa.
Hoehn, A., 26253 Pattow Lane, Roseville, Mich.
Hogan, J. V., 307 Burns St., Forest Hills, L. I., N. Y.
Jacobs, G. G., 302 S. Frankfort, Tulsa, Okla.
Keane, T. P., 6149 Tujunga Ave., N. Hollywood, Calif.
Kephart, D. C., 1725 E. 32 St., Baltimore 18, Md.
Kingwill, H. S., 35 E. Wacker Dr., Chicago 1, Ill.
Kramer, M., 981 Dumont Ave., Brooklyn 8, N. Y.
Krogstad, M. O., 79 Herschel St., Providence, R. I.
Krohn, E. H., Krohn-Hite Instrument Co., 580 Massachusetts Ave., Cambridge 39, Mass.
Lebednik, M. A., 169 Peshine Ave., Newark 8, N. J.
Lendway, J. M., 10565 Vernon Ave., Huntington Woods, Mich.
Lindner, G. H., 41 John St., Red Bank, N. J.
MacDuff, G. R., 110 E. Vernor Hgwy., Detroit, Mich.
Martz, A. F., Jr., 1513 Hawthorne, Grosse Pointe 30, Mich
Mayberry, J. H., Box 4046, Corpus Christi, Tex.
McArthur, C. S., Jr., Box 5214 College Station, Raleigh, N. C.
McLawler, M. S., 908 E. 15 St., Indianapolis 2, Ind.
Milton, L., 21-17 80 St., Jackson Heights, L. I., N. Y.
Morrell, A., 1818 Grandview Ave., Portsmouth, Ohio
Murphy, T. P., Jr., Rt. 2, Box 175, Los Gatos Rd., Santa Clara, Calif.
Nakata, C. K., 2454 N. Orchard St., Chicago 14, Ill.

(Continued on page 60A)

MEMBERSHIP

(Continued from page 59A)

Nedved, O. M., 258 Northwood St., Riverside, Ill.
Newby, H. H., 905 Perry, Wichita, Kan.
Noll, J. W., 502 Boonton Ave., Boonton, N. J.
Padva, P. D., Armour Research Foundation, Technical Center, Chicago 16, Ill.
Page, F. T., 6220 Tramore Rd., Baltimore 14, Md.
Parnell, B., 36 Fisher Ave., Ottawa, Ont., Canada
Paton, I. W., 931 Government St., Penticton, B. C., Canada
Person, L. B., Technical Material Corporation, Box 142, Mamaroneck, N. Y.
Pierce, V. M., 134-12 Sutter Ave., S. Ozone Park, L. I., N. Y.
Randolph, A. M., United Gas Pipeline Company, Box 1407, Shreveport, La.
Robbins, T. N., Jr., 154 Broad St., Eatontown, N. J.
Rosenkoetter, E. A., 5564 A Hebert St., St. Louis, Mo.
Rust, J. J., 1001 Jourdan Ave., New Orleans 17, La.
Saito, S., 3318 Jones Bridge Ct., ChevyChase 15, Md.
Schrader, H. E., Palmer Laboratories, Princeton, N. J.
Schwabel, C. A., R.D. 1, Sandusky, Ohio
Sharp, W. C., CAA, City Hall Bldg., Kansas City 6, Mo.
Sibold, A. P., Jr., U. S. Naval Shipyard, Naval Base, S. C.
Sidwell, R. D., KUBC, Box 279, Montrose, Colo.
Silverman, M., 2052 69 St., Brooklyn 4, N. Y.
Sjoholm, E. M., Jr., Brilyn Park, 714 North St., Falls Church, Va.
Skaggs, E. P., 1441 S. Third St., Louisville 8, Ky.
Smith, A. D., Sr., 727 Maupin Ave., Salisbury, N. C.
Smith, E. C., 228 Maple Ave., Hershey, Pa.
Spencer, G. D., 394 Rugby Rd., Brooklyn 26, N. Y.
Stavrou, S., 36 Gibson Ave., Toronto 5, Ont., Canada
Stiefel, R. C., 206 W. 21 St., New York, N. Y.
Sweigart, A. S., 6243 N. Cardiff St., Philadelphia, Pa.
Toomey, J., Old Croton Rd., Ayer, Mass.
Weiner, M., 257 S. 16 St., Philadelphia 2, Pa.
Weisenfeld, S., 4906 N. Kenmore Ave., Chicago 40, Ill.
Williams, W. F., 4506 Randolph Rd., Silver Spring, Md.
Young, C. E., 309 Walnut St., Bellevue, Ky.

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 30A)

sizes; $4\frac{1}{2}$, $3\frac{1}{2}$, and $2\frac{1}{2}$ inches. The design is identical on all three sizes. Model numbers are, 1029, 1027, and 1127.

The scale provides increased readability and is protected by unbreakable plastic.

Plant Expansion

A new plant for producing ceramic capacitors at the rate of 1,000,000 per week was opened in Olean, N. Y., by **The Electrical Reactance Corp.**

The plant has 70,000 square feet of floor space, and will employ about 1,500 people, according to Charles E. Krampf, president.

(Continued on page 62A)

Let MARION help you...

...seal components *hermetically*
...speed up sub-assemblies

Marion portable, bench-type induction soldering unit....

- SMALL
- COMPACT
- ADAPTABLE
- EFFICIENT
- ECONOMICAL
- EASY-TO-USE

A PRODUCTION TOOL This unit was designed to simplify and improve both the quality and quantity of production of many different assemblies in the electronic and electrical fields. It has been used successfully for many years in the manufacture of Marion hermetically sealed instruments. Assembly applications include magnet assemblies, relay armatures, connectors, capacitors and transformer cans and germanium diode assemblies; also jewelry, watches, toys, automotive parts, household fixtures, etc.

HERMETICALLY SEALED COMPONENTS Because of the present intense interest in hermetically sealed components, Marion offers the benefit of its experience in true glass-to-metal hermetic sealing with the Marion Induction Soldering Unit and Marion metalized (platinum film) glass. Marion platinum film base glass has been developed to permit higher sealing temperatures, greater thermal shock range and resoldering if necessary.

PROFIT FROM EXPERIENCE Investigate now. Submit your requirements. We will be glad to supply samples and quotations. Ask for bulletin. Marion Electrical Instrument Company, 407 Canal Street, Manchester, New Hampshire.

MANUFACTURERS OF MARION RUGGEDIZED INSTRUMENTS

marion meters

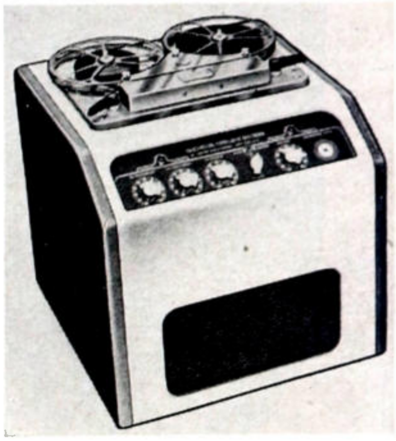
News—New Products

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

(Continued from page 60A)

New Tape Recorder

The Magnemaster Consolette, is the latest addition to the Twin-Trax Recorder series, manufactured and distributed by **Amplifier Corp. of America, 398-1 Broadway, New York 13, N. Y.**



Two-speed two-direction tape travel is available with this recorder. The recorder may be operated at 15 inches per second,

(Continued on page 65A)

CRYSTAL CALIBRATOR

MEASUREMENTS CORPORATION

Model 111

FREQUENCY RANGE: **.25Mc.—1000 Mc.**

FREQUENCY ACCURACY:

±0.001%

A Dual-Purpose Calibrator

- **CRYSTAL-CONTROLLED OSCILLATOR**
- **BUILT-IN DETECTOR 2 Microwatt Sensitivity**

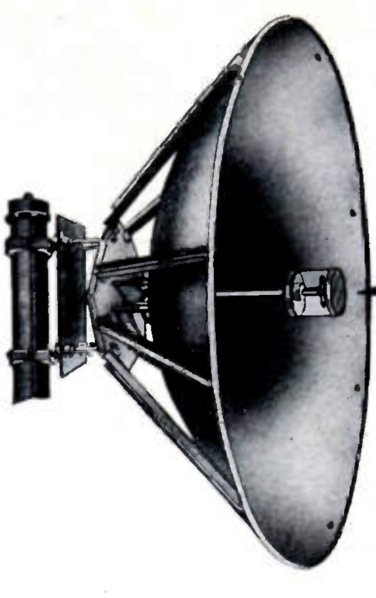
Designed for the Calibration and Frequency Checking of Signal Generators, Transmitters, Receivers, Grid-Dip Meters and other equipment where a high degree of frequency accuracy is required.

Harmonic Range:
.25 Mc. Oscillator: .25-450 Mc.
1 Mc. Oscillator: 1-600 Mc.
10 Mc. Oscillator: 10-1000 Mc.

117 volts, 50/60 cycles; 18 watts, 6" wide, 8" high, 5" deep; 4 lbs.

MEASUREMENTS CORPORATION

Boonton New Jersey



WHAT ARE *Your* REQUIREMENTS IN PARABOLIC ANTENNAS ?

For microwave systems . . . check these advantages of ANDREW Parabolic Antennas:

☐ **DEPENDABILITY** — An actual record of 100% dependability. There has never been a single mechanical or electrical failure on an ANDREW Parabolic Antenna . . . anywhere in the world.

☐ **COST** — Exceptionally low; made possible by high production.

☐ **LIGHT WEIGHT — HIGH STRENGTH** — Achieved by spun aluminum reflectors braced by formed steel struts.

☐ **ADJUSTABLE MOUNTING** — Through ± 10 degrees in azimuth and elevation.

☐ **DEICING KITS** — Thermostatically controlled, available where required.

☐ **CABLE** — 7/8" air dielectric Teflon insulated cable. Radiator is pressure tight. Fittings for solid dielectric cables also available.

SPECIFICATIONS

Frequency Range	. . . 890-960 MCS . . .				. . . 1750-2110 MCS . . .			
Type Number	1002	1004	1006	1010	2002	2004	2006	2010
Diameter of Parabola feet	2	4	6	10	2	4	6	10
Gain Over Half Wave Dipole Decibels	10	15	20	25	15	20	25	29
Beam Width, Half Power Points, Degrees	36°	22°	16°	11°	18°	10°	7°	5°
Net Weight, Pounds	10	64	150	380	10	65	150	380
Thrust Due to Wind Loading at 30 Pounds/FT Pounds	127	509	1145	3200	127	509	1145	3200

Your antenna problems can best be solved by ANDREW—the largest firm of antenna equipment specialists in the world. Write today.

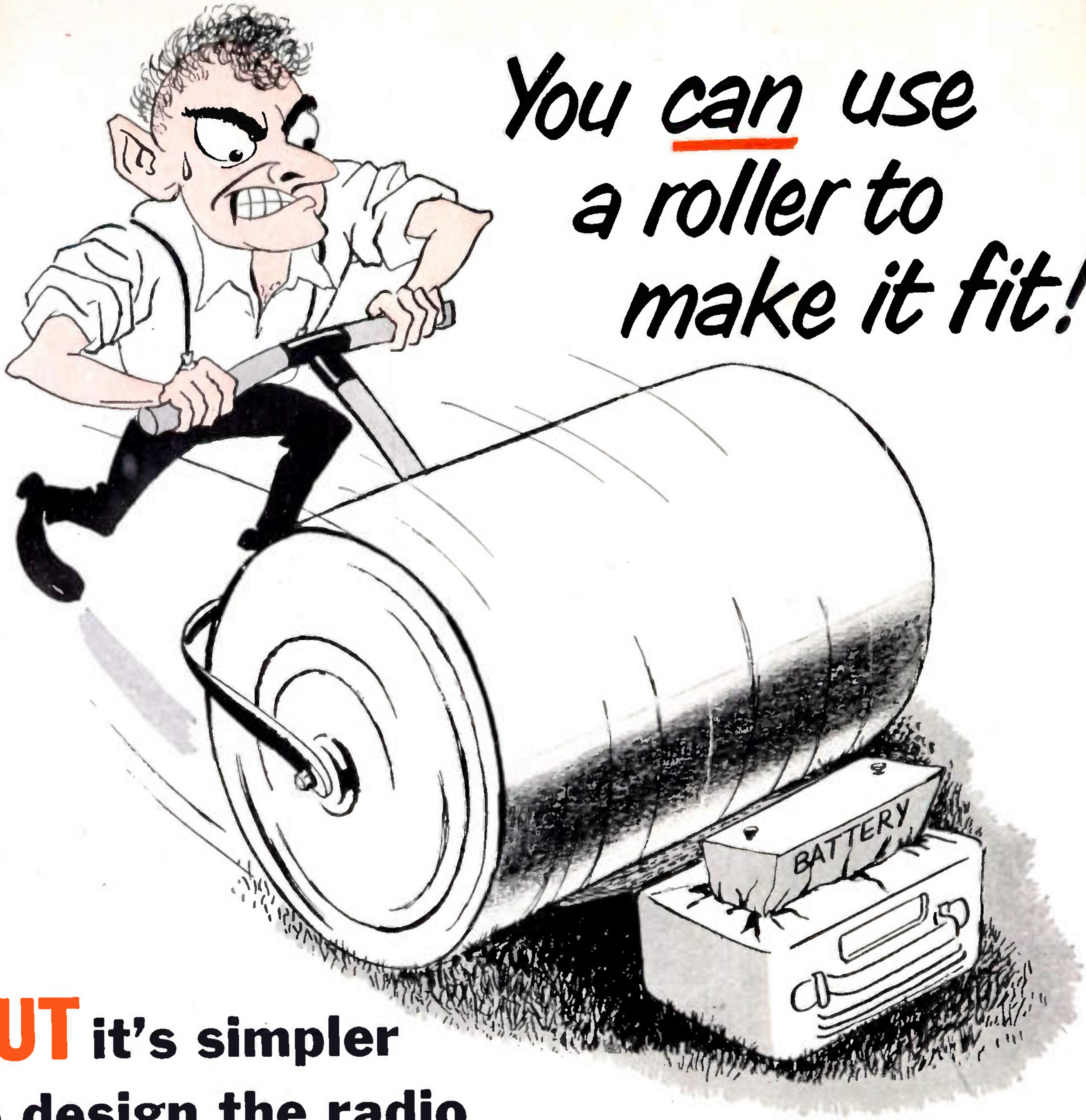
Andrew CORPORATION
363 EAST 75th STREET • CHICAGO 19

ANDREW

WORLD'S LARGEST ANTENNA EQUIPMENT SPECIALISTS

TRANSMISSION LINES FOR AM-FM-TV • ANTENNAS • DIRECTIONAL ANTENNA EQUIPMENT ANTENNA TUNING UNITS • TOWER LIGHTING EQUIPMENT

You can use a roller to make it fit!



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

Regardless of what size portable radio you are designing, you'll find compact, long-lasting "Eveready" batteries to fit it. "Eveready" brand batteries give longer playing life. They are the accepted standard for portable radios. Users can get replacements everywhere—they *prefer* portables that use "Eveready" batteries.

"Eveready" No. 950 "A" batteries and the No. 467 "B" battery make an ideal combination for small portable receivers.

CONSULT OUR BATTERY ENGINEERING DEPARTMENT FOR COMPLETE DATA ON "EVEREADY" BATTERIES

EVEREADY
TRADE-MARK
RADIO BATTERIES

"Eveready", "Mini-Max", "Nine Lives" and the Cat Symbol are trade-marks of

NATIONAL CARBON DIVISION
UNION CARBIDE AND CARBON CORPORATION
30 East 42nd Street, New York 17, N. Y.
District Sales Offices: Atlanta, Chicago, Dallas, Kansas City, New York, Pittsburgh, San Francisco

News—New Products

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

(Continued from page 62A)

with a frequency response of 50 to 15,000 cps; or at 7½ inches per second with a frequency response of 50 to 10,000 cps ±2 db. After the reel has run through, the instrument instantaneously and automatically changes direction of tape travel, and plays an equal length of time in the other direction.

A dynamic range of better than 50 db is attained without isolating the power stage. Flutter is rated at ±0.05 per cent at 15 inches per second; and ±0.1 per cent at 7½ inches per second.

The 10-tube recording amplifier, with supersonic bias oscillator and separate erase amplifier, and 3.5-watt monitor playback amplifier includes individual tone controls for accentuation and attenuation of bass and treble. Inputs are provided for radio-phono, and choice of high or low impedance microphone.

New Seismographic Indicator from 8-Channel Oscilloscope

To answer the seismographic problems of a government experimental agency, **The Electronic Tube Corp.**, 1200 E. Mermaid Lane, Philadelphia 18, Pa. has just developed a new eight-channel oscilloscope. Through the high-frequency response made possible by this instrument, users

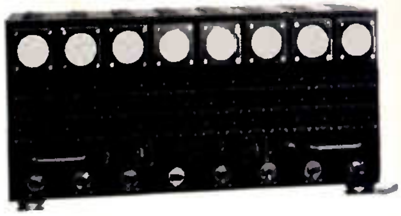
(Continued on page 66A)

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)

will be able to learn more about the characteristics of the earth than was formerly possible with a magnetic seismograph.



Designated Model H-81, the new oscilloscope consists of eight independent channels, each of which contains a single-gun cathode-ray tube, RMA Type 3 JP11; and eight ac amplifiers with a deflection sensitivity of 10 mv/in. Frequency response is 20 to 25,000 cps ± 2 per cent, or 20 to 150,000 cps ± 30 per cent. A dc model eight-channel oscilloscope, Model H-82, has been constructed also with a sensitivity of 2 mv/in. The only factors common to all the channels are intensifying of the cathode-ray tubes and timing marker injection.

Signals are displayed on a horizontal axis for photographing on a film strip or drum with vertical travel. Power supply is independent of the indicator unit.

(Continued on page 67A)

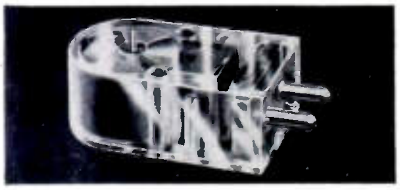
News—New Products

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)

Variable Reluctance Pickup

The Clarkstan Corp., 11921 W. Pico Blvd., Los Angeles 64, Calif., announces a new variable reluctance pickup. It is the junior model of their type RV cartridge which has been distributed during the past few years. This new pickup, model 204, has several features: the stylus is easily removable and interchangeable so that microgroove, standard, and transcription recordings can be played. Styli are available with ball points for all these various types of records.



The instrument offers a new technique in pickups in that the case is made of clear polystyrene which permits the operator to clearly see the inner workings of the pickup at any time. The output is high; only a customary preamplifier is required. The cartridge weighs 14 grams (½ ounce).

(Continued on page 68A)

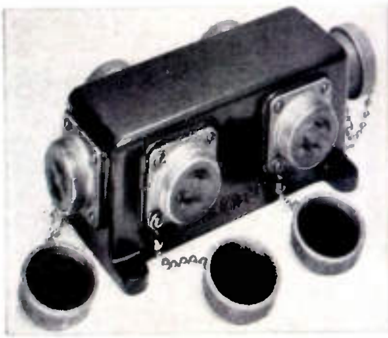
News—New Products

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

(*Continued from page 67A*)

Interruption-Proof All-Weather Outlet

The Type 6005 outlet box, when used with accessory connector, Types 106 and 115, gives the user waterproof, weatherproof and pull-out proof power connections. A complete illustrated bulletin is available from the manufacturer, **Equipment and Service Co.**, 6815 Oriole Dr., Dallas 9, Texas.



The inlet connector, GREENBILT Type 106, is rated for 6,000 watts, and the outlet connector, Type 115, is rated for 1,500 watts.

(*Continued on page 71A*)

News—New Products

(Continued from page 68A)

Minimum Inter-section Coupling in Multi-section Electrolytics

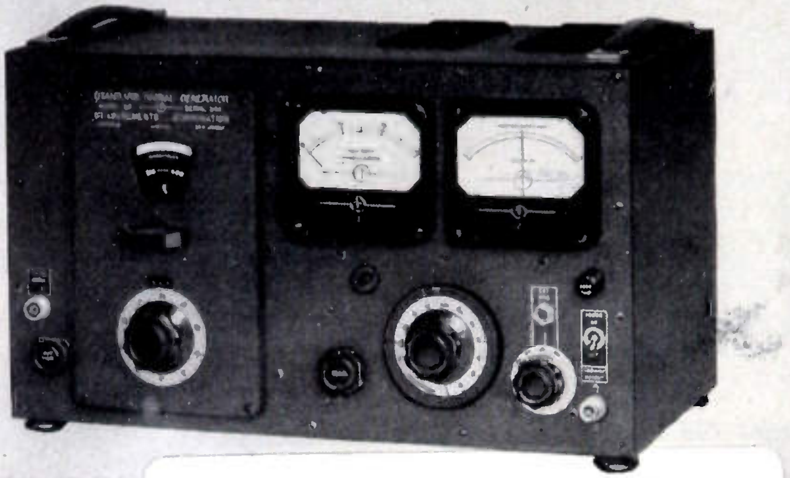
Marked reduction in hum and "hash" often experienced in multisection electrolytics because of interanode coupling and resultant internal cross modulation, is claimed for Type AFH, or twist-prong-base multisection electrolytics currently produced by **Aerovox Corp.**, New Bedford, Mass.

A special internal construction provides low rf impedance and minimum coupling between sections. This feature applies to the large selection of capacitance and voltage combinations in the Type AFH electrolytics which are suited for television applications and will withstand temperatures up to 85°C.

(Continued on page 73A)

MEASUREMENTS CORPORATION *MODEL 80*

STANDARD SIGNAL GENERATOR



2 to 400 MEGACYCLES

MODULATION: Amplitude modulation is continuously variable from 0 to 30%, indicated by a meter on the panel. An internal 400 or 1000 cycle audio oscillator is provided. Modulation may also be applied from an external source. Pulse modulation may be applied to the oscillator from an external source through a special connector. Pulses of 1 microsecond can be obtained at higher carrier frequencies.

FREQUENCY ACCURACY ±.5%

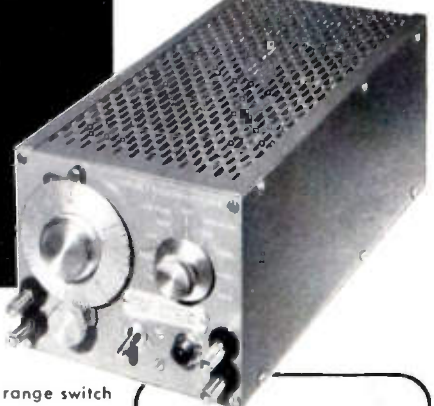
OUTPUT VOLTAGE 0.1 to 100,000 microvolts

OUTPUT IMPEDANCE 50 ohms

MANUFACTURERS OF
Standard Signal Generators
Pulse Generators
FM Signal Generators
Square Wave Generators
Vacuum Tube Voltmeters
UHF Radio Noise & Field Strength Meters
Capacity Bridges
Megohm Meters
Phase Sequence Indicators
Television and FM Test Equipment

MEASUREMENTS CORPORATION
BOONTON NEW JERSEY

MODEL 300
VARIABLE ELECTRONIC FILTER



Two simple controls are all that are necessary to operate the Model 300 Variable Electronic Filter. With the variable frequency dial and range switch any cut-off frequency from 20 cps to 200 KC may be quickly and accurately selected and reselected. With the range switch either low-pass or high-pass filter action may be chosen. In either case the rate of attenuation is 18 db per octave and the insertion loss 0 db. For higher rates of attenuation or continuous band pass operation two or more sections can be cascaded. Its low noise level and flexibility of operation make the Model 300 indispensable in geophysical and acoustic research, industrial noise measurements, in the automotive and aircraft industries as well as the radio broadcasting, recording and motion picture studio.

Write for further information today.

SPECIFICATIONS

- CUT-OFF RANGE — 20 cps to 200 KC
- ATTENUATION RATE — 18 db per octave
- SECTIONS — Single, can be high pass and low pass
- INSERTION LOSS — 0 db
- PASS BAND LIMITS — 2 cycles to 4 MC
- NOISE LEVEL — 80 db below 1 volt

SKL SPENCER-KENNEDY LABORATORIES, INC.
181 MASSACHUSETTS AVE., CAMBRIDGE 39, MASS.

SELENIUM RECTIFIERS

Proven 40 Volt Cells Represent More Watts For Your Rectifier Dollar

KOTRON alone manufactures selenium rectifiers to block from 6 to 40 volts A.C. rms per cell with optimum current rating for each voltage. Kotron is the first to produce commercial rectifiers embodying this relationship. In high voltage application, Kotron uses fewer cells than ordinary rectifiers—thus reducing size and cost. In low voltage applications, Kotron works at higher current densities per sq. in. of rectifying area than ordinary selenium rectifiers, reducing size and lowering cost.

In either direction, Kotron surpasses ordinary standards!

Kotron mechanical features match Kotron superior electrical characteristics. Square or rectangular plates offer maximum wattage in minimum space. Rectifier cells and terminals are mechanically interlocked to prevent twisting out of alignment in stack assembly (pat. pending). Double terminals, where necessary, facilitate connection to external circuits. Echelon terminal structure eliminates soldered or internal stack connections.

KOTRON rectifiers whether for 1 ampere or thousands of amperes, are custom-built and tailored to the application, giving maximum efficiency with minimum cost. Kotron rectifiers are manufactured by selenium rectifier specialists with years of technical experience. Their specialized engineering knowledge and consulting service is available without obligation.

Write for descriptive sheets. Submit specifications on your requirements for quotations.

KOTRON RECTIFIER CORP.
Humboldt 2-2400
54 Clark St. Newark 4, N.J.

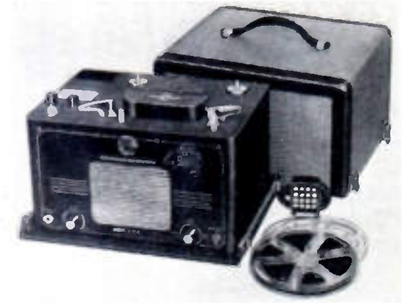
News—New Products

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

(Continued from page 71A)

New Tape Recorders

There are six models of the new "Sound Reel," dual-speed, dual-track, tape recorders manufactured by **Mark Simpson Mfg. Co., Inc.**, 32-28 49 St., Long Island City, L. I., N. Y.



All feature two hours of dual-track recording at 3.75 inches per second or one hour at 7.50 inches per second on a seven-inch reel of 1,200 feet of plastic tape. They come complete with 625 feet of tape, a 7-inch plastic take-up reel and microphone.

(Continued on page 74A)

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 73A)

The six models include three with self-contained AM tuner built into the housing. The tuner can be used as a separate radio when not recording

Aluminized 19-Inch Metal Picture Tube

The Tube Div., General Electric Co., Electronics Park, Syracuse, N. Y., has announced that production will begin soon on an aluminized 19-inch round metal picture tube (19AP4C) which will provide improved picture brightness, contrast, and detail.

This is the second large picture tube to be aluminized within the past month by GE, the other being the 16-inch glass rectangular.

The 19AP4C tube features an electron gun designed to be used with an external ion-trap magnet. The aluminized screen allows the tube to be operated at a lower anode voltage than is feasible with the non-aluminized version of the tube.

Electrical characteristics of the tube include: heater voltage, 6.3 volts; heater current 0.6 ampere ±10 per cent. Maximum ratings include: anode voltage, 19,000 volts; grid-No. 2 voltage, 410 volts.

(Continued on page 76A)

Where the Requirements are Extreme...

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For extraordinary electrical performance

GRAPHALLOY TRADE MARK REG. U. S. PAT. OFF.

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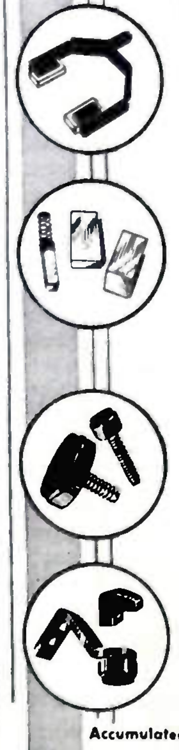

IN BRUSHES

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- low contact drop
- low electrical noise
- self-lubrication

IN CONTACTS

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Designed for low cost NE-51 Neon

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- U/L Listed
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Catalogue Number 521308 — 997 for 110 or 220 volts.

SAMPLES for design purpose NO CHARGE

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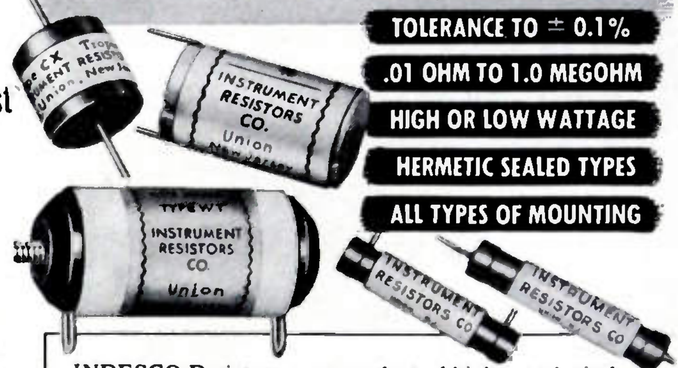

INRESCO Resistors are a product of high-speed winding techniques that introduce a new measure of economy in precision wire wound resistors.

They are available for IMMEDIATE DELIVERY, in diversified types that meet practically every circuit requirement of load, ohmic value, size, and shape.

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Manufacturers and designers of wire wound resistors—exclusively. Estimates on custom built resistors furnished without obligation; inquiries are invited.

INSTRUMENT RESISTORS COMPANY

1036 COMMERCE AVENUE, UNION, NEW JERSEY

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 74A)

Windowless Flow Counter

The first production run of SC-16 Windowless Flow Counters has been completed by **Tracerlab, Inc.,** 130 High St., Boston 10, Mass.

This instrument is a low background radiation counter designed for operation in either the Geiger or proportional regions. It is essentially a shielded counter tube into which solid samples are inserted directly, and through which a constant gas flow is maintained to prevent air contamination.

The absence of a window between the sample and the counting chamber makes this instrument useful for the detection and measurement of alpha radiation and weak beta radiation, such as that emitted by Carbon-14 and Sulfur-35.

A unique feature of this instrument is that it is equipped with a three-position rotating platform which has three recesses for holding standard size sample containers, such as planchets and brass rings and disks for filter paper. One position is for sample loading, one for pre-flushing and the third for counting.

When operated in the Geiger region, the flow counter does not distinguish between the various types of radiation, but when operated in the proportional region it is possible to count alpha particles separately in the presence of beta radiation.

Electrolysis, Corrosion, and Cathodic-Protection Tester

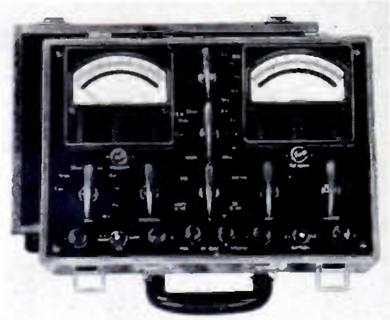
A new model, the B-3, of the multi-combination meter for electrolysis, corrosion, and cathodic-protection testing, is now available from the manufacturer, **M. C. Miller,** 1142 Emerson Ave., West Englewood, N. J. It is small, has a low range of 2 millivolts, and special mirror scales 3.9 inches long. The accuracy is 1 per cent.

(Continued on page 77A)

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 76A)



Wide current and potential ranges and a number of measuring circuits are provided, as follows: low-resistance, 1,000-ohm-per-volt voltmeter with 10 ranges from 2 mv to 100 volts; high-resistance, 62,500-ohm-per-volt voltmeter with six ranges from 0.1 to 20 volts; potentiometer voltmeter with six ranges from 2 mv to 3 volts; dc biasing potential in series with high-resistance voltmeter to "back-to-zero" residual earth, pipe or electrode potentials to permit direct readings of "change-in-potentials" up to 20 volts; milliammeter/ammeter with nine ranges from 2 milliamperes to 20 amperes, low internal resistance, built-in current switch and current control rheostats.

MICROWAVE

ARMY-NAVY TEST SETS

LU-1 Freq. Meter and Test Oscillator. Type CRV-TVN-9HU Power Supply. M.I.T. Rad. Lab.
TVN-8SE Klystron Power Supply. M.I.T. Rad. Lab.
CS60ABW Watt Meter—Wavemeter. 3 centimeter.
APR5 Receiver—1000 to 6000 mcs.
AN/CPN-8—10 centimeter 40 kw. output RF package. Includes magnetron oscillator, complete modulator, complete receiver, complete signal and power analyzer with 5" scope. 115V AC input.
Dehydrator Unit CPD 10187 Automatic cycling.
Compressor to 50 lbs. Compl. for Radar XSMN. Line New $425.00
SO-3 Receiver, 30 mc. IF. 6 stages 6AC7, 10 MC. Band width inpt. 5.1 mc B.W. per stg. 9.6 volt gain per stage as desc. in ch. 13 vol. 23 M.I.T. Rad. Lab. Series $99.50
TS12 3 cm Slotted Line
TS13 3 cm Signal Generator
TS33 3 cm Frequency Meter
TS35 3 cm Signal Generator
TS36 3 cm Power Meter
TS45/AMP3 3 cm Signal Generator
TS63 3 cm Echo Box
TS108 Dummy Load
I—185A oscillator
I—158 Range Calibrator
I—233 Range Calibrator
BC—438 Freq Meter
RF—Preamp
GR—Capacity bridge $216
GR—Uni Galvo Shunt $229
GR—1000 Avd Osc $213
TS 226A/AP Power Meter 0-1000W
Sig Gen 3804 8-330 MC
BC 221 Freq Meter
TS 10/AP
TS 69/AP Freq Meter
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I126
Field Int. Meter 43A
Sig Gen Mod. 78B
Megometer 0-5 mgr
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D-171812	$1.50
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D-168442	$3.00
D-165593	$1.25
D-98836	$2.00
D-161871A	$2.85
D-171121	$1.50
D-162356(308A)	$1.50
D-163357	$2.00
D-99946	$2.95

THERMISTORS

D-167332 (tube)	$1.50
D-170396 (bead)	$1.50
D-167613 (button)	$1.50
D-164699 for MTG in "X" band Guide	$2.50
D-167018 (tube)	$1.50

MAGNETRON MAGNETS

Gauss	Pole Diam.	Spacing	Price
4850	¾ in.	⅝ in.	$12.50
5200	21/32 in.	1 5/16 in.	$17.50
1300	1⅝ in.	1 5/16 in.	$12.50
1860	1⅝ in.	1½ in.	$14.50

Electromagnets for magnetrons ....$24.50 ea.
GE Magnets type M7765115, GI Distance Between pole faces variable. 2 1/16" (1900 Gauss) to 1⅛" (2200 Gauss) Pole Dia. 1⅝". New Part of SCB 584.....$34.50

MAGNETRONS

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QK 60	2J37	2J62	725-A	
2J21	2J38	3J31	730-A	
2J22	2J39	5J30	728	
QK 915	2J26	2J40	714AY	706
QK 62	2J27	2J49	718DY	700
QK 59	2J31	2J34	720BY	

Klystrons 723A, 707B, 417A, 2K41

PULSE TRANSFORMERS

G.E.K.-2745 ....$39.50
G.E.K.-2744-A. 11.5 KV High Voltage, 3.2 KV Low Voltage @ 200 KW oper. (370 KW max.) 1 microsec. or ¼ microsec. @ 600 PPS $39.50
W.E. $166173 Hi-Volt input transformer, W.E. Impedance ratio 50 ohms to 900 ohms. Freq. range: 10 kc. to 2 mc. 2 sections parallel connected, potted in oil $36.00
W.E. KS 9800 Input transformer. Winding ratio between terminals 3-5 and 1.2 is 1.1:1, and between terminals 6-7 and 1.2 is 2:1 Frequency range: 380-520 c.p.s. Permalloy core $6.00
G.E. $K2731 Repetition Rate: 635 PPS. Pri. Imp: Ohms. Sec. Imp: 450 Ohms Pulse Width: 1 Microsec. Pri. Input: 9.5 KV PK Sec. Output: 28 KV PK. Peak Output: 800 KW Bifar 2.75 Amp. $64.50
W.E. $D169271 Hi Volt input pulse Transformer $27.50
G.E. K2450A. Will receive 13KV, 4 micro-second pulse on pri. secondary delivers 14KV Peak power out 100 KW G.E....$34.50
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RAY-WX-4298F $32.50
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22 Ft. Trailer Complete. Tracking & 10 CM Search Write

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RELAY SYSTEM PARABOLIC REFLECTORS: approx. range: 2000 to 6000 mc. Dimensions: 4' x 3' rectangle, now ..$35.00
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SC-13 ANTENNA 24" dish with feed-back dipole 360 deg. rotation, complete with drive motor and selsyn. New $125.00
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DBM ANTENNA. Dual. back-to-back parabolas with dipols. Freq. coverage 1,000-4,500 mc. No drive mechanism $65.00
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140-600 MC CONE type antenna, complete with 25' sectional steel mast, guys, cables, carrying case, etc. New $49.50
ASD 3 cm. antenna, used, ex. cond. $49.50
YAGI ANTENNA AS-46A. APG-4, 5 elements ....$14.50 ea.
DISH FOR PARABOLA 30" $4.85
AN 122 ....$22.50
RC 224 ....$99.50
AS 17/APS
OAK Directional Finder Loops
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200 MC. Bed Spring
LP21-A Airborne Dir. Find. Loop
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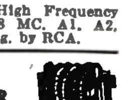
RADIO SYSTEMS

White Radio Telephone Model $WRM55—ship to ship—shore to ship—small airports—mines—plantations — inter island — ranches, 10 channel fix tuned rec. & Xmitter. Xmitter, pwr. output in excess of 100W unmodulated into antenna of 18 & 100 MMF. Freq. Range 2-12 MC. Can be modified to increase range. Xtal controlled. 110V 60 cy or 220V 60 or 25 cy. Meas. 24" H x 19" W x 14" D. 125 lbs. Write or phone for data.
TAJ 500 Watt Low Frequency Transmitter 150-550KC CW-MCW.
TBK 500 Watt High Frequency Transmitter 2-18 MC. A1, A2, A3, Emission Mfg. by RCA.

FILAMENT TRANSFORMER



for above 115V/60 cy
Pri: four 6.3V/4A
Sec. 500VT .$27.50
Magnetron Kit of four QK's 2675-3375 inc. w/trans special $250.00

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3 CM AIRBORNE
APS-15
3 CM AIRBORNE
APS-31
3 CM AIRBORNE
CPN-8
10 CM GROUND BEACON
CEXH
10 CM GROUND BEACON
FD-MARK 4
800 MC SHIP GUNLAYING
MARK-10
10 CM SHIP GUNLAYING
SA
200 MC SHIP AIR SEARCH
SC
200 MC SHIP AIR SEARCH
SD
200 MC SUBMARINE
SE
10 CM SURFACE SEARCH
SF
10 CM SURFACE SEARCH
SG
10 CM SURFACE SEARCH
SI
10 CM SUBMARINE
SK
200 MC AIR SEARCH
SL
10 CM SURFACE SEARCH
SN
10 CM PORTABLE
SO (ALL SERIES)
10 CM SURFACE SEARCH
SQ
10 CM PORTABLE
SW
200 MC PORTABLE
SCR 518
400 MC ALTIMETER
SCR 520
10 CM AIRBORNE
SCR 533
200 MC AIR SEARCH
SCR 545
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SCR 663
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QBG
QC
QCJ
QCL
QCO
QCS
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WEA

3000 MC BENCH TEST PLUMBING

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10 CM Wavemeter WE type B435490 Transmission type. Type N Fittings. Veeder Root Micrometer dial., Gold Plated W/Calib. Chart P/o Freq. Meter X66404A. New $99.50
AS14A/AP-10 CM Pick up Dipole with "N" Cables $4.50
LHTR. LIGHTHOUSE ASSEMBLY. Part of RT-39 APG 5 & APG 15. Receiver and Trans Cavities w/assoc. Tr. Cavity and Type N CPLG. To Revr. Uses 2C40, 2C43, 1B27. Tunable APX 2400-2700 MCS. Silver Plated $49.50
Beacon Lighthouse cavity 10 cm with miniature 28 volt DC FM motor. Mfg. Bernard Rice ....$47.50 ea.
Magnetron to Waveguide Coupler with 721A Duplexer Cavity, gold-plated $45.00
721A TR Box complete with tube and tuning plungers $12.50
McNally Klystron Cavities for 707B or 2K28. Three types available $4.00
F-29/SPR-2 Filters, Type "N", input and output $12.50
WAVEGUIDE TO ⅞" RIGID COAX "DOORKNOB" ADAPT-
ER. CHOKE FLANGE, SILVER PLATED BROAD BAND $32.50
WAVEGUIDE DIRECTIONAL COUPLER, 27 db. Navy type CABV-47AAN, with 4 in. slotted section $32.50
SQ. FLANGE to rd choke adapter, 18 in. long OA 1½ in. x 3 in. guide, type "N" output and sampling probe $27.50
AN/APR5A 10 cm antenna equipment consisting of two 10 cm waveguide sections, each polarized, 45 degrees ..$75.00 per set
POWER SPLITTER: 726 Klystron input dual "N" output ..$5.00

⅞" RIGID COAX

10 CM FEEDBACK DIPOLE ANTENNA, in lucite ball, for use with parabola ⅞" Rigid Coax Input $8.00
721A TR cavities, heavy silver plated ....$2.00 ea.
Magnetron Coupling with TR Loop, ...$7.50
Sperry Rotating Bend, pressurized $22.50
5 Ft. Lengths Stub Supported, gold-plated, per length ..$7.50
Short Right Angle Bends (for above) $2.50

8500 Mc. to 9600 Mc. Bench Test Plumbing

1" x 1½" WAVEGUIDE

VARIABLE STUB TUNER, DB536, 180 degree phase shifting capacity $70.00
LOW POWER TERMINATION, DB381 ....$18.50
UNI-DIRECTIONAL COUPLER, DB390, 23DB type "N" output $18.50
PICK UP HORN, Type "N" output $4.50
WAVEMETER, 8500 to 9400 mcs. with calibration. Micrometer adjust head. Reaction type $95.00
WAVEGUIDE LENGTHS, Plated and fitted with couplings available in 6", 12", 24", 36", 60" sections ....$2.00 per ft.
90 DEGREE ELBOWS, E or H plane, 2½" radius $12.50
90 DEGREE TWIST, 6" long $8.00
BULKHEAD FEED—THRU ASSEMBLY $15.00
PRESSURE GAUGE SECTION, 15 lb. gauge and press. nipple $10.00
PRESSURE GAUGE, 15 lbs. $2.50
DUAL OSCILLATOR-BEACON MOUNT, P/O APS10 Radar for mounting two 723A/B klystrons with crystal mts. matching slugs, shields $42.50
DUAL OSCILLATOR MOUNTS, (Back to back) with crystal mount, tunable termination, attentuating slugs $19.50
Directional Coupler, UG-40/U Take off 20 DB $17.50
Directional coupler, APS-6 type "N" take off 20 DB calibrated $17.50
Rotary Joint Choke to Choke $10.00
2K25/723 AB Receiver local oscillator Klystron Mount, complete with crystal mount, Iris coupling and choke coupling to TR $22.50
TR-ATR Duplexer section for above $8.50

1¼" x ⅝" WAVEGUIDE

TUNABLE TERMINATION, Precision adjust $65.00
LOW POWER TERMINATION $25.00
MAGIC TEE $45.00
OSCILLATOR MOUNT, for four 723AB klystron $38.50
90 DEGREE ELBOWS, E or H plane $12.50
WAVEGUIDE LENGTHS, Cut to size and supplied with 1 choke, 1 cover, per length $2.00 per ft.
WAVEMETER: Absorption type. Precision micrometer adjust. Very high Q $150.00
B1-D1-Coupler WG output calibrated—25 db nominal ..$17.50
Flex sections, 12" Rubber Coated $14.50
Mitred Elbow H Plane Ug51-UG52 ....$12.00
6" St. sect. choke to choke $3.50
APQ 13 Constant Z Rotat Jnt. $22.50

4000 to 6000 mcs. Bench Test Plumbing 2" x 1" WAVEGUIDE

FLAP ATTENUATOR $45.00
VARIABLE STUB TUNER AND LOW POWER TERMINATION $40.00
WAVEMETER TEE $48.00
ADAPTERS:
Choke to choke $18.00
Cover to cover $14.00
Choke to cover $10.00
WAVEGUIDE TO TYPE "N" ADAPTER $45.00
DIRECTIONAL COUPLER, Two hole type, type "N" output $40.00
KLYSTRON MOUNT, Equipped with tunable termination and micrometer adjust. klystron antenna tuning $110.00
CRYSTAL MOUNT, Equipped with tunable termination and micrometer adjust crystal tuning $125.00

TUNABLE TERMINATION, Precision adjust $90.00

6000 Mc. to 8500 Mc. Bench Test Plumbing

1½" x ¾" WAVEGUIDE

KLYSTRON MOUNT, DB356 complete with shield and tunable termination $125.00
FLAP ATTENUATOR, DB361 ....$45.00
VARIABLE STUB TUNER $90.00
WAVEGUIDE TO TYPE "N" ADAPTER $10.50
WAVEMETER TEE DB352 ....$32.50
MAGIC TEE $80.00
DIRECTIONAL COUPLER, Two hole 25DB coupling, type "N" output $25.00
PRECISION CRYSTAL MOUNT, Equipped with tuning slugs and tunable termination $125.00
TUNABLE TERMINATION, Precision adjust $70.00
LOW POWER LOAD $35.00

23,000 TO 27,000 MC. BENCH TEST PLUMBING

½" x ¼" WAVEGUIDE

PRECISION SLOTTED LINE, Adjustable probe $200.00
DIRECTIONAL COUPLER-Wavemeter Mount, 12DB $60.00
PRECISION VARIABLE ATTENUATOR, mfg. Bernard Rice $90.00
LOW POWER LOAD $20.00
SHUNT TEE $35.00
WAVEGUIDE LENGTHS, 2" to 6" long, goldplated with circular flanges and coupling nuts ....$2.25 per inch
APS-34 Rotating Joint $49.50
Right Angle Bend E or H Plane, specify combination of couplings desired $12.00
45° Bend E or H Plane, Choke to cover $12.00
Directional coupler CU-103/APS33 ....$40.50
Mit. Elbow, cov. to cov. ..$4.00
TR-ATR Sect. choke to cov. $4.00
Flex. Sec. 1" choke to choke $5.00
"S" Curve Choke to Cover $4.50
Adaptor, rnd. to sq. cov. ..$5.00
Feedback to Parabola Horn with pressurized window $27.50
Low Power Load, less cards $18.50
K Band Mixer Block $45.00
Waveguide ½ x ¼" $1.00 per ft.
90° Twist $10.00
"K" Band Directional Coupler CU104/APS-34 20 DB $40.50 ea.
3J31 Magnetrons $55.00

High Voltage Power Supply

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INDEX AND DISPLAY ADVERTISERS

DISPLAY ADVERTISERS

PROFESSIONAL CARDS

INDEX AND DISPLAY ADVERTISERS—Continued

CORRECTION NOTICE

The price of the Spencer-Kennedy Laboratories Model 212 T V Amplifier as given on page 81A in December PROCEEDINGS OF THE I.R.E. should be $220.00 f.o.b. Cambridge, Mass.

they may look alike, *but:* there is only one C-D

"VIKANE" IMPREGNATED

"GREY TIGER"

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For further details on the "Grey Tiger" line of paper tubulars, write for Bulletin No. NB116. CORNELL-DUBILIER ELECTRIC CORPORATION, Dept. M-11 South Plainfield, New Jersey. Other plants in New Bedford, Brookline and Worcester, Mass.; Providence, R. I., Indianapolis, Ind., and subsidiary, The Radiart Corp., Cleveland, Ohio.

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TYPE 1303-A
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Ideal for Non-Linear Tests On: ★ *Audio Amplifiers* ★ *Hearing Aids* ★ *Filter Networks* ★ *Noise Suppressors* ★ *High-Efficiency Speech Reproducing Systems* ★ *Loudspeakers* ★ *F-M Systems with Pre-Emphasis* ★ *Recording Systems* ★ *Any System of Restricted Frequency Range*

The new G-R Type 1303-A Two-Signal Audio Generator supplies signals by the beat-frequency method. Three oscillators and three mixers are used to provide a number of output-signal combinations. The output of the mixers are combined in a linear adding network and then amplified through a very low-distortion power amplifier. The output from the amplifier is fed into a 600-ohm attenuator system, with a voltmeter to monitor the level at the input of the attenuator. The harmonic content and inter-modulation products in the final output are at a very low level. High stability of voltage and frequency are provided. The frequency drift from cold start is only a few cycles.

This A-F Signal Generator will supply the following signals:

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- Two low-distortion sinusoidal voltages, each separately adjustable, one to 20 kc and the other to 10 kc.
- Two low-distortion sinusoidal voltages with fixed difference in frequency maintained between them as the frequency of one is varied. The fixed difference frequency is adjustable up to 10 kc, and the lower of the two frequencies is adjustable up to 20 kc.

The output is continuously adjustable and is calibrated both in volts and in db with respect to 1 mw into 600 ohms. The frequency calibration can be standardized within one cycle at any time. Its accuracy is ± (1% + 0.5 cycle).

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2. The intermodulation method that evaluates distortion in terms of the resultant modulation of a high-frequency tone by a low-frequency tone.

3. The difference-frequency intermodulation test, which evaluates distortion in terms of the amplitude of the difference-frequency components produced by intermodulation of two sinusoidal test signals of equal amplitude.

Write for Complete Information

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