Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

June, 1950
Volume 38 Number 6

Andrew Corp., Chicago, Ill.

RADIO "PIPE FITTERS"

Television coaxial transmission lines are produced in a plant bearing little resemblance to earlier conventional facilities for making any radio component.

PROCEEDINGS OF THE I.R.E.

Table of Contents, Indicated by Black-and-White Margin, Follows Page 32A

The Institute of Radio Engineers

for every application

While the catalogue line of UTC components covers a wide variety of applications, many people are not familiar with the full range of products produced by UTC. It is impossible to describe the thousands of special UTC designs as they become available. The illustrations below are intended to indicate some of the range in size of these special products.

A This 100 cubic foot modulation transformer is for 50 Kw. broadcast service. Frequency response flat from 30 to 20,000 cycles.

B The high Q toroid coil shown is 12" in diameter. It operates in a 50 Kw. circuit at supersonic frequency.

C This sub-miniature (.18 cubic inch) output transformer is intended for hearing aid and other extreme compact service. While the dimensions are only 7/16" x 9/16" x 3/4", the fidelity is ample for voice frequency requirements.

D This sub-miniature (.18 cubic inch) permalloy dust core toroid is available in a wide range of inductances, and for frequencies from 1,000 cycles to 50 Kc.

Write for catalog 500

United Transformer Co.

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

Chicago IRE's 25 Years of Progress Meeting

Remember this picture? It is the 1949 N.E.C. Committee which laid the groundwork for the 1950 Conference at Edgewater Beach Hotel.

THE NATIONAL ELECTRONICS CONFERENCE

September 25-27, 1950
Edgewater Beach Hotel
Chicago, Illinois

Technical Sessions on:

- Quality Control by Electronics
- Application of Magnetic Amplifiers to Electronic Control
- Nucleonics and Nuclear Instrumentation
- Use of New Circuits in Electronics
- Electro-Acoustics
- Applications of Analog and other Electronic Computors
- Industrial Television
- Flight Control by Electronics
- Dielectric Heating
- Micro-wave Spectroscopy

Born as a cooperative venture to further the war effort in 1945, The National Electronics Conference is jointly sponsored by the Chicago Sections of the IRE and AIEE, together with the Illinois Institute of Technology, and Northwestern University, and other cooperating organizations. This Sixth Conference promises to be the highlight of the 1950 Silver Anniversary of the Chicago IRE Section.

EXHIBITS with a purpose

Increased exhibit space is provided this year to permit manufacturers in or selling to Chicago to show their own contributions in equipment to Chicagolands 25 Years of Progress. About four units are open. Contact Kipling Adams, Exhibits Chairman, Room 212, at 920 South Michigan Avenue, Chicago 5, Phone: WAbash 2-3820. The Exhibition will be 38% larger than ever before!

Old Timers' Nite

Tuesday Evening, September 26

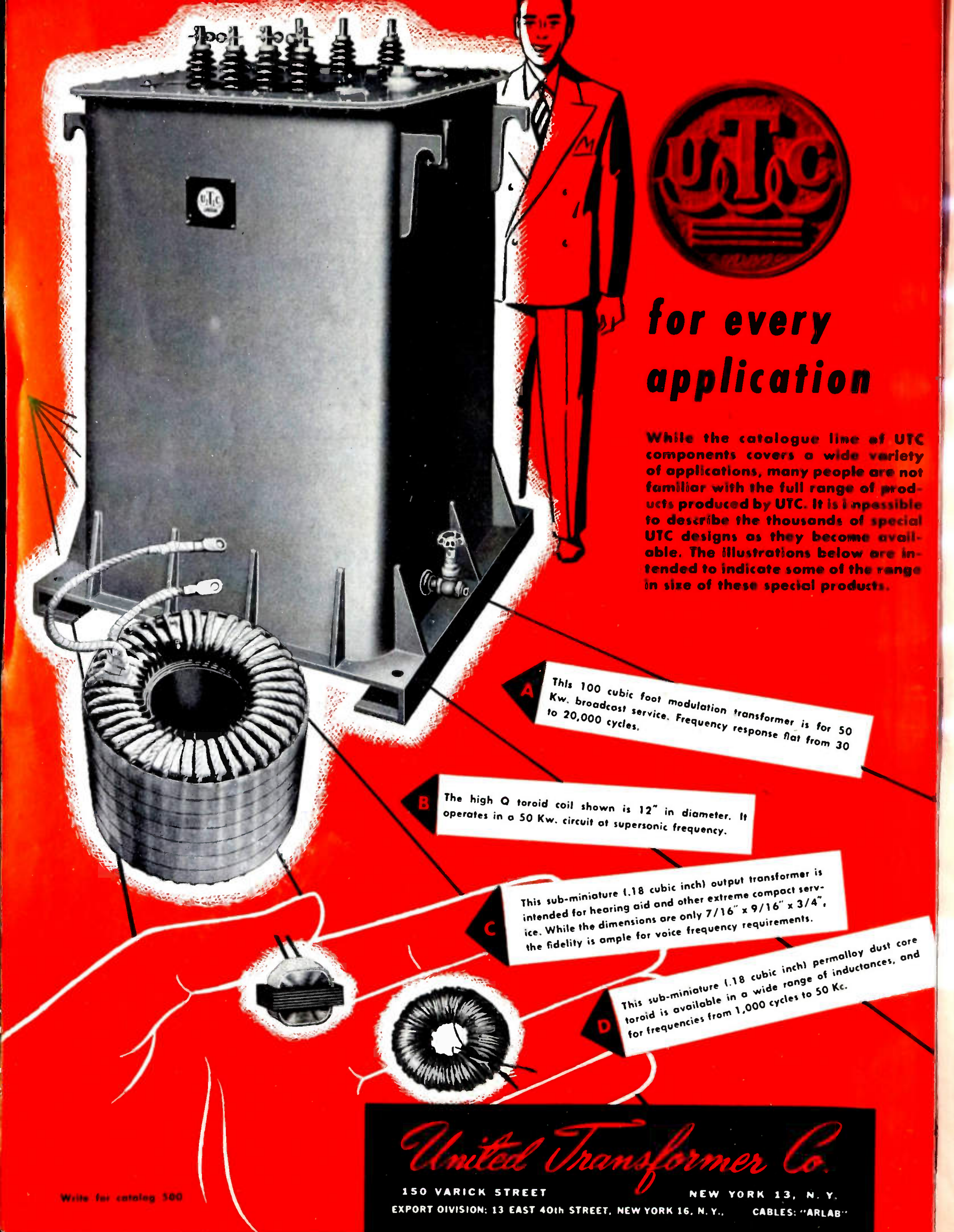
Well deserved, good fun is added to the Conference by a social event of true relation to IRE's 25 Years of Progress in Chicagoland. As described, "It is an opportunity to meet honored guests whose names are milestones in the history of electronics." Nathan Cohn is President of N.E.C. for 1950. Karl Kramer is Executive Vice-President. E. H. Schulz is I.R.E. Section Chairman and Alois W. Graf is Chairman of the Directory Committee.

INSTITUTE OF RADIO ENGINEERS
25 YEARS PROGRESS 1950
CHICAGO SECTION

IRE Regional Meetings Promote Electronic Progress

PROCEEDINGS OF THE I.R.E. June, 1950, Vol. 38, No. 6. Published monthly by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price $2.25 per copy. Subscriptions: United States and Canada, $18.00 a year; foreign countries $19.00 a year. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

Table of Contents will be found following page 32A



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INSTITUTE OF RADIO ENGINEERS · CHICAGO · SECTION
25
YEARS
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1950

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This picture was taken in the Bridgeport office of The Southern New England Telephone Company, one of the 22 operating telephone companies which the Laboratories serve through the Bell System.

Cutting Board for Telephone Costs

Few of these tools have sharp edges. But they are powerful cost cutters. Whenever a telephone craftsman reaches for one, he finds the right tool ready to his hand. There's no time wasted trying to do a complicated job with makeshift equipment.

Most telephone tools are highly specialized. 90% of dial system tools were designed by Bell Laboratories. Each saves time in maintenance, installation or construction.

There are tools with lights and mirrors to work deep within relay bays; tools to brush, burnish and polish; tools that vacuum clean—even a tool to weld on new contact points without dismantling a relay. There are gauges to time dial speeds, others to check spring tension. Some look like a dentist's instruments. Some you have never seen.

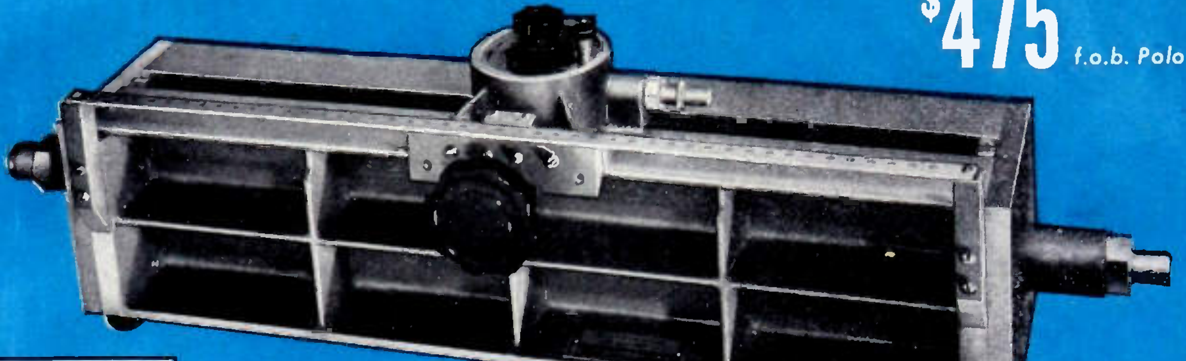
Keeping the telephone tool kit abreast of improvements is a continuing job for Bell Telephone Laboratories. It's another example of how the Laboratories help keep the value of your telephone service high, the cost low.

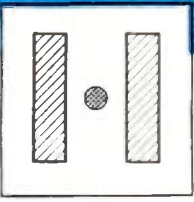
BELL TELEPHONE LABORATORIES

WORKING CONTINUALLY TO KEEP YOUR TELEPHONE SERVICE BIG IN VALUE AND LOW IN COST hp 805A SLOTTED LINE

PRECISION ACCURACY FOR STANDING WAVE MEASUREMENTS

$475 f.o.b. Palo Alto





RADICAL NEW "PARALLEL-PLANE" DESIGN GIVES -hp- SLOTTED LINE UTMOST ELECTRICAL STABILITY

The new *-hp-* 805A Slotted Line employs two parallel planes and a large, circular central conductor, instead of the conventional coaxial configuration. This new design makes possible an electrically stable precision instrument capable of fast, easy measurements of unvarying accuracy. Parallel planes and central conductor are both mechanically rigid. Penetration depth of the probe is less critical than in coaxial slotted lines, and leakage is low because the effective slot opening is less than .001 referred to the coaxial system. Residual VSWR is held to less than 1.04. Probe position may be read to 0.1 mm.

This new approach to the Slotted Line problem makes possible the manufacture of an instrument of maximum accuracy at moderate cost.

SPECIFICATIONS

Frequency Range: 500 to 4,000 mc.

Impedance: 50 ohms.

Connections: Special Type "N" fittings designed for minimum VSWR.

Residual VSWR: 1.04 or better.

Slope: Negligible.

Calibration: Metric, in cm and mm. Vernier reads to 0.1 mm.

Size: 27" long, 8" high, 6" wide.

Carriage: Ball-bearing probe movement. Probe depth adjustable. Probe resonant circuit tunable over freq. range of line. Detector may be standard crystal or employ barretters.

Data subject to change without notice.

WRITE FOR DETAILS

HEWLETT-PACKARD COMPANY
1824-D Page Mill Road · Palo Alto, California

1824

hp laboratory instruments
FOR SPEED AND ACCURACY

NEW -hp- 415A

Standing Wave Indicator

The new *-hp-* 415A Standing Wave Indicator is used with the *-hp-* Slotted Line to determine coaxial flatness or measure impedance. It consists of a high gain amplifier of low noise level, operating at a fixed audio frequency. Amplifier output is measured by a voltmeter with a square-law calibration in db and voltage standing wave ratio. The *-hp-* 415A is direct reading, compact and easy to use.

SPECIFICATIONS

Frequency: Fixed at 1,000 cps, ± 2%. Other frequencies 300 to 2,000 cps supplied on special order. Amplifier "Q" is 20 ± 5.

Sensitivity: 0.3 uv gives full scale deflection. Noise-level-to-input equivalent is 0.04 uv.

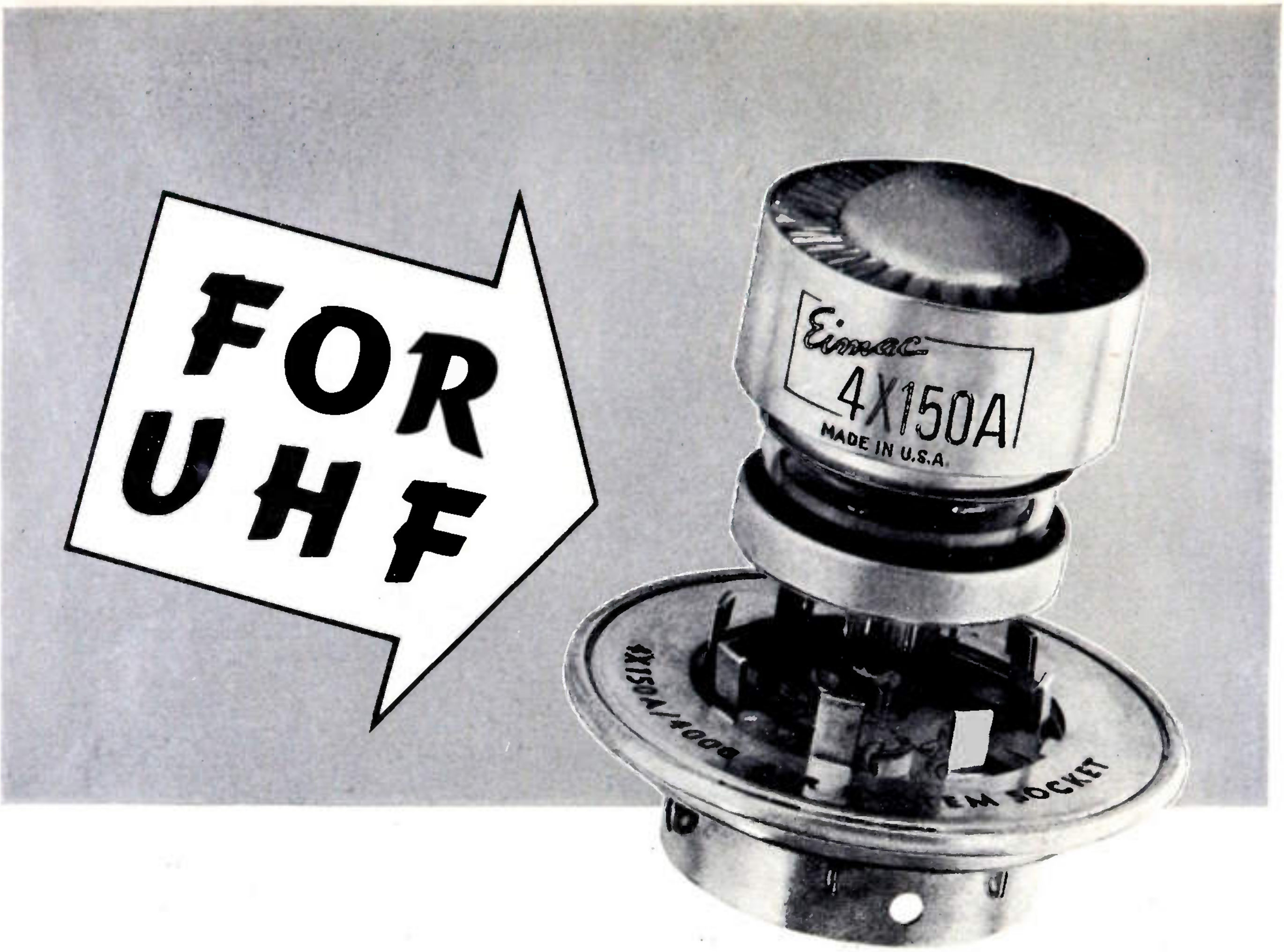
Calibration: For use with square-law detector. 60 db level covered in 6 ranges. Accuracy ± 0.1 db per 10 db step.

Gain Control: Adjusts meter to convenient level. Range is approx. 30 db.

Detector Input: Connects to Xtal rectifier or bolometer. Bias of 8 v. ± .5 v. delivers approx. 8.75 ma. to a 200 ohm barretter.

Size: 12" long, 9" wide, 9" high.

Data subject to change without notice.



Now available . . . a UHF, tube and socket package to solve your UHF tube and tube-cooling problems.

The combined use of the Eimac 4X150A tetrode and the new Eimac 4X150A socket makes possible improved circuit arrangement especially at frequencies between 100 and 500 Mc. and also simplifies mechanical design of the tube cooling system.

The tube . . . type 4X150A is a highly efficient beam-power Eimac tetrode capable of handling 150 watts of plate dissipation and delivering as high as 140 watts of useful output power per tube in conventional coaxial amplifier circuits. Its high degree of stability, high power-gain, and high ratio of transconductance to capacitance make it ideally suited for service as a video amplifier, TV sound amplifier, FM & TV r-f amplifier, or in UHF communications, and in STL and dielectric heating applications.

The socket . . . type 4X150A/4000, in addition to insuring adequate cooling of the 4X150A, simplifies circuit construction. It incorporates a 3750 $\mu\mu$f screen bypass capacitor and its terminal design reduces lead inductance to a minimum. The 4X150A/4000 socket is engineered for service in either coaxial line or chassis construction.

Take advantage of the tetrode engineering experience of America's foremost manufacturer . . . Eimac. Write today for complete data on the 4X150A, 4X150A/4000 socket and other high performance tubes contained in the new Eimac tube catalogue.

255

Follow the Leaders to
Eimac
TUBES
The Power for R-F

EITEL-McCULLOUGH, INC.
San Bruno, California
Export Agents: Frazar & Hansen, 301 Clay St., San Francisco, California

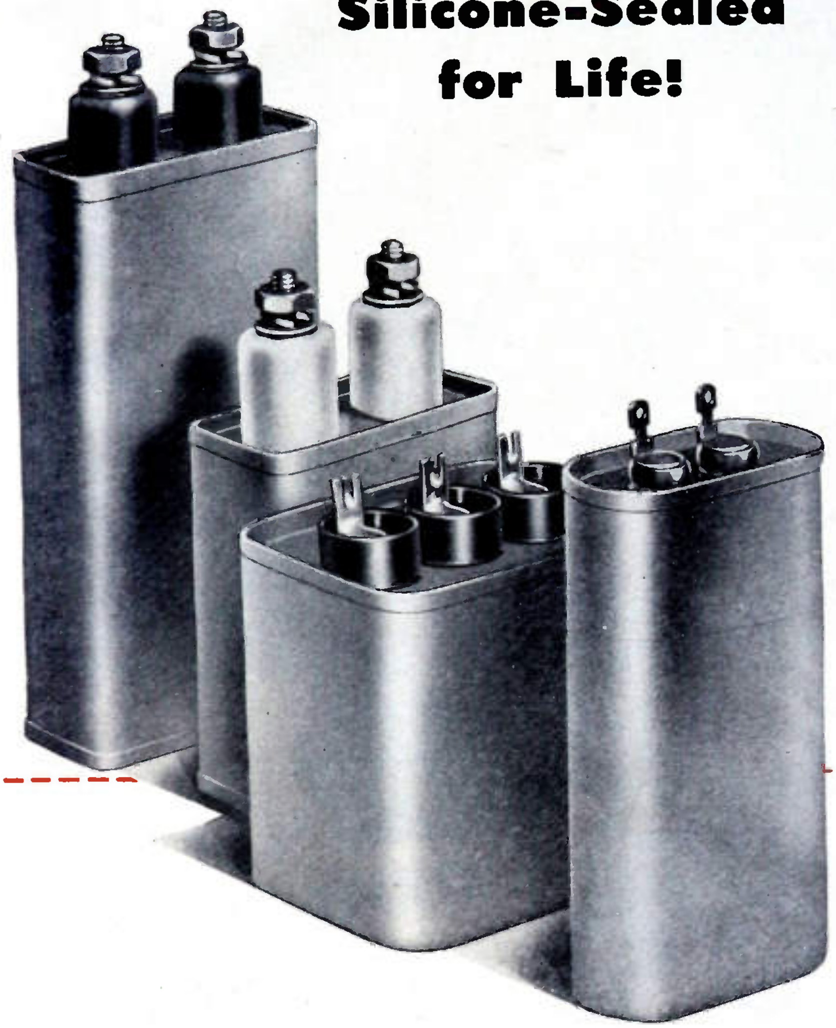
CAPACITORS

Silicone-Sealed for Life!

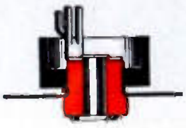
Silicone—the amazing new synthetic—made headlines when General Electric brought it out during the war. It's news again today—for G.E. has now made Silicone bushings and gaskets a *standard feature* of all its specialty capacitors up through 5000 volts.

This means that your new G-E capacitor is sealed positively, permanently—for maximum life. For Silicone seals by compression alone, without the use of contaminating adhesives. It will never shrink, loosen or pull away—it remains elastic at any operating temperature a capacitor will ever meet. Moreover, it is impervious to oils, alkalies and acids, and its dielectric strength is permanently high.

This exclusive G-E feature—with the use of highest grade materials, with strictest quality control and individual testing—make General Electric capacitors finer and more dependable than ever before. *Apparatus Dept., General Electric Company, Schenectady 5, N. Y.*



Silicone bushings used with capacitors 660-v a-c, or 1500-v d-c and lower.



Silicone bushings and plastic cups used with capacitors 660-v a-c, or 1500-v d-c and lower.

Silicone gaskets and plastic stand-offs used with capacitors rated 2000-v d-c and lower.

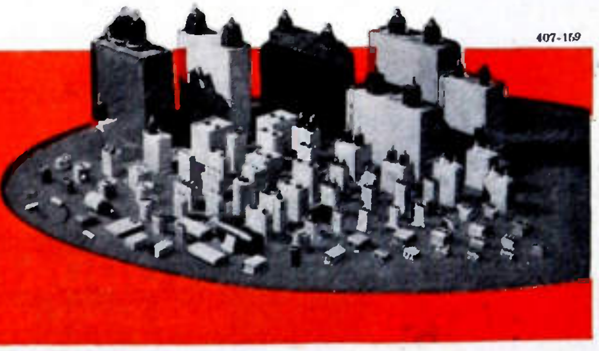
Silicone gaskets and porcelain stand-offs used with capacitors rated 2500-v to 5000-v d-c.

GENERAL ELECTRIC

Specialty Capacitors **FOR**

- Motors
- Luminous-tube transformers
- Fluorescent lamp ballasts
- Industrial control
- Radio filters
- Radar
- Electronic equipment
- Communication systems
- Capacitor discharge welding
- Flash photography
- Stroboscopic equipment
- Television
- Dust precipitators
- Radio interference suppression
- Impulse generators

AND MANY OTHER APPLICATIONS



407-169

PRECISION ATTENUATION to 3000 mc!

*Patents applied for

- VSWR less than 1.2 at all frequencies to 3000 mc.
- **Turret Attenuator*** featuring **"Pull — Turn — Push"** action with 0, 10, 20, 30, 40, 50 DB steps.
- Accuracy ± .5 DB, no correction charts necessary.
- 50 ohm coaxial circuit. Type N connectors.

Inquiries are invited concerning single pads and turrets having other characteristics

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Hillside 9294

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Trimmers • Choke Coils
Wire Wound Resistors

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- ✓ DEPENDABILITY
- ✓ MINIATURIZATION

● There is *no* variation in quality or high performance characteristics among the million of **Hi-Q** Components manufactured every month. Strict production control, engineering watchfulness and individual testing of every single unit guarantee that each of them maintains the uniform precision standards for which **Hi-Q** has long been noted. This never failing dependability is just one of many reasons why you will find **Hi-Q** Components the best that you can use.

The new **HI-Q** *Datalog is now ready. You are invited to write for a copy.*

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HI-Q

Electrical Reactance Corp.
FRANKLINVILLE, N. Y.

SALES OFFICES: New York, Philadelphia
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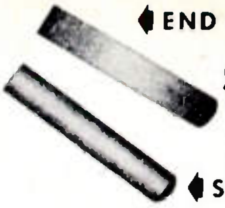
PLANTS: Franklinville, N. Y., Olean, N. Y.
Jessup, Pa., Myrtle Beach, S. C.

STANDARD CORES

. . . in a wide range of sizes, shapes and frequencies

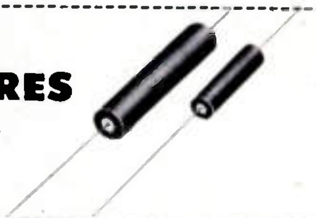


SIDE-MOLDED CORES

Outstandingly superior for permeability tuning

CHOKE COIL CORES

Insulated or non-insulated types



MOLDED IRON TRANSFORMER CORES

The ideal core for filter cores in carrier-frequency equipment

IRON SLEEVE CORES

Smaller cores of *any* standard material provide higher Q.

THREADED CORES

Permit higher Q, smaller assemblies, simplified design and AM or FM tuning

CUP CORES

Space savers de luxe. Dozens of shapes and sizes

Maximum Permeability...

...unaffected by operating conditions

STACKPOLE IRON CORES

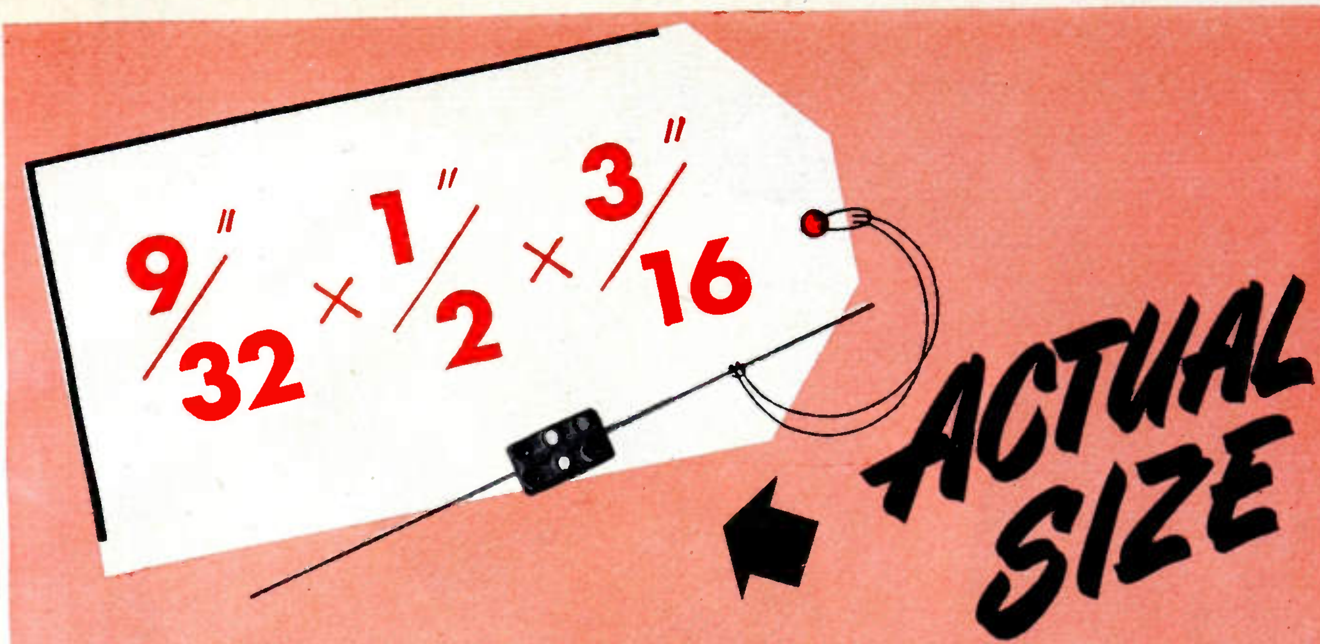
..and now Ceramag®

HIGH PERMEABILITY CERAMIC CORES FOR TELEVISION

Stackpole Ceramag TV flyback transformer cores are half the size of conventional types—assure permeability on the order of 10 to 1 by comparison. Width control types give ratios of from 1 to 8 or more compared with 1-5 for previous high permeability types assuring more positive width control in low voltage areas.

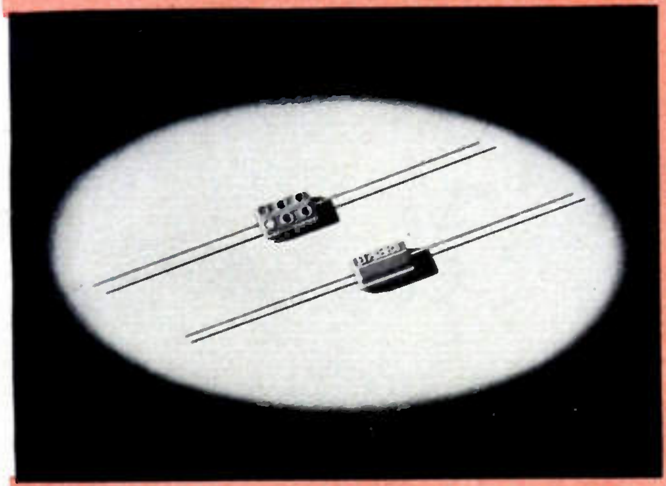
Electronic Components Division

STACKPOLE CARBON COMPANY • ST. MARYS, PENNA.



EL-MENCO CAPACITOR - CM-15

Miniature



CM 15 MINIATURE CAPACITOR

Actual Size 9/32" x ½" x 3/16"
For Television, Radio and other Electronic Applications
2 — 420 mmf. cap. at 500v DCw
2 — 525 mmf. cap. at 300v DCw
Temp. Co-efficient ±50 parts per million per degree C for most capacity values.
6-dot color coded.

This small-sized, high-capacity fixed mica condenser meets and beats strict Army-Navy standards. Like all El-Menco capacitors, the CM-15 must pass severe tests before leaving the factory. It is tested for dielectric strength at *double* working voltage; for insulation resistance and capacity value. You can always depend on the tiny CM-15 to give positive product performance under the most critical climate and operating conditions.

MANUFACTURERS WHO MAINTAIN REPUTATIONS for high-quality electrical equipment, demand and get high-quality El-Menco capacitors.

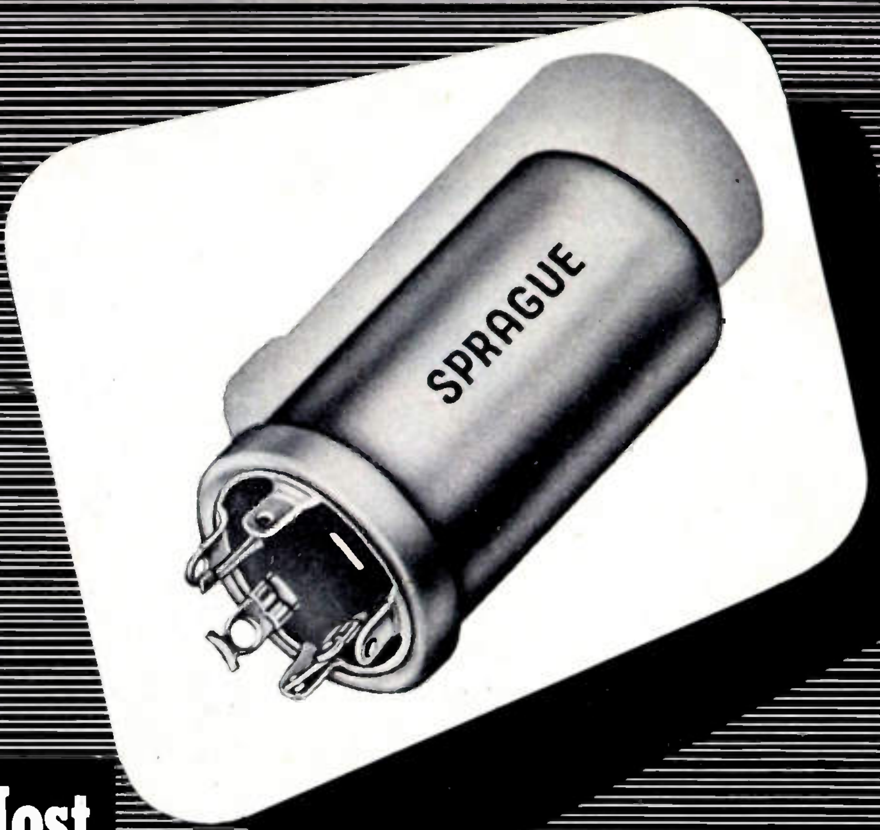
THE
ELECTRO MOTIVE MFG. CO., Inc.
WILLIMANTIC CONNECTICUT

Write on your firm letterhead for Catalog and Samples

MOLDED MICA El-Menco MICA TRIMMER
CAPACITORS

FOREIGN RADIO AND ELECTRONIC MANUFACTURERS COMMUNICATE DIRECT WITH OUR EXPORT DEPT. AT WILLIMANTIC, CONN. FOR INFORMATION.

ARCO ELECTRONICS, INC. 135 Liberty St., New York, N. Y.- Sole Agent for Jobbers and Distributors in U.S. and Canada


SPRAGUE

New Dust-Tight Plug-in Enclosure for CLARE TYPE "J" TWIN CONTACT RELAY

To Meet Severe Operating Conditions

Steel cover, securely held by a slotted-head screw, is easily removed for inspection.

● This new CLARE dust-tight plug-in enclosure for the small Type "J" Relay offers designers a number of unusual features for installation on industrial equipment.

Entrance of dust is prevented by the steel cover and by use of a Neoprene gasket which is closely fitted at the factory to the relay terminals. The dust-tight cover is easily removed for inspection. Use of standard radio plug simplifies installation and cuts wiring costs. Base is secured to chassis to prevent plug from being jarred or accidentally pulled from its socket.

Exclusive design of the CLARE Type "J" Relay allows the twin contacts to operate independently of each other. One contact is sure to close, reducing contact failure to the practical limit. This relay combines all the best features of the conventional telephone-type relay with small size and light weight. It provides unusually high current-carrying capacity, large contact spring capacity, extreme sensitivity and high operating speed.

This new dust-tight enclosed relay is one of many outstanding CLARE contributions in the development of new and better relay components for industry. CLARE Sales Engineers are located in principal cities to consult with you on your relay problems. Call them direct or write: C. P. Clare & Co., 4719 West Sunnyside Avenue, Chicago 30, Illinois. Cable Address: CLARELAY. In Canada: Canadian Line Materials Ltd., Toronto 13.

Write for Bulletin No. 108

Neoprene gasket, closely fitted at factory to relay terminals, between base and cover, effectively occludes dust.

Plug is standard radio-type plug. Standard finishes are silver lustre lacquer for cover, cadmium for base. Retaining screws hold base securely to panel.

CLARE RELAYS

First in the Industrial Field

Specialization is

for resistors too!

Specialization—and *only* specialization—can keep manufacturers abreast of today's resistance needs. The constantly-growing multitude of resistor applications demands full-time concentration on resistance products. *IRC has concentrated*—for 25 years! Result:—The widest line of resistance products in the industry; parts designed to suit specific circuit requirements in virtually every type of application; unbiased recommendations.

LOW-WATTAGE WIRE WOUND REQUIREMENTS are met efficiently by IRC Type BW Wire Wound Resistors. Exceptional low-range stability and economy suit these small, completely insulated resistors to use in meters, analyzers, cathode bias resistors, television circuits, low-range bridge circuits, high stability attenuators, low-power ignition circuits. Check coupon for Bulletin B-5.

important

IN HIGH VOLTAGE APPLICATIONS where high resistance and power are required, Type MVX high ohmic, high voltage resistors afford exceptional stability. Construction is similar to that of Type MV, but distinctive terminal permits mounting through a hole in mounting block of insulating material without terminal interference. Long resistance path permits use of high voltage on resistor while keeping voltage per unit length of path comparatively low. Check coupon for Catalog G-2.

IN CRITICAL HIGH-FREQUENCY CIRCUITS, Type MP High Frequency Resistors offer dependable performance and unusual stability. Special resistance film on a steatite ceramic form provides a stable resistor with low inherent inductance and capacity—entirely suitable for broad band RF amplifiers, RF probes, dummy loads for transmitters, television sideband filters, radar pulse equipment, and other circuits involving steep wave fronts. Send coupon for Bulletin F-1.

NEW TYPE Q CONTROL offers many advantages to engineers and purchasing agents. Its modern 15/16'' diameter size features a one-piece dual contactor of thin, high-stress alloy; simplified single-unit collector ring; molded voltage baffles; and special brass element terminals that will not loosen or become noisy when bent or soldered. Increased arc of rotation provides same resistance ratio as larger IRC controls. Salt-spray materials are employed. Complete mechanization in manufacture assures absolute uniformity and provides a dependable source of supply for small control requirements. Coupon brings you full details in Catalog A-4.

When you have special need of maintenance or experimental quantities of standard resistors in a hurry, simply phone your local IRC Distributor. IRC's Industrial Service Plan keeps him fully stocked with the most popular types and ranges—enables him to give you 'round-the-corner delivery of small order requirements. We'll gladly send you his name and address.

IRC

Wherever the Circuit Says -ᴧᴧᴧ-

Power Resistors • Voltmeter Multipliers • Insulated Composition Resistors • Low Wattage Wire Wounds • Controls • Rheostats • Voltage Dividers • Precisions • Deposited Carbon Precistors • HF and High Voltage Resistors • Insulated Chokes.

INTERNATIONAL RESISTANCE COMPANY

401 N. Broad Street, Philadelphia 8, Pa.

In Canada: International Resistance Co., Ltd., Toronto, Licensee

INTERNATIONAL RESISTANCE CO.
405 N. BROAD ST., PHILADELPHIA 8, PA.

A

Please send me complete information on the items checked below:

☐ New "Q" Controls	☐ MP High Frequency Resistors
☐ BW Insulated Wire Wounds	☐ MVX High Voltage Resistors

☐ Name of Local IRC Distributor

NAME ....

TITLE ....

COMPANY ....

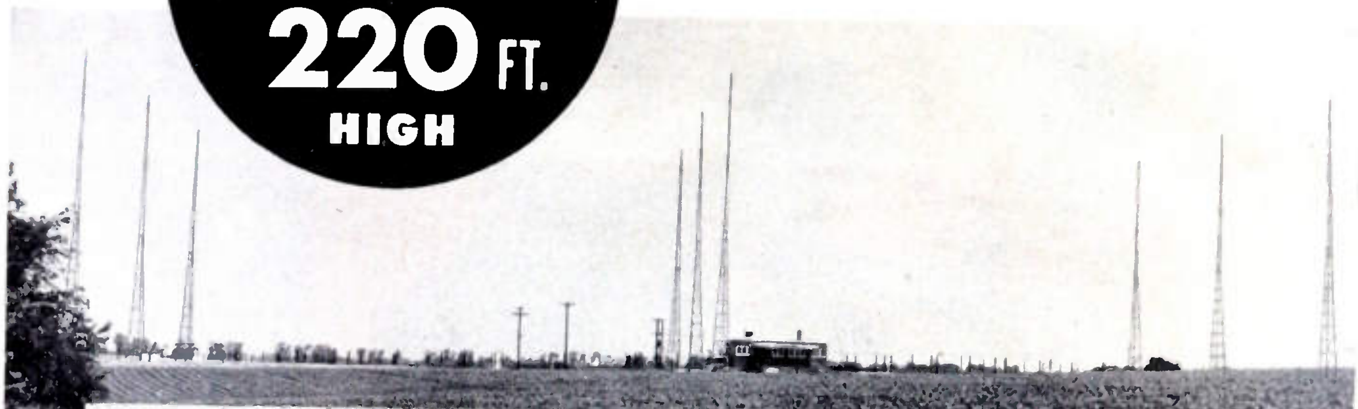
ADDRESS ....

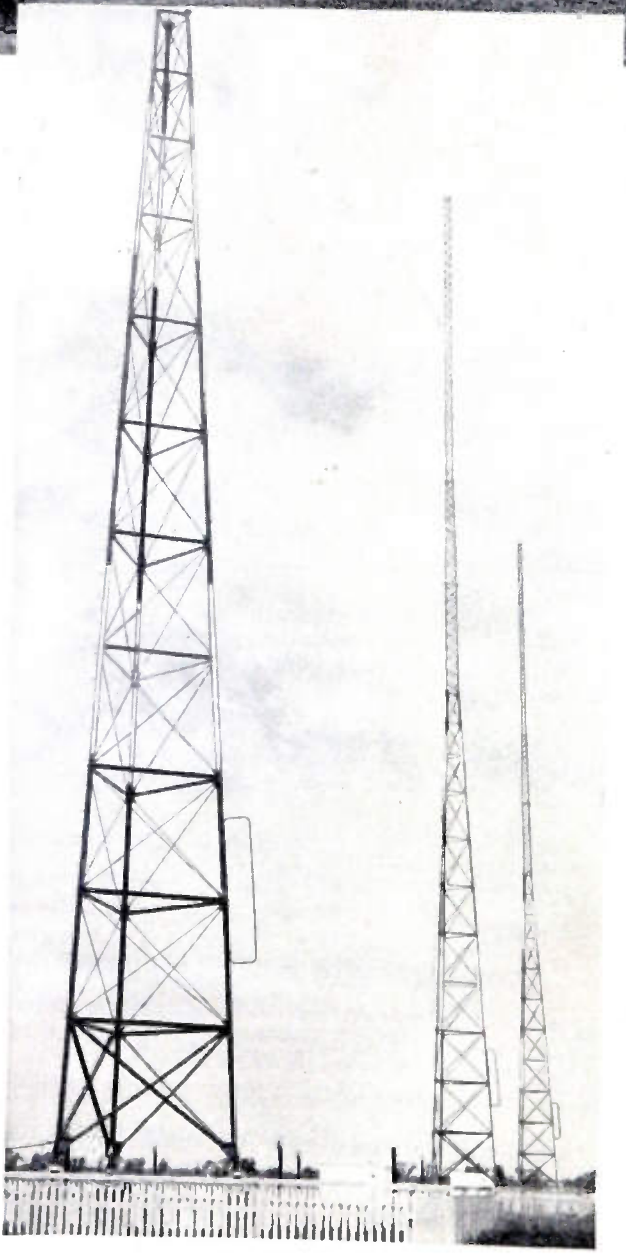
....

J. F. ARNOT & CO., ADV. AGENCY

Nine More
TRUSCON
TOWERS OF STRENGTH
220 FT.
HIGH

WDGY has Nine Truscon Radio Towers in the Milling Capital of America...





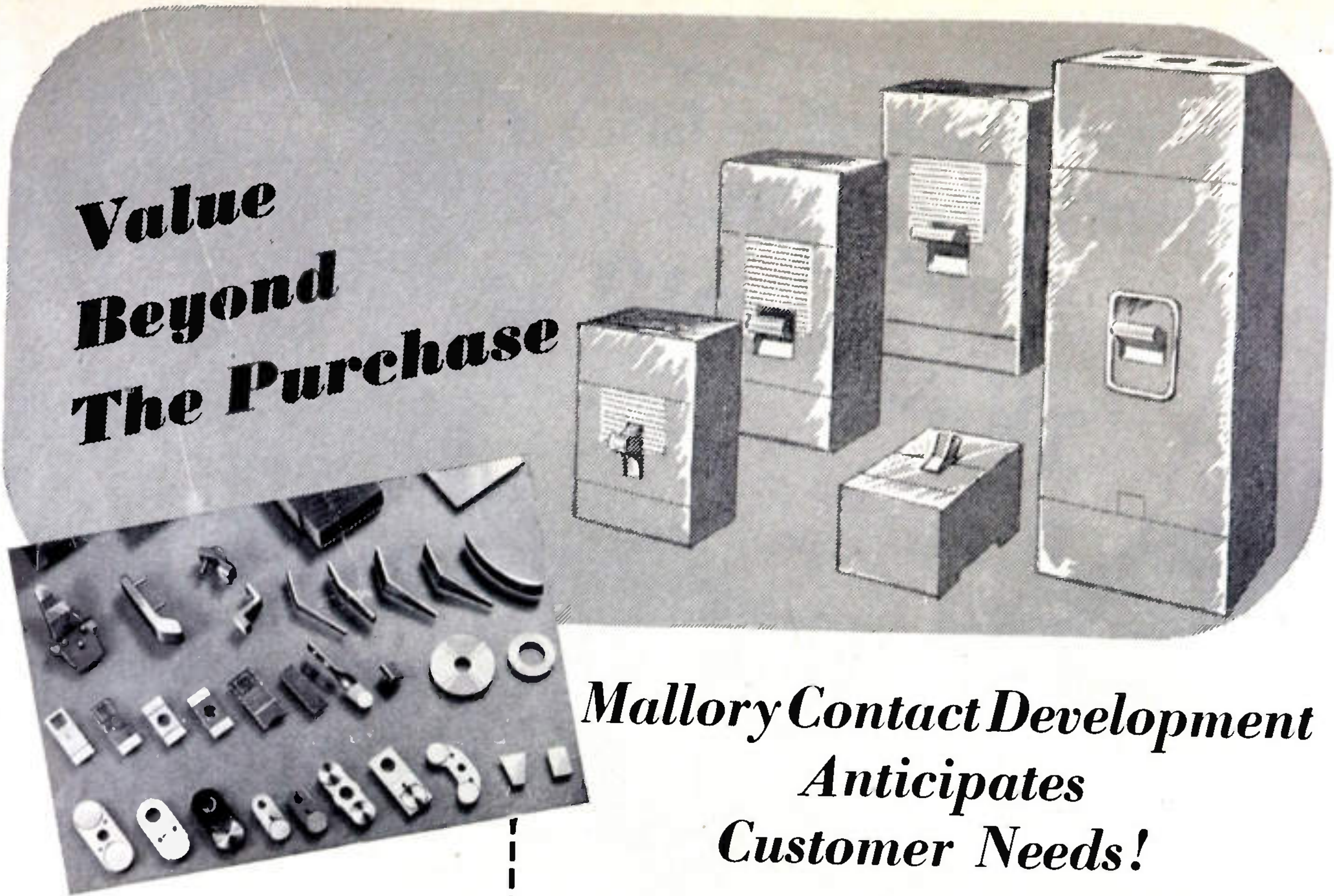
WDGY, Minneapolis, Minnesota now represents a powerful new selling force in the great northwest. It has 50,000 watts power on 1130 kilocycles, reaching 55% of Minnesota radio homes within its daytime 0.5 Mv/m. contour. It carries an effective power signal into 96 counties in three states, representing nearly a million radio homes.

The nine WDGY self-supporting Truscon Radio Towers typify Truscon's world-wide experience in designing towers to fit individual needs. Whether your own plans call for new or enlarged AM, FM, or TV transmission, Truscon will assume all responsibility for tower design and erection . . . tall or small . . . guyed or self-supporting . . . tapered or uniform in cross-section. Your phone call or letter to our home office in Youngstown, Ohio—or to any convenient Truscon District Service Office—will rate immediate, interested attention . . . and action. There is no obligation on your part, of course.

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

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


Tops For
TV and RADIO



Mallory Contact Development Anticipates Customer Needs!

Elkonite* Contacts

Elkonite is the trade name for a series of contact materials developed by Mallory and manufactured from metal powders. They are best known for their hardness, resistance to mechanical wear and impact, resistance to erosion by arcing, and resistance to sticking.

Elkonite contacts are made by the only method which permits the combining of the desirable features of basic metals which cannot be alloyed. By this means, the high melting points of tungsten, molybdenum, or their carbides, can be combined with the current-carrying ability of silver and copper.

Customers' contact problems are solved rapidly and effectively due to Mallory's precise attention to every detail of design, material and production.

A manufacturer of small industrial circuit breakers recently asked Mallory to study his contact assembly . . . with an eye to reducing costs. Investigation proved that new Mallory equipment coupled with unique Mallory production techniques would eliminate certain expensive finishing operations. As a result, the problem was solved in rapid-fire order . . . and the contacts delivered at a price that is 21% less than the customer previously had been paying.

That's value beyond the purchase!

Mallory contact know-how is at your disposal. What Mallory has done for others can be done for you!

In Canada, made and sold by Johnson Matthey & Mallory, Ltd., 110 Industry St., Toronto 15, Ontario

**Reg. U. S. Pat. Off.*

Electrical Contacts and Contact Assemblies

P. R. MALLORY & CO., Inc.

MALLORY

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

SERVING INDUSTRY WITH

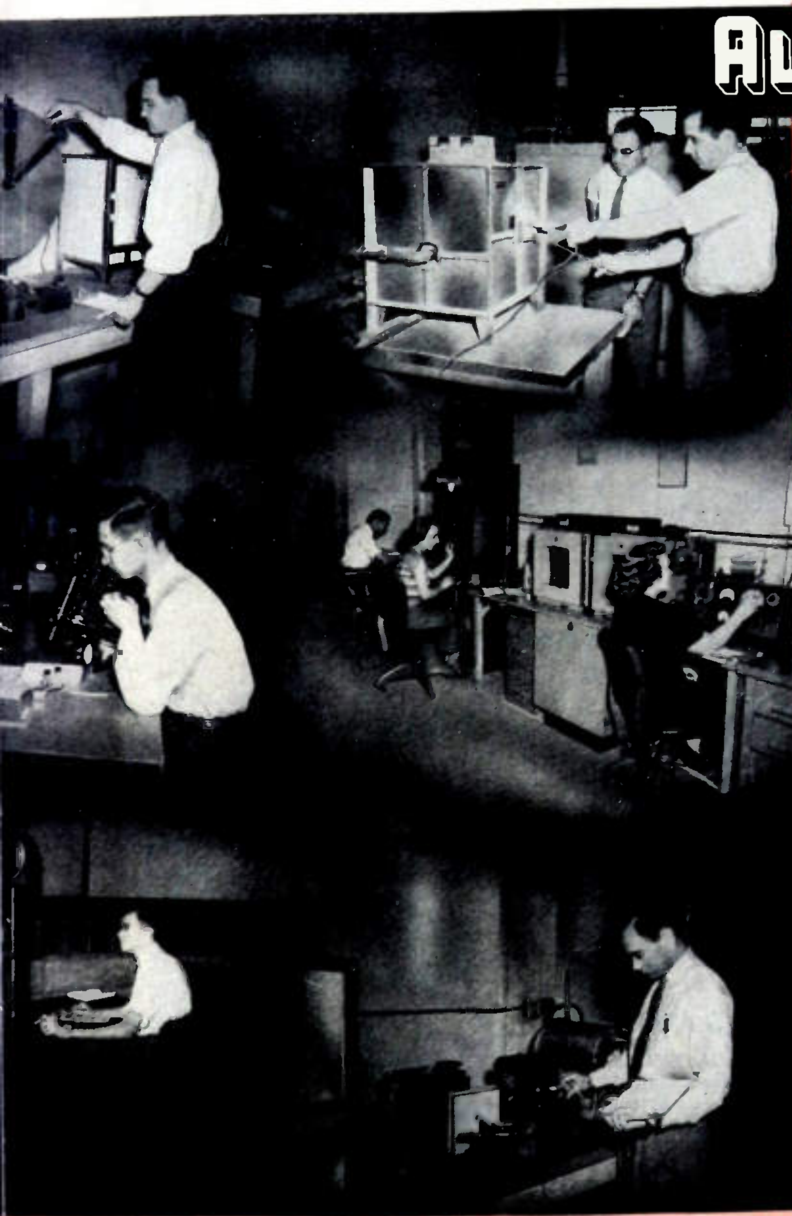
Capacitors	Contacts
Controls	Resistors
Rectifiers	Vibrators
Special Switches	Power Supplies

Resistance Welding Materials

research

at American Lava Corporation is continuous,
to anticipate and answer your problems
on Custom Made Technical Ceramics

ALSIMAG
Reg. U. S. Pat. Off.



Here are AlSiMag custom made components for radar, radio, television, electric appliance, textile, chemical, metallurgical, gas, petroleum, rubber, carbon and foundry applications.

Here you are most apt to find the answer to any question involving technical ceramics. The Research Division has developed hundreds of highly successful compositions which combine special physical characteristics needed for unusual requirements. These compositions are custom fabricated to your specifications.

AlSiMag's Research Division maintains accurate, cross-indexed records of all research findings to give you a prompt answer on the possibilities of furnishing you any special combinations of physical characteristics you may find desirable.

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AMERICAN LAVA CORPORATION

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OFFICES: METROPOLITAN AREA: 671 Broad St., Newark, N. J., Mitchell 2-8159 • CHICAGO, 9 South Clinton St., Central 6-1721
PHILADELPHIA, 1649 North Broad St., Stevenson 4-2823 • LOS ANGELES, 232 South Hill St., Mutual 9076
NEW ENGLAND, 38-B Brattle St., Cambridge, Mass., Kirkland 7-4498 • ST. LOUIS, 1123 Washington Ave., Garfield 4959

WILCOX ...FIRST CHOICE of HAWAIIAN AIRLINES

VHF AIR-BORNE COMMUNICATIONS

Hawaiian Airlines selected the WILCOX TYPE 361A COMMUNICATIONS SYSTEM for all aircraft. This consists of a 50 watt transmitter, a high sensitivity receiver, and a compact power supply, each contained in a separate ½ ATR chassis. Transmitter and receiver contain frequency selector with provisions for 70 channels . . . ample for both present and future needs.

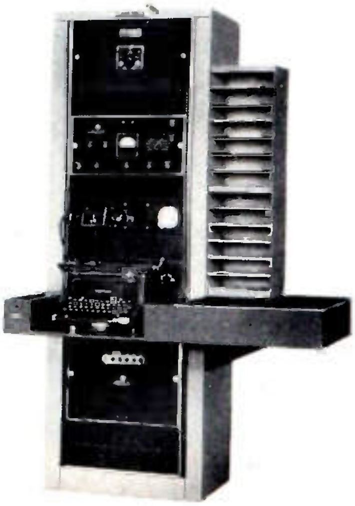
VHF GROUND STATION PACKAGED RADIO

Hawaiian Airlines selected the WILCOX TYPE 428A FACTORY PACKAGED STATION for all ground stations. This consists of the WILCOX 406A fixed frequency 50 watt transmitter, the WILCOX 305A fixed frequency receiver, the WILCOX 407A power supply, the WILCOX 614A VHF antenna, telephone handset, loudspeaker, desk front, typewriter well, and message rack.

DEPENDABLE COMMUNICATIONS FOR THE WORLD'S AIRLINES

During recent months, many of the world's foremost airlines, UNITED, EASTERN, TWA, MID-CONTINENT, BRANIFF, PIONEER, ROBINSON, and WISCONSIN CENTRAL have placed volume orders for similar communications equipment. No greater compliment could be paid to the performance, dependability, and economy of WILCOX equipment than to be "**FIRST CHOICE**" of this distinguished group.

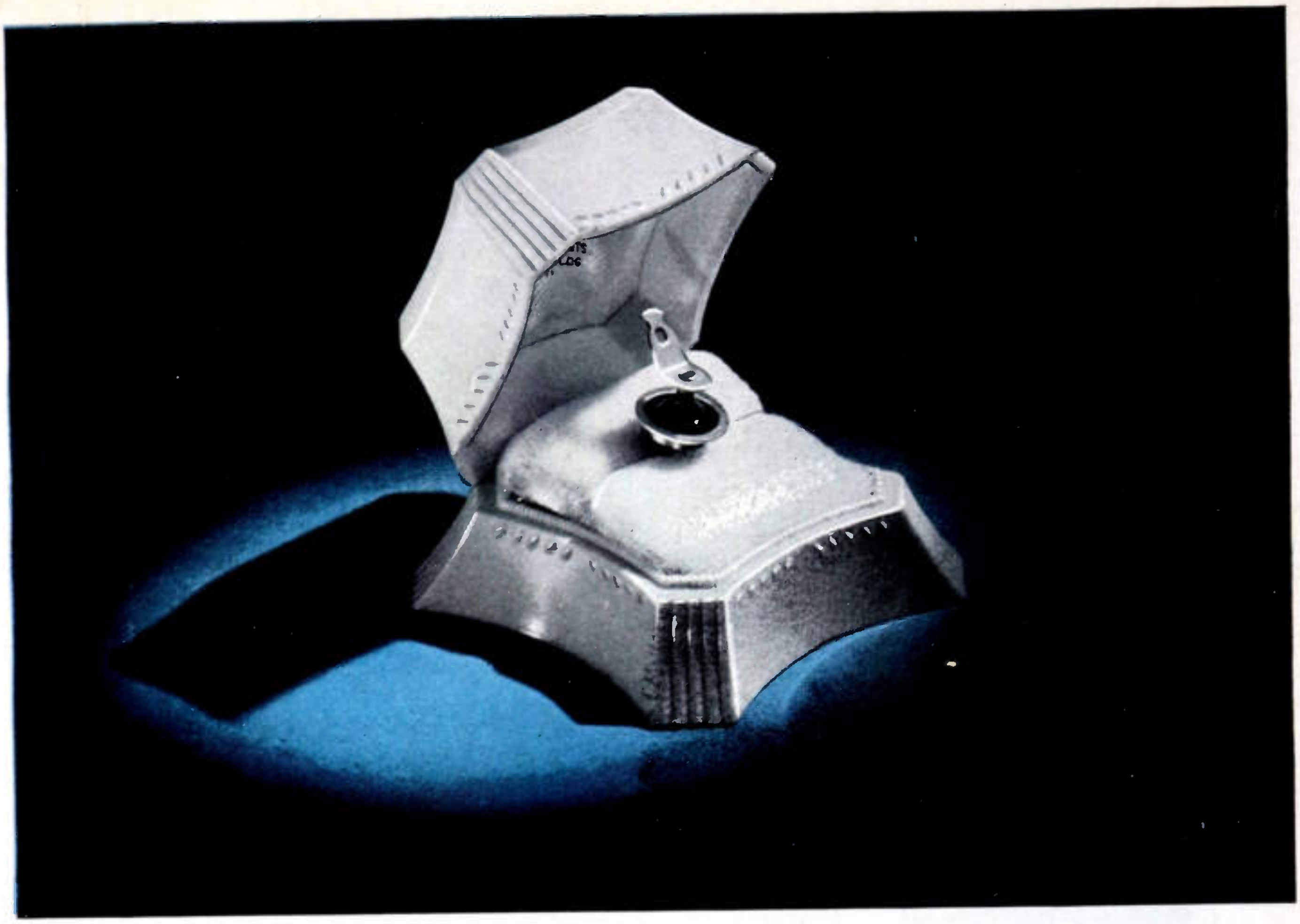
Write Today for complete information on the Type 361A VHF Air-borne Communications System and the Type 428 Packaged VHF Ground Station.



Type 428 Packaged VHF Station

WILCOX ELECTRIC COMPANY

KANSAS CITY 1, MISSOURI, U.S.A.



Why a Fusite Terminal Where a Diamond Ought To Be?

A Fusite Terminal would look much more natural performing its vital function in the hermetic sealing of your electrical product. But since it's every bit as valuable for 1000 other products that should be fusion sealed, we aren't playing favorites.

The smooth uniform interfusion of steel and inorganic glass that is a Fusite Terminal is as beautiful as a flawless diamond to any design engineer. In its own way, it's as rugged as the diamond used on the tip of a heavy duty drill.

It withstands the thermal shock of tortuous heat from soldering or welding and the rapid cooling that follows. It will carry up to 3000 A.C. volts (RMS) with a 10,000 megohms insulation factor after salt water immersion.

This is just one of a wide line of standard Fusite single and multiple electrode terminals.

Would you like to know more and see samples? Write to Dept. P

TERMINAL ILLUSTRATED 112 HTL
SINGLE—HOLLOW TUBE ELECTRODE WITH LUG

INSULATE ELECTRICALLY
GLASS TO METAL
FUSITE
TERMINALS
PROTECT PRODUCT
PERFORMANCE
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THE FUSITE CORPORATION
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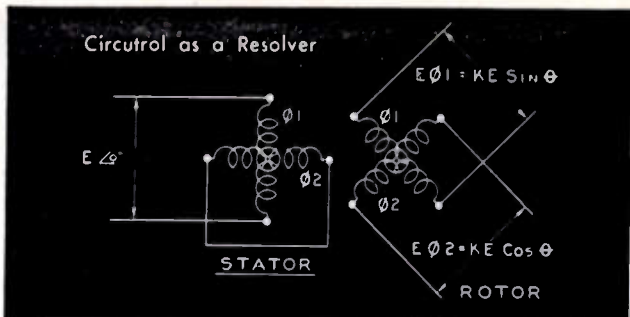
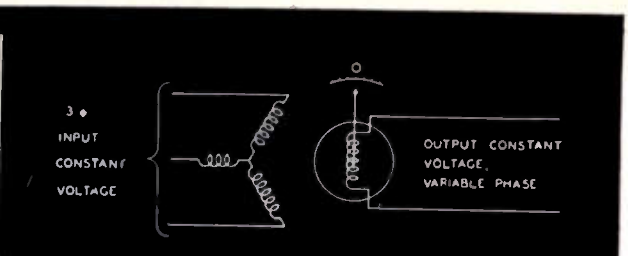
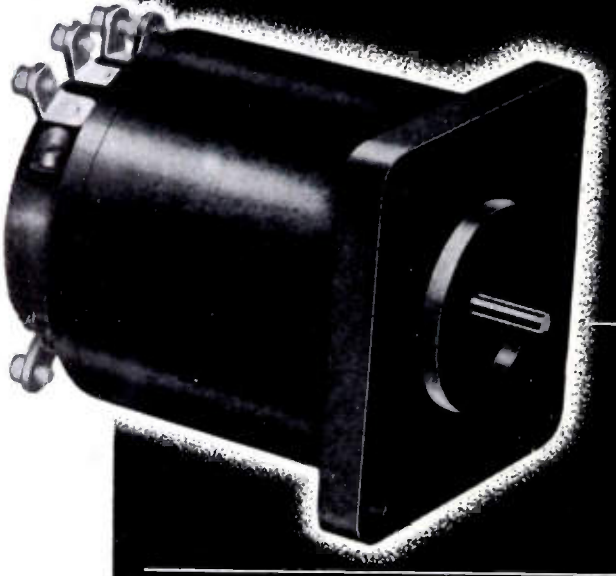
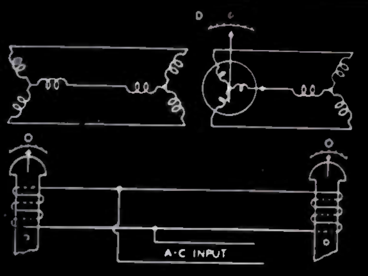


Circutrol as a Resolver

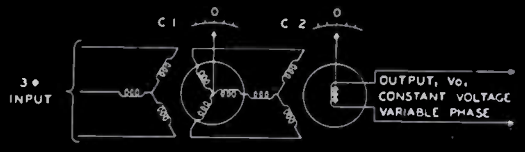


Circutrol as a Phase Shifter

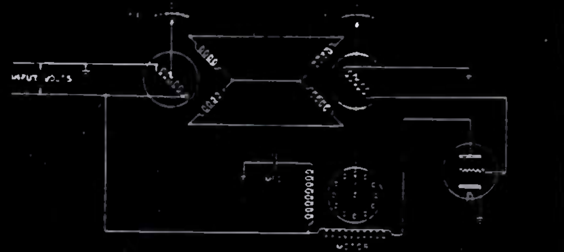




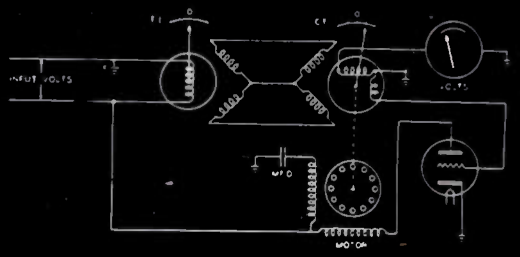
Circutrol as a Differential Unit



Double Phase-Shifter Circuit



Control-Synchro System

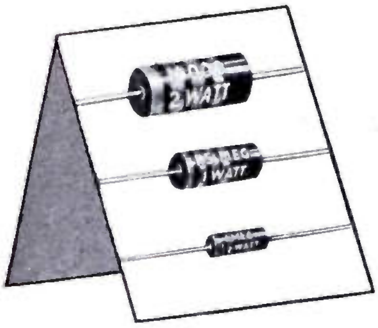


Resolver-Control System


17" RECTANGULAR

OHMITE *presents...*

two dependable performers*

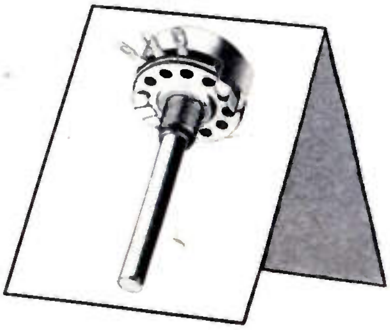


"Little Devil" Composition Resistors

For quick, easy identification, resistance and wattage are clearly marked on every one of these tiny, rugged insulated composition resistors. In three sizes—½, 1, and 2-watt, and all RMA values. Tolerances ±5 and ±10%.

Here are two Ohmite stars that can always be depended upon to give a good performance—anywhere, anytime. Made to the high standards that characterize all Ohmite products, they are built to stand up under the most severe test and service conditions. In laboratory and development components, as well as maintenance parts, you *need* that dependable preformance. Follow the lead of thousands of engineers and designers—"Be Right with Ohmite!"

*So that these two exceptionally high-quality products will be universally obtainable, Ohmite Manufacturing Company, in co-operation with the Allen-Bradley Company, has arranged for the Type AB (Allen-Bradley Type J) control, and Little Devil Molded Composition Resistors (Allen-Bradley Types EB, GB, and HB) to be available from stock at Ohmite distributors.



Type AB Potentiometer

It's quiet! This composition potentiometer has a resistance unit that's solid-molded. As a result, the noise level often becomes less with use. Has 2-watt rating, good safety factor.

Available at all OHMITE Distributors
WRITE FOR CATALOG 40 ON COMPANY LETTERHEAD

OHMITE MANUFACTURING CO.
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Be Right with **OHMITE**

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RHEOSTATS • RESISTORS • TAP SWITCHES

SHERRON WILL INITIATE OR DEVELOP AND MANUFACTURE PROTOTYPES FROM YOUR SPECIFICATIONS

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- Transmitter — Receiver: A.M.; F.M.; P.M.; TV
- Navigational Aids and devices — Homing and Direction Finding Equipment
- Telemetering; Radar; Sonar
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ELECTRONIC CONTROL EQUIPMENT FOR:

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- Computers and Calculators • Servo Links
- Velocity Propagation Measurement
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VACUUM TUBE CIRCUIT DEVELOPMENT

- New applications for existing and newly developed vacuum tubes
- Precision Test and Processing Equipment for all types of vacuum tubes, laboratory, production or composite

ELECTRONIC MEASURING DEVICES

- Flow indicators:—Liquids, gases, solid components
- Sorting, counting and inspection
- Chemical process control; titration; ionization; diffusion
- Flow detection — strains, stresses, inner faults
- X-Ray, supersonic and radio frequency applications to measurement

INSTRUMENTATION

- DC; AC; Audio; RF; Microwaves; Infra-Red; Visible Spectrum; Ultra-Violet; Soft and Hard X-Ray; Cosmic Radiation
- Vacuum Tube Meters • Null Detectors
- RF; Audio frequency generators • Bridges
- Multi-Wave Shape Generators
- High Gain Amplifiers • Oscilloscopes
- Power Supplies, Regulated, High & Low Level

TELEVISION

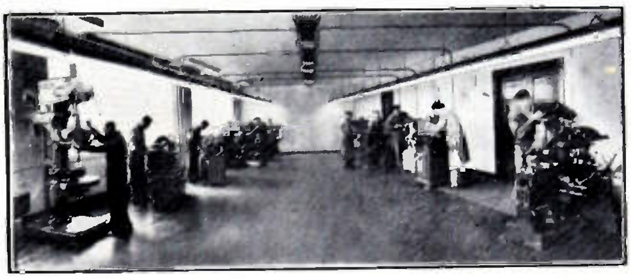
- Television Signal Synthesizer
- Shapers; Timers; Delay Circuits
- Air Monitors • Sync Generators
- Field Intensity Equipment

NUCLEONICS

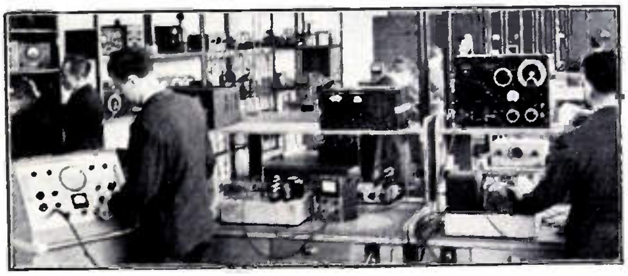
- Counters — Geiger-Muller, Scintillation, and crystal types
- Computers — Mechanical linkages, complex electrical analogues digital computers
- Servo-Mechanisms — Velocity and position control to specified practical accuracy limits
- Amplifiers — SHF; UHF; VHF; LF; D.C. to fit any application
- Oscillators — All frequencies, audio to the extremely high microwaves
- Power Supplies — High & Low power and voltage, D.C. output hum-free to your limits
- Regulators — Electronic or electro-mechanical, Regulation, drift, range of control, to your specifications
- Measurement—Devices for measurement and control of all parameters capable of being controlled and capable of producing proportional electrical, optical or other physical indication
- Particle Accelerator Controls — Grouping of controls, supplementary apparatus, and experimental system into a compact integrated unit

SHERRON OFFERS AN OVER-ALL, *START-TO-FINISH* ELECTRONICS SERVICE

RESEARCH · DESIGN · DEVELOPMENT



You get all the benefits of an experienced, integrated, all inclusive electronics service — when you work with Sherron. This definitive service includes . . . research and development in our electronics and electro-mechanical laboratories right through sheet metal fabrication in our block-long plant.



At the top of the Sherron personnel pyramid are physicists with electronic knowledge who will provide the necessary research for your project . . . electronic and mechanical engineers who will develop it...technicians who will accredit its workability. These men work hand in hand with our production engineers, who have at their command the necessary sheet metal fabrication facilities and manpower to turn out a finished prototype . . . ready for you to manufacture. Whether we initiate the design — or work from your prints and specifications — our service is confidential.

ELECTRO MECHANICAL LAB.

To design, style, develop and manufacture prototypes and high precision mechanical sub-assemblies, or assemblies involving part manufacture and assembly of units incorporating machine elements of bases, springs, levers, shafts, bearings, cams, valves, regulators, drives, transmissions and controls for:

- ELECTRO MECHANICAL TEST EQUIPMENT
- SPECIAL HYDRAULIC UNITS
- SPECIAL MEASURING INSTRUMENTS
- EQUIPMENT FOR NUCLEAR PHYSICS
- SERVO MECHANISMS
- KEYERS & COUPLERS
- DRIVING AND CONTROLLING EQUIPMENT
- REGULATORS
- MICRO WAVE APPARATUS
- MECHANICAL ELEMENTS FOR SPECIAL ATOMIC ENERGY EQUIPMENT
- RADAR MECHANISMS
- COMPUTER MECHANISMS
- MECHANICAL OPTICAL SPECIALTIES
- MECHANISMS OF NAVIGATIONAL DEVICES
- VIBRATION CONTROL MOUNTINGS
- AUTOMATIC SCREW MACHINE PARTS
- PRECISION GEAR TRAINS
- MECHANICAL PORTIONS OF MEDICAL AND OPERATING EQUIPMENT
- MECHANICAL PORTIONS OF ULTRA HIGH FREQUENCY EQUIPMENT

Sherron Electronics

SHERRON ELECTRONICS COMPANY

Division of Sherron Metallic Corporation

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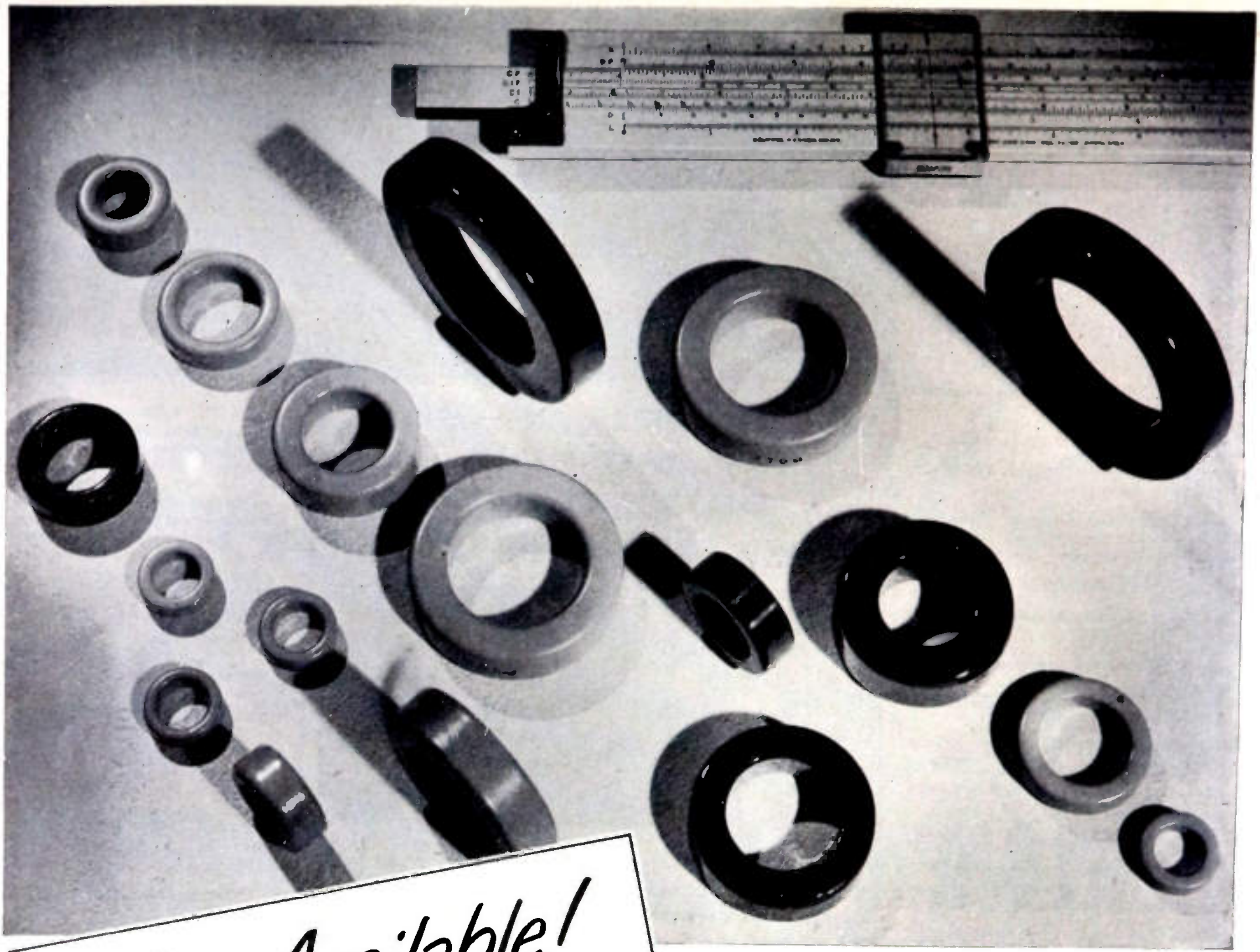

Now Available!

MOLYBDENUM PERMALLOY POWDER CORES*

HIGH Q TOROIDS for use in Loading Coils, Filters, Broadband Carrier Systems and Networks—for frequencies up to 200 K C

COMPLETE LINE OF CORES TO MEET YOUR NEEDS

- ★ Furnished in four standard permeabilities—125, 60, 26 and 14.
- ★ Available in a wide range of sizes to obtain nominal inductances as high as 281 mh/1000 turns.
- ★ These toroidal cores are given various types of enamel and varnish finishes, some of which permit winding with heavy Formex insulated wire without supplementary insulation over the core.

For high Q in a small volume, characterized by low eddy current and hysteresis losses, ARNOLD Moly Permalloy Powder Toroidal Cores are commercially available to meet high standards of physical and electrical requirements. They provide constant permeability over a wide range of flux density. The 125 Mu cores are recommended for use up to 15 kc, 60 Mu at 10 to 50 kc, 26 Mu at 30 to 75 kc, and 14 Mu at 50 to 200 kc. Many of these cores may be furnished stabilized to provide constant permeability (±0.1%) over a specific temperature range.

* *Manufactured under licensing arrangements with Western Electric Company.*

W&D 2930

THE ARNOLD ENGINEERING COMPANY

SUBSIDIARY OF ALLEGHENY LUDLUM STEEL CORPORATION

147 EAST ONTARIO STREET, CHICAGO 11, ILLINOIS

Let **Bendix** Solve Your Problems

SPECIALIZED DYNAMOTORS

. . . Promptly and at Moderate Cost!

Bendix dynamotors are built to supply the *exact* power requirements of your equipment—to work from any input voltage and to deliver the necessary power at any output voltage. Dual or triple output voltages are available for high and low-level portions of the circuit, or for biasing. For critical circuits, regulated outputs will simplify your design problems, especially since a regulated filament supply can be obtained as a bonus when regulating the high voltage output. Bendix will build your dynamotors to the usual military specifications or to meet even more rigid requirements, such as operation at higher temperature, or altitudes in experimental equipment.

Samples or production units of special dynamotors are priced competitively. A definite proposal will be made upon receipt of the details of your problem. For immediate information call our Engineering Staff—Red Bank 6-3600, Red Bank, New Jersey.

THE RIGHT DYNAMOTOR FOR EVERY PURPOSE

- Sizes—2¾" to 5¼" diameter
- Power Range—10 to 500 watts
- Input Voltage—6 to 115 volts
- Output Voltage—6 to 1500 volts
- Single and multiple output and input
- Plain and regulated types

RED BANK DIVISION

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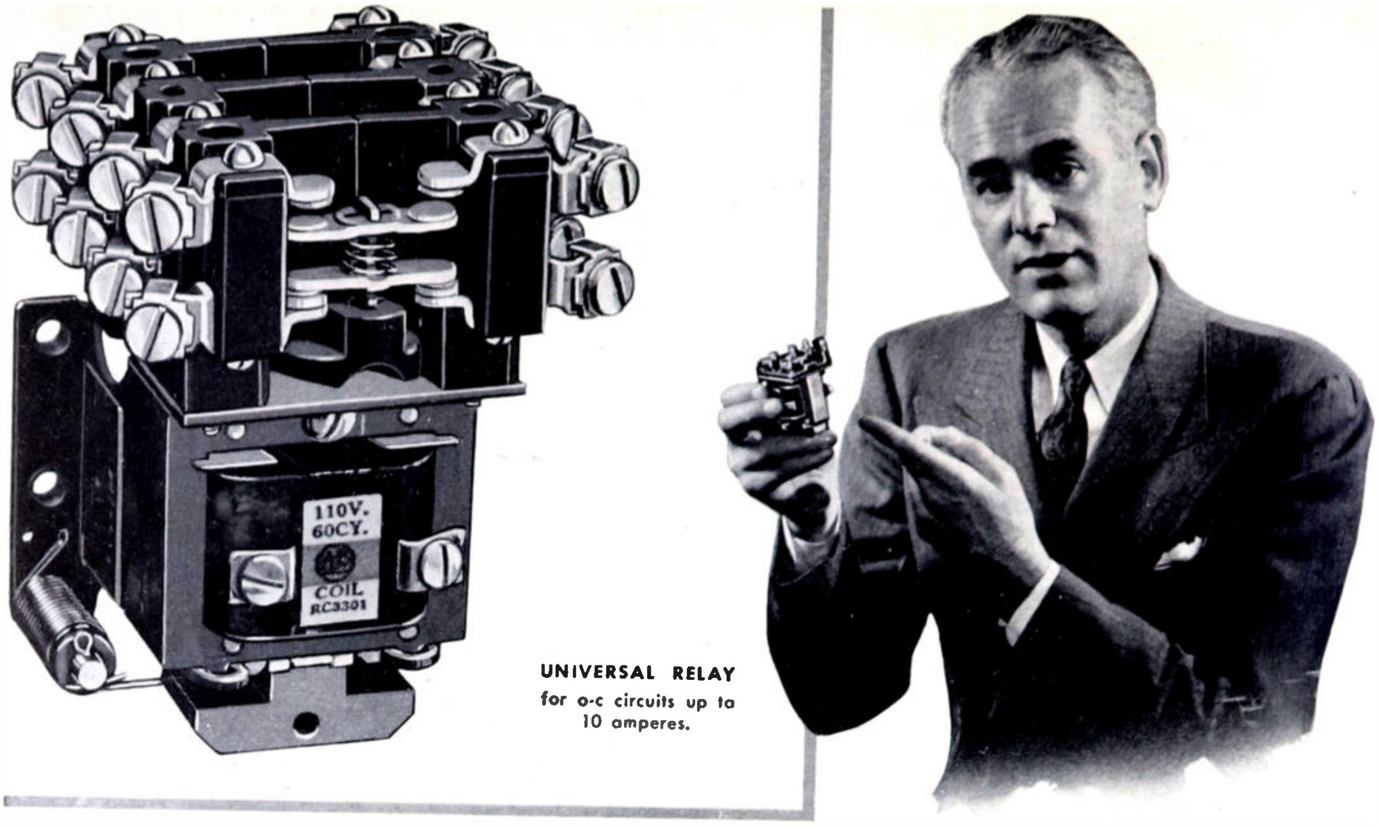


UNIVERSAL RELAY
for a-c circuits up to 10 amperes.

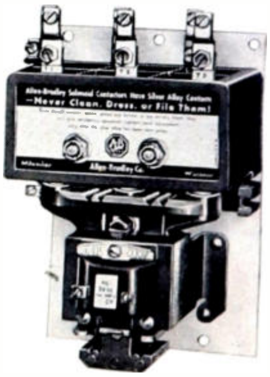
SOLENOID CONTACTORS from 10 to 900 Amperes

When power supply circuits carry substantial currents . . . or are switched frequently . . . or their functioning must be foolproof . . . the relays and contactors used in such circuits must be rugged, consistent in action . . . and trouble free.

Allen-Bradley relays and contactors are extremely compact for their ratings . . . but designed for tough service. They are built up to a high quality standard . . . not down to a price. They have but one moving part . . . there are no trouble breeding pins, pivots, levers, or flexible shunts. The double break, silver alloy contacts are maintenance free.

For sturdy, long lived switching units, specify Allen-Bradley solenoid contactors . . . made in a full range of sizes up to 900 amperes. Send for catalog, today.

Allen-Bradley Co.
114 W. Greenfield Ave., Milwaukee 4, Wis.



100 AMPERE CONTACTOR
Allen-Bradley solenoid contactor for circuits up to 100 amperes. Double break, silver alloy contacts are totally enclosed. Simple, straight line solenoid action means long, trouble free operating life.

LIMIT SWITCHES OF ALL TYPES
The Allen-Bradley line of Bulletin 801-802 limit switches covers a remarkable assortment of pilot controls for automatic limiting of control circuits. All types of standard and precision limit switches are available with lever arms, rollers, forks, and chain controls.

A-B Quality

ALLEN-BRADLEY
RELAYS • RESISTORS

Sold exclusively to manufacturers of radio and electronic equipment

QUALITY

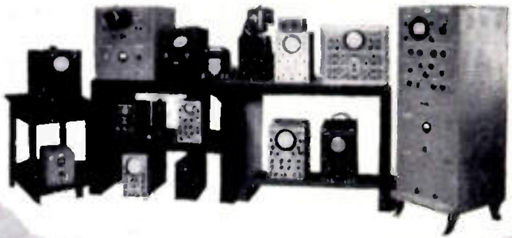
FOR CONGESTED AREAS
specify BLAW-KNOX

Whether it's for a spectacular TV antenna or a modest mobile communication system, a Blaw-Knox Tower *designed for the job* is your best—and safest—buy. In congested areas and cities throughout the country hundreds of Blaw-Knox Towers—both large and small—are proving the worth of their superior construction. For this assurance of safety and performance you might expect a premium price but—Blaw-Knox Towers *cost no more* than those of doubtful stability! Specify Blaw-Knox, and be sure.

BLAW-KNOX DIVISION OF BLAW-KNOX CO.
2037 Farmers Bank Bldg., Pittsburgh, Pa.

BLAW-KNOX *ANTENNA* TOWERS

DUMONT *Oscillography..*



. . . the outstanding heritage of another great performer

DEFINING THE OSCILLOGRAPHIC SPECTRUM **from 10 cps. to 15 megacycles**

THE NEW DU MONT TYPE 294 CATHODE-RAY OSCILLOGRAPH

➧ The Type 294 is an extremely versatile cathode-ray oscillograph combining high-voltage operation with precise high-frequency circuit design, extending its general-purpose utility to meet the specialized needs of high-speed transient study.

Stable operation of the high-gain, wide-band amplifier of the Y axis over the entire frequency range from 10 cps. to 15 megacycles includes the performance of a signal-delay line built into the Y-axis circuit to insure full display of short-duration pulses. An input pulse rise time of 0.01 μs. will be reproduced with a rise time not exceeding 0.03 μs.

Available undistorted deflection of both symmetrical signals and unidirectional pulses of either positive or negative polarity exceeds the usable vertical scan of the cathode-ray tube. A built-in high-voltage unit supplies 12 kv. accelerating potential to the Du Mont Type 5XP- cathode-ray tube; rear-panel selection of a lower potential may be made for increased sensitivity and deflection.

A flexible sweep circuit provides continuously variable driven and recurrent sweeps with sweep calibration being provided by internal timing markers applied through the Z-axis amplifier.

• • •

Permanent records of phenomena studied with the Type 294 may be made with either the Du Mont Type 271-A or 314-A Oscillograph-record Camera.

GENERAL SPECIFICATIONS

Cathode-ray Tube . . . . . . Du Mont Type 5XP-
Accelerating potential . . . . . . 12,000 volts
. . . . . . 7,000 volts

Y-axis Amplifier
Frequency response
10 cps. to 15 megacycles
Sensitivity . . . . . 0.15 rms volt/in. at 7 kv.
. . . . 0.20 rms volt/in. at 12 kv.
Rise time . . . . 0.03 μs. from 10% to 90%

Signal Delay . . . . 0.25 μs.

X-axis Amplifier
Frequency response . . . . . 2 cps. to 700 kc.
Sensitivity . . . . . . 0.4 rms volt/in. at 7 kv.
. . . . . 0.5 rms volt/in. at 12 kv.
Rise time . . . . . . 0.5 μs. from 10% to 90%

Driven Sweep Range . . . . . . 0.1 sec. to 2 μs.

Recurrent Sweep Range . . 10 cps. to 150 kc.

Z-axis Amplifier
Polarity selection—3 volts peak to blank trace of normal intensity.

Timing-Marker Intervals
100 μs., 10 μs., 1 μs.

Trigger Generator
Repetition rate . . . . 200 to 3600 p.p.s.
Output amplitude . . . . 50 volts peak
Output polarity . . . . positive or negative

Physical Specifications
Indicator Unit
24½" d.—15¾" h.—12¾" w.—62 lbs.
Power Supply
19¾" d.—15¾" h.—12¾" w.—100 lbs.

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

...with New

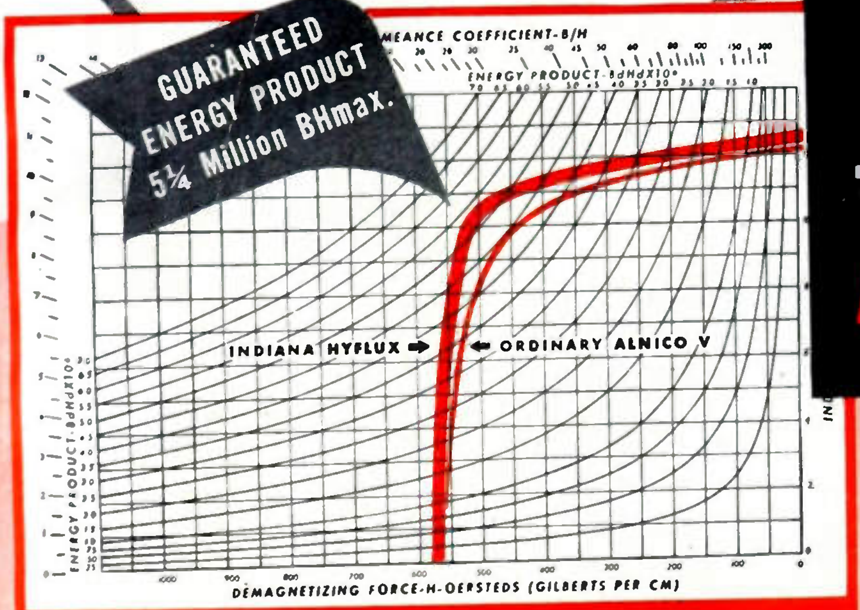
INDIANA HYFLUX

PERMANENT MAGNETS

Now 22% More Energy

for Alnico V

AT NO EXTRA COST



DEMAGNETIZATION AND ENERGY PRODUCT CURVE

The Stronger Magnets you've been hoping for ARE HERE!

Now it can be told! After years of research and months of field-testing, INDIANA announces exclusive new *super-strength permanent magnets* made of HYFLUX Alnico V.

The industry's highest published guaranteed energy product for standard Alnico V has been 4½ million BHmax. Now, INDIANA guarantees much greater strength—5¼ million BHmax, and the *average* energy product reaches 5½ million BHmax, or more. Yet, for these higher-energy HYFLUX magnets, you pay *not a penny* premium.

What is HYFLUX Alnico V? INDIANA HYFLUX is *not* a new alloy. It's the result of a new precision technique applied to *dependable* Alnico V . . . new procedures, controls, instrumentation, and equipment, and *precise supervision* over every step of production. Add to this the 42 years of permanent magnet experience and a long-term training program for personnel by the world's *largest* exclusive producer, and you have the background and reasons why INDIANA HYFLUX is so outstanding in *both* performance and value.

Find Out What HYFLUX Can Do! For *greater strength* . . . more compact designing . . . for the lower production costs these smaller, better magnets can bring to *your own* products, get all the facts today on amazing INDIANA HYFLUX. It's the *most important development* in permanent magnets since the introduction of Alnico V.

See what HYFLUX *does!*

When this standard R. M. A. No. 3 loud speaker magnet is INDIANA HYFLUX Alnico V with the minimum guaranteed energy product of 5¼ million BHmax, it has .7 decibels greater output than when made with 4½ million BHmax *regular* Alnico V. Similar improvements—in strength or size—apply to *all* applications. INDIANA HYFLUX is ready now to bring *you* these advantages.

For Cost-Cutting Engineering Aid,
Put Your Magnet Problems up to INDIANA.

THE INDIANA STEEL PRODUCTS COMPANY
Specialists in Packaged Energy Since 1908
SALES OFFICES: BOSTON, CHICAGO, CLEVELAND, LOS ANGELES, NEW YORK, PHILADELPHIA, ROCHESTER

INDIANA
PERMANENT
MAGNETS

THE INDIANA STEEL PRODUCTS COMPANY
DEPT. N-60, VALPARAISO, INDIANA

Please send me all the facts on INDIANA HYFLUX. I am interested in permanent magnets for:

. . . .

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Name . . . .

Company . . . .

Street . . . .

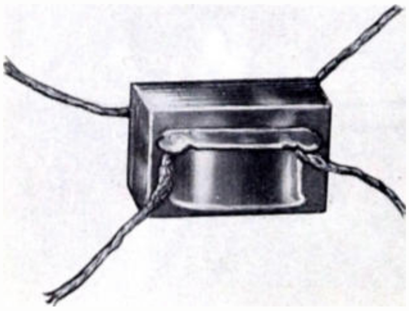
City . . . . *Zone* . . . . . *State* . . . .

NEWS and NEW PRODUCTS

JUNE, 1950

Extremely Small Transformers

United Transformer Co., 150 Varick St., New York 13, N. Y. claims to have manufactured the smallest standard audio transformer in the world. This new unit is so small that 30 will fit into a cigarette pack.



The Type SSO transformer's dimensions are only 0.4×0.75×0.56 inch. It weighs 0.28 ounce. Five stock types cover input, interstage, output, and reactor applications.

Especially suitable for hearing aid, aircraft, and all other instances where size and weight are the prime consideration, the SSO is suited for the miniaturization program in Government aircraft and for Navy emergency miniature transmitters. Designs are available for all types of low-level applications requiring wide frequency range.

Great dependability is provided in this minute structure through the use of a molded nylon bobbin and nonhygroscopic insulation throughout. All SSO transformers are vacuum impregnated to assure dependable operation under high humidity conditions.

Recent Catalogs

• • •A new folder "On Air," to be published every other month by **Broadcast Equipment Section, Radio Corp. of America,** Camden 2, N. J., is now available.

NOTICE

"Information for our News and New Products section is warmly welcomed. News releases should be addressed to Industry Research Division, Proceedings of The I.R.E., Room 707, 303 West 42nd Street, New York 18, New York Photographs and electrotypes, if not over two inches wide, are helpful. Stories should pertain to products of interest specifically to Radio Engineers."

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

Two-Variable Plotting Board

The Model 205 Variplotter, a new two-variable, graphic recorder, has been introduced by **Electronic Associates, Inc.,** Long Branch, N. J., to the instrumentation field.

The Variplotter graphically presents one variable as a function of another wherever the variables can be expressed in terms of dc voltages. Information such as stress versus strain, pressure versus temperature, speed versus torque, frequency response, antenna pattern, hysteresis loops, tube characteristics, and process control is readily obtained with the instrument.

Another feature of the Variplotter is its 30-inch plotting surface, which is equipped with back-lighting to aid viewing. This does not restrict use to 30-inch square charts. If smaller standard forms are presently employed, these may be used.

The static accuracy is 0.05 per cent of full scale at 70° Fahrenheit. The dynamic accuracy averages 0.05 per cent of full scale plus the static accuracy at a writing speed of 8.5 inches per second. The maximum pen and arm accelerations are 350 and 150 inches per second squared, respectively. Slewing speeds of both pen and arm are 10 inches per second.

Accessories are available which will increase the usefulness of the Variplotter; for example, conversion kits which add an additional arm and pen to the plotting board, enabling it to present the equivalent of four variables, analog computer components or systems to meet a specific requirement; and a standard line of components, such as dc amplifiers, resolvers, reference voltage standards, and power supplies, etc.

Precision Attenuators

Precision decade attenuators for laboratory use and for building into other equipment are now available from the **General Radio Co.,** 275 Massachusetts Ave., Cambridge 39, Mass.

Designed to operate at audio- and low-radio frequencies, these attenuators are built into a compartmented metal casting, which affords such excellent shielding between pads that some of the units can be operated at frequencies as high as 1 Mc, with errors of less than 2 per cent.

The Type 829 decade attenuator units can be built into speech and ultrasonic equipment, recording channels, measuring devices, etc. Characteristic impedance is 600 ohms for both H and T types. Tapered units are available for matching to other impedances.

The Type 1450 decade attenuators are assemblies of attenuator units in metal cabinets for laboratory bench use. Two- and three-dial boxes are available with maximum attenuation of 110 db.

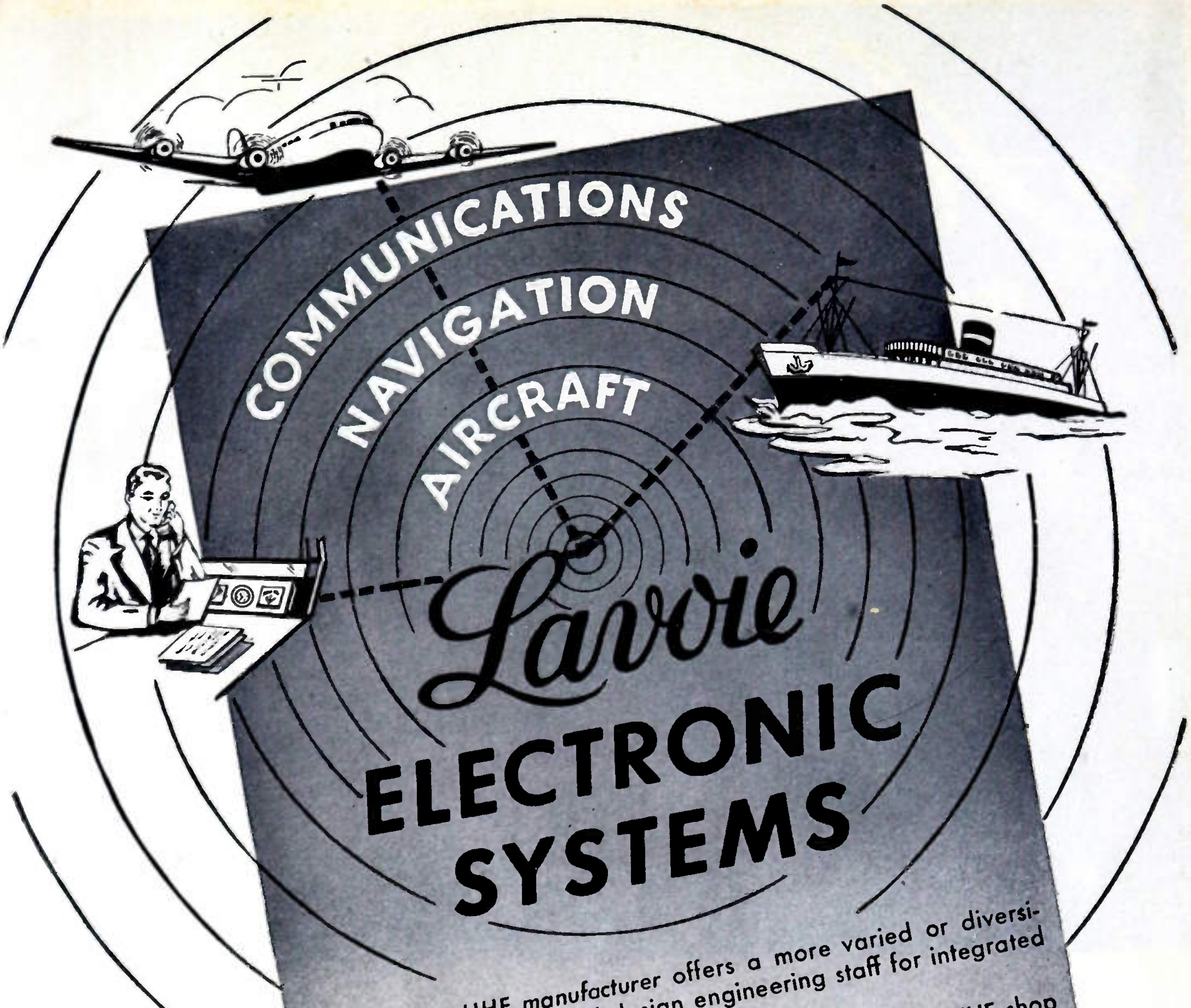
Omni Test Equipment

To fill a need for test equipment, required for field overhaul and major service on omnirange navigational equipment, the **National Aeronautical Corp.,** Wings Field, Ambler, Pa., has designed the Model T-3 test set.

The T-3 generates all of the signal components of an omnistation, and the omnitrack, which is transmitted, may be set on any bearing by means of an accurately calibrated dial. This permits a complete calibration test to be made on any omni equipment, without removing the receiving equipment from the plane.

For major overhaul work on the bench, the test set produces the necessary signals to make a complete adjustment of all the necessary balancers and compensations that are required. The T-3 also produces the signals necessary for testing the adjustment of phase localizer approach equipment and provides the necessary signals for tone localizer and VAR equipment.

(Continued on page 64A)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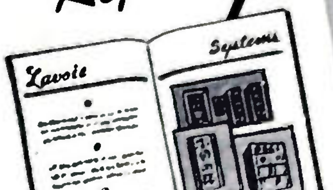
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... a major advance in studio-type image orthicons

A NOTABLE PRODUCT of RCA leadership in tube research and engineering—the new RCA-5826 image orthicon provides important refinements over previous types of television camera tubes for studio use.

The new RCA-5826 combines exceptionally high sensitivity, a resolution capability of better than 500 lines, high signal-to-noise ratio—about twice that of outdoor camera types—and improved gray-scale rendition in the vicinity of the "blacks."

Having the same spectral response as the companion outdoor pickup type RCA-5820—a response closely approaching that of the eye—this new studio camera tube permits portrayal of colors in nearly their true tone gradation. The use of the RCA-5826 in the studio and the RCA-5820 outdoors facilitates the combination of indoor and outdoor pickups on the same program . . . improvements that are automatically extended to every receiver.

ANOTHER new RCA tube . . .

...the RCA 6AX5-GT Heater-Cathode Type Full-Wave Vacuum Rectifier: Designed to operate from a common 6.3-volt heater supply in ac-operated sets or auto receivers. Has the same heating time as other heater-cathode types, thus permitting the use of filter capacitors having lower peak voltage ratings than required for filament-type rectifiers. Delivers 125 ma. at 350 volts to a capacitor-input filter.

THE FOUNTAINHEAD OF MODERN TUBE DEVELOPMENT IS RCA

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ELECTRON TUBES
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PROCEEDINGS OF THE I.R.E.

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 38 — *June, 1950* — NUMBER 6

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John D. Reid

Regional Director, 1950–1951

John D. Reid, Regional Director of the Central Region, was born in Morristown, New Jersey, on March 18, 1907. Mr. Reid attended the University of Pennsylvania during the years 1923–1925, taking a combined Wharton-Engineering course. After spending a year with the Arcturus Radio Tube Company as a circuit application engineer, he returned to the University as a special student during 1928 and 1929.

He then joined the Norden-Hauck Radio Manufacturing Company as chief engineer, and developed the "Admirality Super 12," the first all-wave superheterodyne with a high frequency intermediate frequency (475 kc) and built-in band-spread short-wave ranges.

Mr. Reid was associated with the Radio Corporation of America as a radio and television development engineer in 1930. At that time he was responsible for the development of all wave tuning systems ("Magic Brain"), FM detector systems, and the television input tuner used on their prewar receivers. In 1934 Mr. Reid took part in a radio field survey trip made through the Caribbean Islands, the northern part of South America and Central America. During 1937, Mr. Reid spent five months in Russia where he was assigned to service as a special consultant to the Soviet Union.

He became affiliated with Crosley Division, Avco Manufacturing Company, Cincinnati, Ohio, in October, 1940, heading an Advance Development Group. Prewar developments included the large radius phonograph needle and the decimal tuning double superheterodyne.

During the war he was in charge of Crosley's development work on the proximity fuze, for which he was awarded the Naval Ordnance Development Award and the President's Certificate of Merit. Since the war Mr. Reid has been Manager of Research for the Crosley Division.

Largely responsible for the inception of the Cincinnati Spring Technical Session on Television, Mr. Reid has taken part in numerous IRE activities, including service as Chairman of the Cincinnati Section during 1945 to 1947; Chairman of the Cincinnati Spring Technical Meeting, 1947; Chairman of the IRE Papers Procurement Committee, 1948 and 1949; and membership on the following committees: Papers Procurement, Radio Receivers, Board of Editors, Membership, Papers Review, and Professional Groups. Mr. Reid was made a Fellow of the Institute in 1948.

In Government and industry alike, the skilled and respected engineer is increasingly offered executive responsibility. A great obligation rests upon those engineers who reach such positions of authority where they must make thoughtful decisions based on human needs and reactions, as well as on measurable data. These men must meet the definition of a great engineer, namely, that he is a man who, given incomplete and partially incorrect premises, can draw correct conclusions!

Fortunately there are such engineers, distinguished alike by technical ability, good judgment, social consciousness, and sincerity of purpose. Their thoughts are of major importance to their fellow engineers. The following guest editorial by a member of the Federal Communications Commission is particularly noteworthy. It was written by one whose Fellow award in the IRE carries the citation: "For his contributions to the development of the maritime mobile radio services and his leadership in promoting measures for enhancing the safety of life and property at sea."—*The Editor.*

Our New Environment of Decision

E. M. WEBSTER

About three years ago I found myself a Commissioner on the Federal Communications Commission, and thus in the position of "high level administrator" within our Government. I was appointed to this position from the ranks of the radio engineering profession. Immediately, I was faced with heavy responsibilities and many complex and divergent problems. Out of this experience I pose this question: "How can technicians achieve the breadth of view that is required on the part of administrators concerned with some of today's public policy problems?"

The need for assuming this heightened responsibility stems, I believe, from two sources—the enlarged role of our government in technological fields as the result of postwar developments, and the new position of the United States as a leader in world affairs. How can a technical man adjust himself to such responsibilities? As Harvard's Sumner Slichter recently pointed out, "Is it not asking the impossible to expect men to be both able administrators and important thinkers in the field of public policy? How can men find time both to manage an enterprise in a highly competitive and rapidly changing economy and to become sufficiently well informed in the fields of economics, political science, sociology, and other disciplines to do first-class thinking in the field of public policy?"[1]

To illustrate the need for going beyond a technical viewpoint in many radio engineering matters, let's take the television problem, for example. In establishing a national television policy, the serious technical problems which the Federal Communications Commission is of necessity appraising today may only be a minor element in the final national television policy that eventually will be promulgated.

The Commission must, in addition, consider many intangible economic aspects of our proposed television industry. For instance, it has been estimated that it may cost well over a billion dollars per year to provide a minimum of national television service. Will that kind of money be available to the industry? If we take the position that "the public interest, convenience, and necessity" require that we have a national television service for all the people, must not the government be just as concerned with where this money is coming from as it is with television's various technical problems?

I trust that you have already recognized that this economic question also closely ties into a political problem. What constitutes a national television service? What elements of the population must be served before we can justly say that we have a national television service? These are questions I think our politicians will ask. There is a further question, too, of the kind of national television service the public is entitled to. Do we escape the responsibility to see that a minimum standard of programming efficiency is maintained to protect the public from the social impact of this new instrument of communication? To achieve this, should we authorize some new system of "phonevision" or metered television which allows the public a voice in programming? Yet, by virtue of the fact that the public then pays for some of the programs it hears, will not the government perhaps be forced to take on greater responsibilities to see that the demands of this paying public are answered?

Development of breadth of view is not easy. However, a previous editorial in this space called "Feedback" pointed up one possible error in our thinking, I believe. That editorial implied that the evolutionary integration of social and natural science would enable engineers to do a better job and to broaden their thinking. I do not believe that the application of more science to problems already raised by natural science is necessarily the total answer. We are continually, in the formation of national policies, making decisions in which intangible facts and certain assumptions must be inherently accepted as elements in the situation: there often is just no "good" choice among the alternatives, and time alone can tell whether or not the chosen course of action will work out. It is impossible to apply a "slide-rule" type of thinking to such decisions. To make such decisions requires a certain open-mindedness, a willingness to set up a program which has the risk of failing, and the fortitude to reconstruct that failure—in the light of new developments—into a positive policy. It is this integration of open-mindedness and willingness to take calculated risks that we technicians must develop if we are fully to meet our responsibilities in the new environment of decision in which United States citizens are working today.

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[1] *Harvard Business School Alumni Bulletin.* p. 132, Autumn, 1949.

Distant Electric Vision*

J. D. McGEE†

***Summary*—An outline of the history of television is followed by a detailed description of the design and development of some English television pickup tubes.**

JUST OVER FORTY YEARS AGO a short letter appeared in *Nature* under the title "Distant Electric Vision." It was signed A. A. Campbell-Swinton.[1] I have ventured to borrow that same title for my paper, since I would like you to regard what I have to write as a part of the story of the development of the brilliant idea first proposed in that letter before even the word "Television" had been coined. I shall return to Campbell-Swinton's proposal presently, but first I must review briefly what had gone before. The photoelectric effect, discovered in 1873, was very soon realized to be the key to "distant electric vision." The first practical device for generation of television signals was, in fact, proposed only eleven years later by Nipkow.[2] The apparatus proposed by Nipkow is shown in Fig. 1. An image *4* of the scene to be transmitted *5* is formed by the lens *6* on the surface of a disk *1* which can rotate about its axis *2*. Around the periphery of this disk is a spiral of small apertures *3* spaced at equal angular intervals and each spaced radially from the center of the disk at distances increasing successively by the height of an aperture. At any given time light can pass through one aperture only and enter the photocell *7*, where it releases an electron current proportional to the intensity of the light; that is, proportional to the brightness of the image at the point where the aperture happens to be at that instant. These electrons are collected on the anode of the photocell by the electric field maintained by the battery *8* and the current passing through the resistance *R* produces a voltage change across it. As the disk rotates, the aperture will scan across the optical image, and the light passing through it will fluctuate proportionally to the light and shade along that line of the image. The voltage fluctuations across the resistance *R* will be also proportional to the light fluctuations. Thus, a "picture signal" will have been generated for that line of the image. As one aperture completes its scan of one "line" of the image, the next aperture begins to scan the next line of the picture, and so on, until the whole image has been scanned in one complete revolution of the disk.

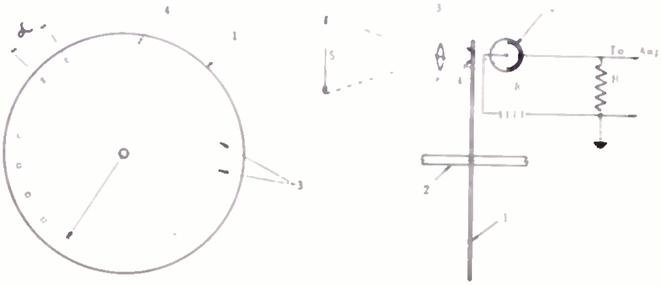

Fig. 1—The Nipkow disk.

For modern television pictures of, say, 405 lines and 25 pictures per second, the disk must rotate at 1,500 rpm. The diameter of the disk must be as large as possible in order that the apertures may be made of a reasonable size; otherwise, the signal current will be very small and the signal-to-noise ratio very poor. However, a disk of, say, 3 feet in diameter rotating with high precision at 1,500 rpm, is not exactly a convenient or mobile piece of equipment. Moreover, an amplifier of sufficient bandwidth for a 405-line television picture signal (say, 3 Mc) requires a peak input signal of about 0.1 μa to override by an order of magnitude the noise of the first stage of the amplifier. If we assume a high photoelectric efficiency of 100 μa per lumen for the photocell it follows that 10^{-3} lumens must fall into the photocell to give this signal current in the peak whites of the image. Since there are 200,000 picture points in such a picture, this corresponds to a light flux of 200 lumens in the whole image, assuming it to be of uniform brightness. Again, assuming an area of the image on the disk of 4 × 5 inches, this corresponds to light flux of 1,440 lumens per square foot or 1,440 foot-candles illumination.

The illumination in the plane of the image I_i is related to the illumination on the scene I_s by the well-known formula

$$I_i = \frac{I_s \times \gamma}{4\,A^2},$$

where γ is the reflection coefficient of the scene which may be taken as unity for a perfect white diffusing surface, and A is the numerical aperture of the lens, say $f/2$.

Then

$$I_s = \frac{I_i \times 4\,A^2}{\gamma} = 23{,}000 \text{ foot-candles.}$$

Such illumination is about a hundred times greater than that used in television studios today and would be quite intolerable.

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* Decimal classification: R095×R583.6. Paper received by the Institute, November 4, 1949. Presented, I.R.E. Radio Engineering Convention, Sydney, Australia, November, 1948. (Since Dr. McGee was unable to present this paper personally to the Convention, the entire manuscript was delivered by means of a magnetic tape recorder.) Reprinted from the *Proceedings of the Institution of Radio Engineers*, Australia, vol. 10, pp. 211–223; August, 1949.

† E.M.I. Research Laboratories, Ltd., Hayes, Middlesex, England.

[1] A. A. Campbell-Swinton, "Distant electric vision," *Nature*, (London), vol. 78, p. 151; June, 1908.

[2] P. Nipkow, Deutsches Reichs Patent No. 30,105; January 6, 1884.

The advent of the electron multiplier made it possible to multiply the initial photoelectric current by a very large factor, 10^6 or more, without seriously changing the signal-to-noise ratio of the initial current which is due to probability fluctuations in the photoelectric emission. In this way the input signal to the amplifier can be made large compared with the noise of the first stage, and the limiting noise then becomes that of the primary photoelectric current. A simple calculation shows that the illumination required is then very much less but still about 1,000 foot-candles, even when the most favorable conditions are assumed.

Thus, the inconvenience of the mechanical scanner, together with its very low sensitivity, has doomed it to oblivion except possibly for film scanning, but, even here, all-electronic systems have found greatest favor.

CAMPBELL-SWINTON'S PROPOSAL

I never cease to wonder at the vision and imagination of that great television engineer, the late A. A. Campbell-Swinton, F.R.S., who as long ago as 1908 not only saw clearly the fundamental limitations of the mechanical systems, but actually proposed the all-electronic system which, with modern technical embellishments, has become the television system of today. I believe I need not apologize to you for showing you in Fig. 2 the actual schematic diagram of the system for "distant electric vision" which he proposed in 1908 and amplified in a paper in 1911.[3] In this figure, you see the transmitter on the left; the receiver on the right. The transmitter is a cathode-ray tube *A*, in which an electron beam from the cathode *B* is scanned by crossed magnetic fields *D* and *E* line by line over a mosaic of photoelectric cells *J*. An optical image of a scene *N* is focused by the lens *M* on to the other side of this mosaic of photocells which become charged by the loss of photoelectrons. These charges are discharged at each scan of the electron beam to the grid *L*. These fluctuating "picture signals" are conducted over a line to modulate an electron beam which scans the fluorescent screen of the receiving cathode-ray tube in synchronism with that of the transmitting tube. Here are two ideas of capital importance; the use of inertia-less cathode-ray beams for scanning at very high speed, and the mosaic of photocells in the pickup tube. I think one may well claim that this was the seed from which present-day television engineering has grown.

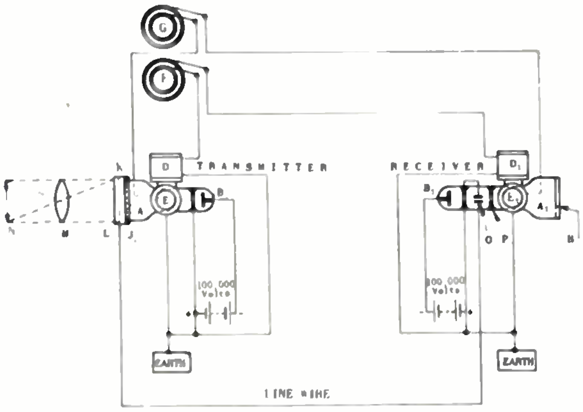

Fig. 2—Electronic television system proposed by Campbell-Swinton.

FARNSWORTH DISSECTOR

One feature of Campbell-Swinton's scheme was realized in the Farnsworth dissector tube,[4] namely, that of all-electric operation. This tube was first described in 1934 and, because it still may find an application in film scanning, it is worthy of a brief description here.

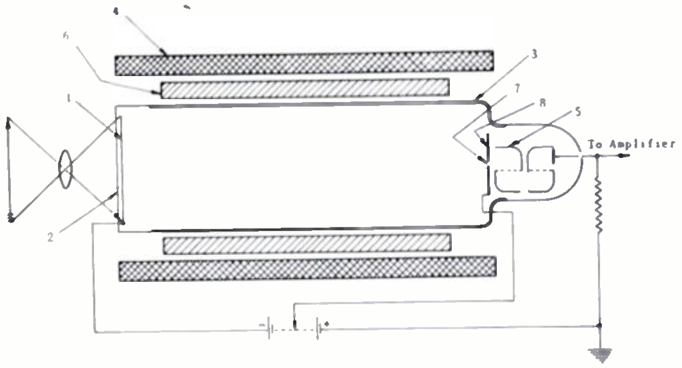

Fig. 3—Diagram of Farnsworth dissector tube.

The tube and associated equipment are shown in Fig. 3. An optical image of the scene is focused on a transparent conducting photoelectric layer *1* on the inside surface of the flat end wall *2* of the high-vacuum tube *3*. Electrons are emitted from this surface under the influence of, and proportional in number to, the incident light. These electrons are accelerated by an axial electrostatic field and focused by means of a uniform axial magnetic field produced by the solenoid *4* to form an electron image in the plane of the electrode *8*. A pencil of electrons passes through the small aperture *7* and falls on the first plate *5* of a multistage electron multiplier, the output of which is fed to a suitable amplifier. Two pairs of coils, one of which is shown at *6*, actuated by frame and line frequency currents of sawtooth wave form, produce two transverse magnetic fields at right angles to one another which scan the whole electron image across the aperture in a series of lines. Since the aperture *7* must allow only those electrons from one picture point to pass through to the multiplier, it follows that for a 400-line television picture the linear dimensions of the aperture must be 1/400th of the dimensions of the electron image, and, since there are 200,000 picture points in such a picture, only one part in 200,000 of the electron emission from the photocathode passes through the aperture. In practice, this current is so small (of the order of 10^{-11} amperes) that if it were fed directly to the input of an amplifier of the necessary bandwidth,

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[3] A. A. Campbell-Swinton, "Presidential address," *Jour. Roentgen Soc.*, vol. 8, p. 7; January, 1912.

[4] P. T. Farnsworth, "Television by electron image scanning," *Jour. Frank. Inst.*, vol. 218, p. 411; October, 1934.

the signal produced would be far smaller than the amplifier noise. However, by multiplying the primary electron current in a multiplier by a factor of 10^6 or more the output can be increased until the signal is much greater than the amplifier noise. The signal-to-noise ratio of the signal is then determined by the random fluctuations of the primary photoelectric current which is not changed appreciably in the process of multiplication. Thus, the Farnsworth dissector is the electric analogue of the Nipkow disk and to a first approximation, the sensitivities of both signal-generating systems are the same. Minor advantages make the Farnsworth dissector slightly more efficient, but it still requires a scene illumination of about 1,000 foot-candles. This is, of course, a hundred times the illumination required by modern television cameras.

CHARGE STORAGE PRINCIPLE

It will have been noticed that, in both the Nipkow and the Farnsworth methods, only a minute fraction of the light from the scene is effective in producing picture signal—in fact, the light from one picture element. Less than 10^{-5} of the light or photoelectric current is used. It is obvious that if all of the photoelectrons released by the light in the Farnsworth dissector could be stored during the frame-scanning period and discharged once per frame period, the gain in efficiency would be enormous—a factor of 10^5 or more.

The obvious line of attack on this problem was clearly suggested by Campbell-Swinton's mosaic of photoelectric cells. If an optical image is formed on a mosaic of minute photoelectric cathodes, each of which is associated with a minute condenser, and the photoelectrons liberated from the cathodes by the light are saturated to a common anode, charges will be built up on the condensers which are at all points proportional to the incident light. Thus, a reproduction of the optical image is built up in electric charges. If now this mosaic of condensers is scanned by a suitable commutator, e.g., a beam of electrons, which discharges those condensers in the area of one picture point simultaneously through a signal resistance, the potential fluctuations across this resistance will be the required picture signal.

A single cell of such a system is shown diagrammatically in Fig. 4. Light L falling on the cathode of the photocell P liberates electrons in proportion to its intensity. These are collected by the anode which is maintained at a positive potential by the battery B. This current flowing through the resistance R charges the condenser C. If now, once per frame period, this condenser is discharged by the commutator A (which may be an electron beam with an impedance r) a pulse of charge will pass through the resistance R proportional to the integrated light flux on the photocell during the period of storage. The potential change across R will be proportional to the discharge current and may be applied to the grid of the first valve of an amplifier chain. The discharge time of the circuit consisting of C, r, and R must be not less than the time required to scan one picture element, i.e., 2×10^{-7} seconds. The potential drop across R during the charging period is the picture signal that would be obtained without storage; hence, the increase in signal strength due to storage is the ratio of the charging time (0.04 second) to the discharging time 2×10^{-7} seconds of the condenser C. That is, a gain of 2×10^5 is achieved. If in practice only a small part of this theoretical gain could be realized, say 1 per cent, it would still represent an increase in sensitivity by a factor of 2,000.

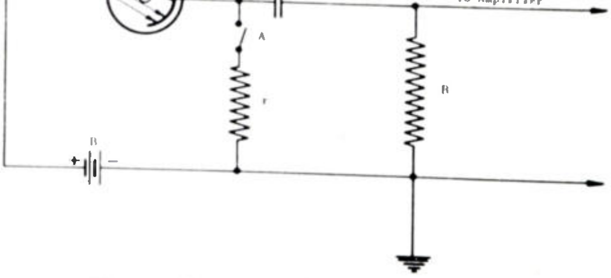

Fig. 4—Elementary charge storage circuit.

The first account of a television pickup tube employing charge storage was published by Zworykin in 1933.[5] This tube was named the "iconoscope" and it has been the main tube employed in American television cameras until quite recently. Work had been in progress in the Electric and Musical Industries Ltd. Laboratories and, in 1932, Tedham and the author succeeded in producing television signals with a tube which proved to be fundamentally the same as the iconoscope but which differed from it in many important practical details. Continuation of this work led to the development of the emitron[6,7] which was used from the beginning of the first regular public television service instituted by the BBC in London in 1936. It is still the only tube used for studio broadcasts by the BBC. Further development led to the super-emitron.[7,8]

I will now give a brief outline of the operation of the emitron and super-emitron before going on to describe the CPS emitron which has been developed since the end of the war and has only recently gone into service.

THE EMITRON

The tube with its immediate operating circuits is shown in Fig. 5, while a very much enlarged section of the target is shown in the inset. The highly evacuated envelope *1* carries a side tube in which a more or less conventional electron gun is mounted. The gun consists

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[5] V. K. Zworykin, "Television with cathode-ray tubes," *Jour. IEE*, vol. 73, p. 437; 1933.

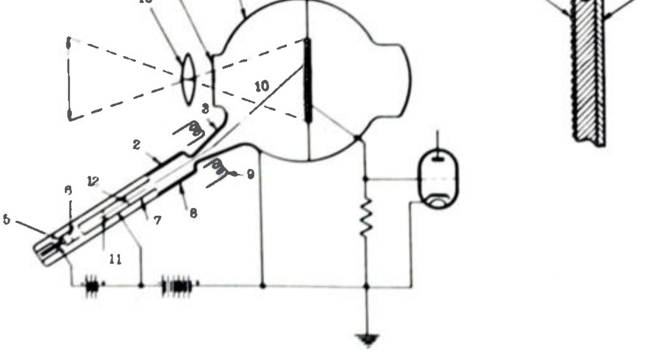
[6] J. D. McGee and H. G. Lubszynski, "E.M.I. cathode-ray television transmission tubes," *Jour. IEE*, vol. 84, p. 468; April, 1939.

[7] J. D. McGee, "The development of high definition electronic television in Great Britain," *Proc. I.R.E.* (Australia), *Proc. World Radio Convention*, Sidney, Australia; April, 1938.

[8] H. G. Lubszynski and S. Rodda, British Patent No. 442,666; May 12, 1934.

of a cathode *5*, modulator *6*, first anode *7*, and second anode *8*. The latter is held at earth potential and extends some distance into the spherical part of the tube where it acts as anode for the photoelectric mosaic also. The cathode of the electron gun is held at about −1,500 volts and the electron beam is focused by the electrostatic electron lens between the first and second anodes, *7* and *8*. The beam is scanned over the mosaic by two pairs of coils, one of which is shown at *9*. Since the beam falls obliquely onto the target, a conventional scanning raster would appear as a trapezium or keystone shape. This must be corrected by modulating the line-scan amplitude by the frame-scan amplitude. This is termed "keystone correction." Further, the frame-scan wave form must be slightly distorted from linear to obtain even spacing of the lines from top to bottom of the mosaic. The electron beam comes to a focus on the surface of a sphere, the center of which is the point of deflection of the beam. This sphere intersects the plane of the target in a circle and at all points off this circle the beam is more or less out of focus. The loss of definition due to this difficulty is minimized by restricting the beam to a very narrow pencil by limiting apertures in the gun so that its depth of focus is increased. This can be done because only a very small beam current of a few tenths of a microampere is required. All these inconveniences in design are accepted in order to be able to project the optical image normally onto the target. The lens *13* forms an image of the scene through the flat window *4* on the surface of the target, which is also scanned by the electron beam. The lens is usually 6½″ focal length with a numerical aperture of *f*/3.

The target is 4×5 inches (if the agreed picture aspect ratio is to be 5 to 4). A small section of it, greatly magnified, is shown in the inset of Fig. 5. It consists of a sheet of highly insulating dielectric *D*—for example, mica—coated on the side which is not scanned by the electron beam with a conducting layer *S*, known as the signal plate. This signal plate is connected to earth through a signal resistance and to the grid of the first tube of the amplifier. On the side of the dielectric which is scanned by the electron beam is formed the photoelectric mosaic *M*. This may be regarded as a vast number of minute islands of photosensitized metal each separated and highly insulated from its neighbors. Also, each mosaic element forms a small condenser with the signal plate.



(a) (b)

Fig. 5—(a) The emitron. (b) Cross section of emitron target.

The basic material of the mosaic elements is usually silver which can be formed into a mosaic by evaporating a very thin layer and aggregating this by heating. The mosaic may also be formed by evaporating the metal through a fine metal mesh which acts as a stencil and in several other ways. The silver mosaic is made photosensitive by oxidation and subsequent treatment with cesium. Here a difficulty is encountered. It is difficult to obtain high photosensitivity of the mosaic elements while preserving good insulation between them because the metallic cesium tends to form a slightly conducting film over the surface of the dielectric between them. A compromise has to be accepted by which we obtain about half the maximum possible sensitivity. Also, the mosaic elements cannot cover more than about 50 per cent of the target surface so that another factor of two is lost in photoelectric efficiency. We can obtain an efficiency of about 10 μa per lumen instead of over 30 μa per lumen that can be obtained from a normal photocell. These points are important and will be referred to when we consider the super-emitron.

We must next consider the mechanism of signal generation in the emitron. This is a complex problem and I ask your pardon if my treatment of it in this paper leaves many unsatisfactory loose ends.

In its simplest form we may be sure that the light of the image which falls on the mosaic liberates photoelectrons at all points in numbers proportional to the light intensity. If these negative electrons are removed from the mosaic, for example, to the anode of the tube, a distribution of positive charges will be built up continually on the minute condensers formed by the mosaic elements and the signal plate. If now the electron beam scans the mosaic line by line, neutralizing these positive charges point by point, equal and opposite (i.e., negative) charges will flow to earth through the signal resistance *R*. These charges and the voltages produced across *R* will be proportional to the integrated intensity at the corresponding points in the image over the previous frame period, i.e., 1/25th of a second. Thus a rapidly moving object should be reproduced with a loss of definition corresponding to that seen in a photograph taken with a time exposure of 1/25th of a second. In fact, this is not found to be the case. A television picture of a rapidly moving object produced by an emitron shows it as a series of sharply defined images. This means that uniform charge storage cannot be taking place during the whole frame period. Moreover spurious signals are produced which cannot be explained on this simple theory.

These experimental observations can only be explained when we take into account the fact that the scanning beam electrons, which reach the mosaic with an energy of 1,500 electron volts, each releases between 5 and 10 secondary electrons. These relatively slow secondary electrons stabilize the potential of the mosaic at approximately that of the nearest electrode of fixed potential, i.e., the second anode. There is therefore no electrical field of sufficient strength to saturate the photoelectrons away from the mosaic. If we now consider the mosaic being scanned by the beam but without light falling on it, it is clear that over a period of time it can neither gain nor lose charge; that is, only one secondary electron can leave the mosaic for each primary falling on it. The remaining secondary electrons must return to some part of the mosaic, but not necessarily the point from which they were released.

Now the point on the mosaic on which the beam is falling at any instant will initially lose between 5 and 10 secondary electrons for every incident primary electron. Thus, its potential will be driven positive until it can recapture all but one of the secondaries released by each primary electron. This charge and discharge of mosaic elements is illustrated for two successive elements *A* and *B* in Fig. 6. It is estimated that the potential rise of the mosaic during scanning is about +6 volts relative to the surrounding areas. As the scanning beam passes on to the next point on the mosaic the secondary electrons liberated from it tend to be deflected towards, and captured by, those areas which have just been scanned since they are the most positive areas in the neighborhood.

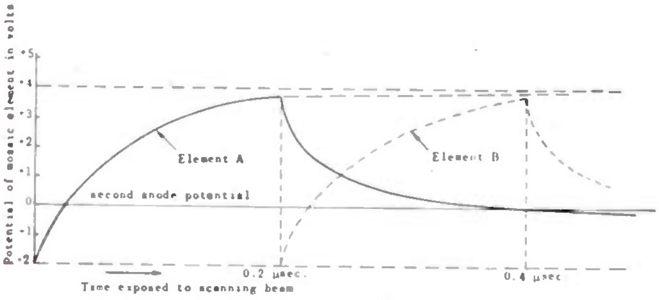


Fig. 6—Charge and discharge curves of mosaic elements.

It will, I think, be clear that this exchange mechanism cannot be uniform over the whole scanning cycle. The charges on the surrounding areas of the mosaic must be very different at the beginning and end of line and frame scans and this will influence locally the sharing of secondary electrons between the second anode and the surrounding mosaic. Other factors such as the asymmetry of the tube, variations in the focus of the scanning beam, local charges due to the light image, etc., can also affect this sharing of electrons between the mosaic and the second anode. It is this variation in the sharing of the secondary electrons between the mosaic and the second anode that causes the spurious signals. These are illustrated in Fig. 7, in which line *A* shows the wave form of the uncorrected signals for three successive lines. There are large spurious pulses at the beginning and end of each line, and a spurious low-frequency signal known as "shading" is superimposed on the true signal. In line *B* is shown the artificially generated correcting signals, "Tilt" and "Bend," which when mixed with the crude signals result in the train of signals shown in line *C*. The large spurious signals between lines are then suppressed, leaving the corrected signal as it appears in line *D*. Finally, synchronizing signals are inserted between the lines as shown in line *E*.

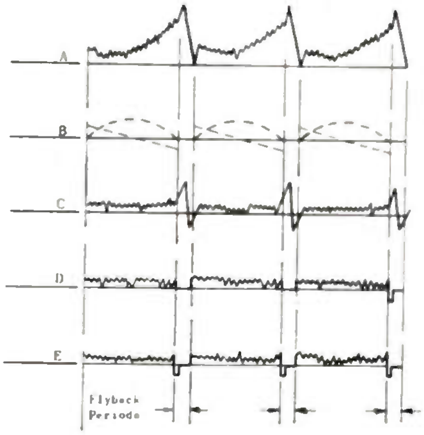


Fig. 7—Emitron signals.

If, when once corrected, the shading signals remained constant, they would be of little consequence. Unfortunately they do not, but change appreciably with distribution and intensity of light falling on the mosaic. Hence, the shading controls must be constantly adjusted as the scene changes in front of the camera and, at the lower limit of light levels when the amplifier gain must be increased, the satisfactory correction of shading becomes almost impossible. In fact, the lowest level of illumination at which an emitron camera will work satisfactorily is usually determined by uncorrected shading signals rather than by amplifier noise. Another important result of this mechanism of operation is the fact that at no part of the scanning cycle is a signal generated which can be regarded as corresponding to the true black level of the picture. This is a great disadvantage, since there is then no absolute method of re-establishing the true black level of a picture after it has passed through an alternating-current system. It must be left to the arbitrary adjustment by some operator.

We have seen how the unwanted signals are generated. We must now consider the true picture signals which are wanted. As I have explained above, each element of the mosaic is driven strongly positive by the scanning beam and then gradually falls in potential until the scanning beam reaches it on the next cycle. Thus, there is an area of the mosaic that has just been scanned which is about 6 volts positive relative to those elements which are about to be scanned. Hence, any photoelectrons liberated in a narrow band across the mosaic in front of the scanning beam will find themselves in a quite strong electric field attracting them to those areas of the mosaic which have just been scanned. This narrow band within which the photoemission is saturated is estimated to extend something like 1 centimeter beyond the last line to be scanned. Most of the photoelectrons liberated in this area leave the mosaic element from which they were liberated and consequently charge them positively. They are collected in a random distribution on the mosaic areas that have just been scanned which they drive more negative in potential. Beyond the narrow band of mosaic from which the photoelectrons are completely removed, the percentage of photoelectrons collected will decrease with distance. Over the greater part of the rest of the mosaic the photoelectrons which are liberated by the light will fall back onto the mosaic at random and so will not contribute appreciably to the stored charges. As the scanning beam passes over these elements on which varying positive charges have been stored, they will all be driven up to a fairly constant peak positive potential. Thus, at a point which has been exposed to light and on which positive charge has been stored, the potential change on being scanned will be less than for an element which has not been exposed to light. It is, in fact, the differences in these potential changes as the beam scans the mosaic that constitute the picture signal.

It will be clear from the above that the emitron operates in a manner very similar to a camera using a focal-plane shutter, and the charges that give rise to picture signal are built up in a very short period of time, about 1/250th of a second, before the scanning beam reaches each part of the mosaic. This explains why fast-moving objects appear as a succession of sharp images rather than as a continuous blurred image.

This same mechanism of operation explains the low sensitivity of an emitron which gives only about 5 per cent of the sensitivity to be expected from a full storage tube. It will be seen that the photoelectric emission is used to build up charges on the mosaic only during a small fraction of each frame period.

Another consequence of this mechanism is that the signal output is not proportional to the light input. This is to be expected since, as a mosaic element which is exposed to light becomes more positive, it tends to collect secondary electrons itself. If we express the signal S in terms of the illumination of the mosaic I by the formula

$$S = I^{\gamma},$$

we find that γ is not greater than about 0.75. This was rather a lucky accident, since this reduction in contrast of the generated signal compensates the inevitable increase in contrast which occurs when such signals are displayed on a normal cathode-ray receiving tube. However, when the scene to be televised lacks contrast there is nothing that can be done to improve its appearance to the viewer.

One final curiosity which can frequently be noticed in an emitron picture can be explained on this theory. If the picture to be televised has horizontal lines that are emphasized, it is found that a white horizontal bar is followed by a black streak extending possibly half the width of the picture in the direction of scanning and vice versa. This is believed to be due to the charges on the mosaic that are about to be scanned influencing the level of the peak positive potential to which the mosaic elements are driven by the scanning beam. This may be regarded as a loss in response at the lower frequencies and it can be corrected to some extent by introducing a bass boost in the amplifier chain between 5 and 10 kc.

The emitron tube mounted on the base of the camera is shown in Fig. 8. The first few stages of the amplifier are built into the camera chassis in order to keep the input capacity of the amplifier as low as possible. Also, the line and frame scan amplifiers are built into this chassis. The complete camera is shown in Fig. 9, showing the objective lens and the view-finding and focusing lens. Such a camera will give a very satisfactory picture

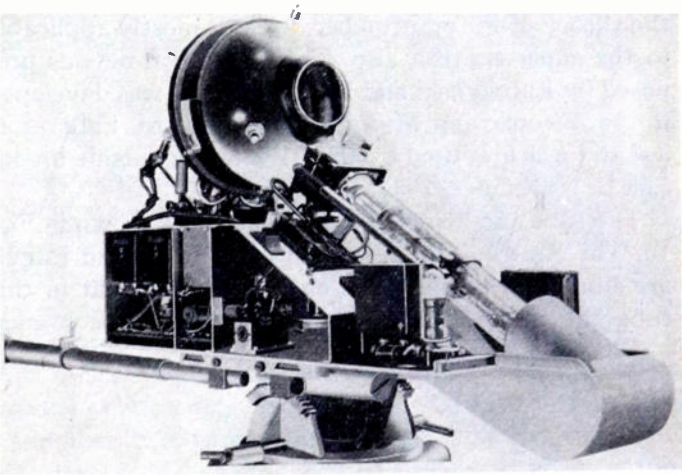


Fig. 8—Emitron tube mounted on base.

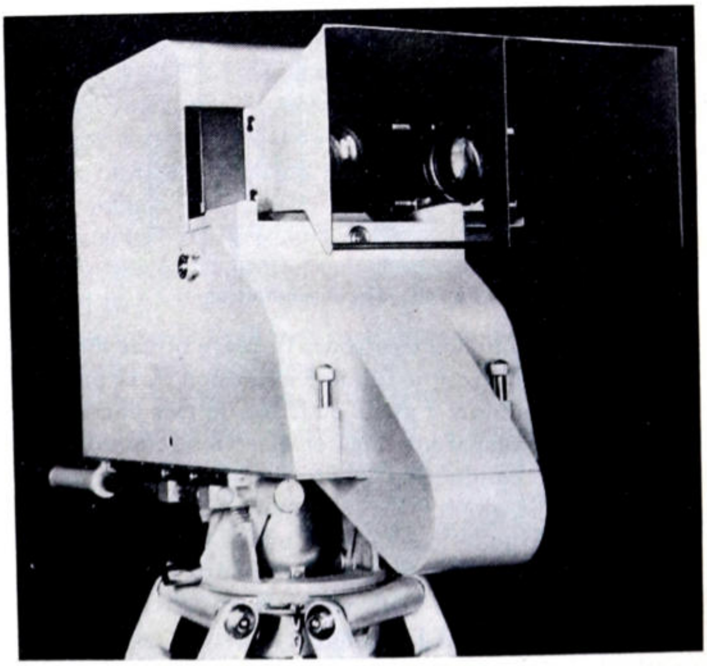


Fig. 9—Emitron camera.

with about 200 foot-candles of illumination incident on a studio scene when the lens is at full aperture (*f*/3). However, under these conditions the depth of focus is very small, which imposes restrictions on the studio production. This can be improved only by increasing the light and stopping down the lens. A worthwhile increase in light becomes expensive and uncomfortable. The definition of the picture is excellent and the geometrical distortion of the image practically negligible. However, the pictures are frequently marred by uncorrected shading and the absence of a definite black level. Moving objects do not appear to lose definition and very rapidly moving objects may even appear as a series of well-defined separate images. It is a doubtful point whether this is preferable to a single blurred image which is more nearly what is seen by the eye.

THE SUPER-EMITRON

I have spent a considerable time over the details of the theory of the emitron because it is mostly applicable to the super-emitron also. This type of tube was proposed by Lubszynski and Rodda in 1934, was developed in the Electric and Musical Industries Ltd. Laboratories[8] and was first used by the BBC for an outside broadcast in November, 1937.

The tube and associated equipment are shown in Fig. 10. The electron gun *1*, scanning system *2*, and target, are similar to those of the emitron except that in this tube the mosaic is not photosensitive but is a good emitter of secondary electrons. In place of the flat window of the emitron, a tubular extension *4*, is provided with a flat window *5* at its end. Inside this flat window a sheet of transparent mica or glass is mounted, and a transparent conducting photocathode *6* is formed on that side of it which faces the target. Contact is made to this photocathode from a solid metal border *7* and it is held at about 500 volts negative relative to the metal coating on the tube wall, which is an extension of the gun second anode. A light image is formed on the photocathode by the lens *8* and the photoelectrons liberated by the light are accelerated by the axial electric field towards the target. A magnetic field produced by a current in the coil *9* brings these electrons to a focus on the surface of the target. Each of these photoelectrons impinging on the mosaic surface of the target with an energy of about 500 electrons volts will liberate between 5 and 10 secondary electrons. These relatively low-velocity secondary electrons play much the same role in the generation of picture signals as do the photoelectrons liberated from the mosaic of the emitron by the direct action of light.

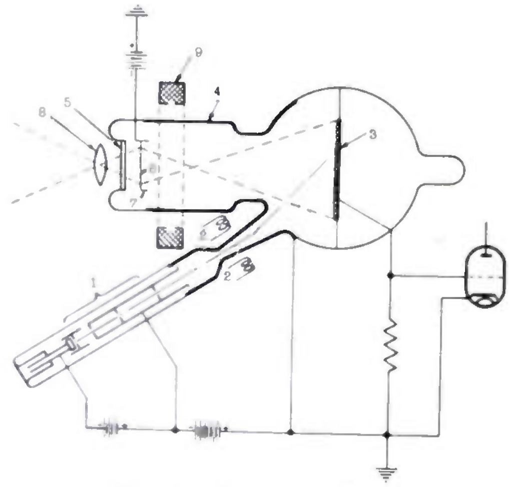


Fig. 10—The super-emitron.

Two significant advantages have been gained in this tube. First, the functions of photoelectric emission and insulating mosaic have been separated. The photoelectric sensitivity can be pushed up as high as possible without ruining the mosaic insulation. In fact, photosensitivity of 30 to 40 μa per lumen can be reached as compared with 10 μa per lumen for the emitron. Second, each photoelectron that impinges on the mosaic liberates at least 5 secondaries. Thus, for the same amount of light gathered by the objective lens the number of electrons liberated from the mosaic is between 15 and 20 times greater in the super-emitron than in the emitron. Since the average energy of these secondary electrons is about ten times that of photoelectrons, it is not to be expected that the mechanism of signal generation would be the same in both tubes. In fact, noticeable differences are observed. The super-emitron shows less evidence of the very short-term charge storage, which is so characteristic of the emitron, and more evidence of uniform charge storage during the whole frame period. This is shown by rather more blurring of moving objects in pictures transmitted by a super-emitron as compared with those transmitted by an emitron. The increase in efficiency of the super-emitron over the emitron does agree fairly well with the factor of 15 to 20 given above.

Another useful feature of the super-emitron is that the electron image can be magnified between the photocathode and the mosaic. Thus, we usually start with an optical image 1 $\times$0.8 inch and magnify it to cover a mosaic 5$\times$4 inches. Then for normal work a lens of 2-inch focal length and *f*/2 or better can be used. This lens has a light gathering capacity of about 25 per cent of the normal lens of the emitron camera, but it gives a much greater depth of focus. Also, telephoto lenses may be used conveniently and a series of lenses can be mounted on a rotating turret.

Since, of the total increase in efficiency of the super-emitron over the emitron by a factor of 15 to 20, a factor of 4 has been used in improving the optical conditions, we are left with a factor of between 4 and 5, as the increase in the sensitivity of the super-emitron camera over the emitron camera. That is, a super-emitron camera will give the same quality of picture at 40 to 50 foot-candles as an emitron camera will give at 200 foot-candles incident illumination.

A practical advantage of the super-emitron is due to the fact that the mosaic is not photosensitive as in the emitron, where the action of the electron beam slowly reduces the photosensitivity and so shortens the life of the tube. The scanning beam has no noticeable effect on the super-emitron mosaic and the photocathode has a very long life.

In focusing the electron image onto the mosaic, the whole image is rotated through an angle of 20° to 30° about the axis of the electron lens. This is easily compensated by rotating the whole tube by the same amount about the same axis. More important is the geometrical distortion introduced by aberrations in the electron lens, which result in peripheral points of the electron image being rotated slightly more than points nearer the center. By careful design it has been found possible to reduce this distortion to a tolerable magnitude.

Very efficient magnetic screening between the focusing field and the scanning fields is necessary. Otherwise, the former will distort the scanning raster and the latter will cause the electron image to oscillate slightly on the mosaic and definition will be impaired.

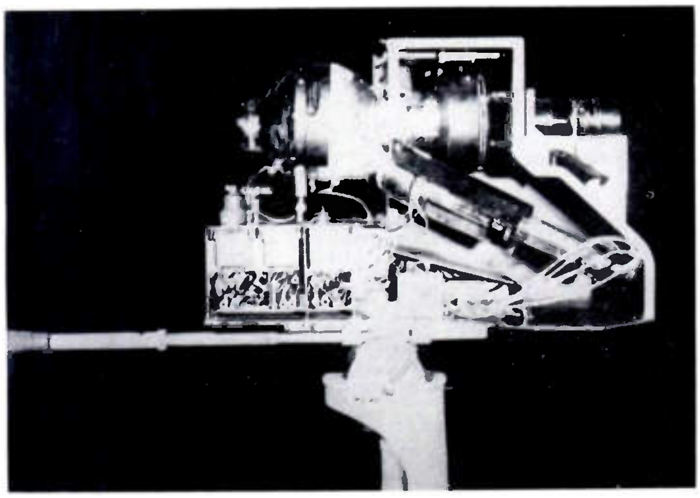


Fig. 11—Super-emitron camera without cover.

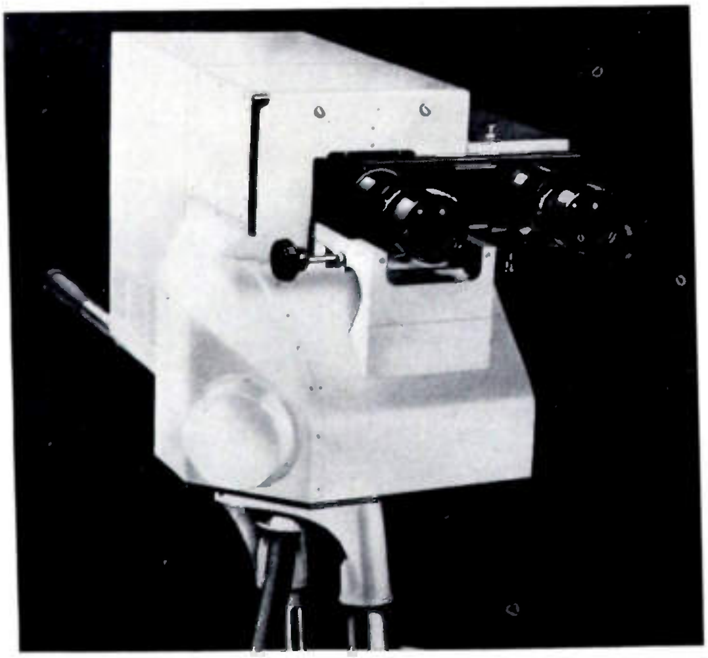


Fig. 12—Complete super-emitron camera.

The super-emitron suffers from the same inconveniences due to oblique scanning and generates the same type of spurious signal, shading signal, and streaking. Also, it does not produce a signal corresponding to true black.

A super-emitron camera with its cover removed is shown in Fig. 11. The magnetic focusing coil is easily visible and also the rotation of the tube to compensate for the rotation of the electron image. The complete super-emitron camera is shown in Fig. 12. Its greater sensitivity and adaptability to telephoto shots are of great value in outside broadcasts.

A small-size super-emitron tube has recently been developed which has a very similar performance to that of the standard tube. Its smaller size facilitates camera design, especially in regard to the lens system, since shorter focal length lenses can be used to give the same angle of view.

THE CPS EMITRON

We had realized as early as 1934 that it is the uncontrolled secondary electrons that are responsible for the spurious signals and low sensitivities of the emitron. In that year the late A. D. Blumlein and the author proposed a method which, it was hoped, would eliminate these troubles.[9] The fundamental principle of this scheme was to scan the mosaic with a beam of electrons which have so little energy when they reach its surface that they liberate substantially no secondary electrons. Under these conditions the mosaic must be driven negative until it reaches the potential of the cathode from which the electron beam originates. Since there are no secondary electrons, we argued, there can be no spurious signals and this has, in fact, proved to be true. Furthermore, since a strong field gradient is established in front of the mosaic which decelerates the approaching beam electrons, this must also serve to saturate away from the mosaic all photoelectrons that are liberated by the light of the image. Thus, there should be full storage of the photoelectric charges during the whole scanning cycle, leading to a great improvement in efficiency, which also proved to be true.

Considerable work had been done in our Laboratories before the war on tubes of this type and it had been realized by my colleague, Lubszynski, that the essential condition for stable operation of tubes in this manner is that the electron beam should fall almost normally onto the mosaic.[10] That is to say, the mosaic must be scanned orthogonally. However it is not because of this feature that the name "Orthicon" has been coined, but

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[9] A. D. Blumlein and J. D. McGee, British Patent No. 446,661; August 3, 1934.

[10] H. G. Lubszynski, British Patent No. 468,965; January 15, 1936.

because the signals generated are strictly proportional to the incident light.[11] We prefer to call this type of tube by the initials "CPS" of the earlier and even more fundamental principle employed of "cathode potential stabilization" of the mosaic.

Pre-war tubes of this type did not prove good enough to displace the emitron or super-emitron, but since the end of the war we have returned to this line of work and now have a CPS emitron with characteristics which I believe will interest you. An experimental camera using this tube was first used by the BBC to televise the royal wedding procession in November, 1947, and a properly engineered three-camera mobile unit was used for the first time in August, 1948, to televise those events of the Olympic Games that took place in the Empire Swimming Pool at Wembley, near London.

The tube and its associated driving coils is shown diagrammatically in Fig. 13. The cylindrical glass tube *1*, about 2½ inches in diameter, has a narrow neck at one end in which a simple electron gun *2* is mounted. This consists of a cathode held at earth potential, a modulator slightly negative, and a first anode held at about +300 volts. This latter electrode accelerates the electron beam and restricts it by an aperture at its center to a diameter of about 0.002 inch. The electrons are projected along the axis of the middle section of the tube which has a metal coating *3* on the walls extending from the first anode to slightly beyond a fine metal mesh *4*. This "wall anode" and the metal mesh are held at the same potential, about +200 volts. On the wall near the end of the large diameter section of the tube is a short cylindrical electrode *5* known as the decelerator. It is held at approximately earth potential. This end of the tube is closed with a flat polished window *6* and the target *7* is mounted on the flat inner surface of this window. This target is similar to the emitron target. It consists of a thin sheet of transparent dielectric which may be mica or glass between 0.002 and 0.004 inch thick. The surface of this dielectric facing towards the window is coated with a conducting but transparent metal layer, the signal plate. Contact is made to this layer through a metal seal at the edge of the window from a small metal plate *8* on the front surface of the window. Here a spring makes contact to the first stage of the head amplifier *9*.

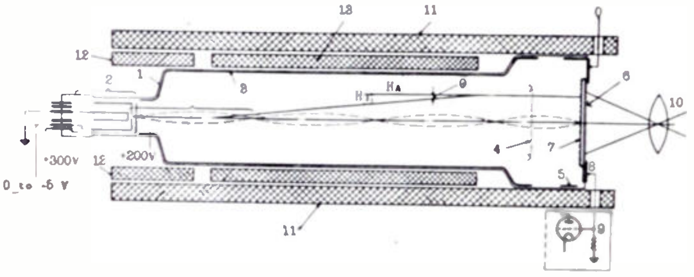

Fig. 13—Sectional diagram of CPS emitron.

The photosensitive mosaic is formed on that side of the dielectric which is scanned by the beam. It is 35 mm ×44 mm in standard tubes and 20 mm×25 mm in special purpose tubes. The mosaic is framed by an opaque metal border which is connected electrically to the signal plate. The optical image is focused onto this mosaic by the lens system *10* through the glass window, the transparent metal signal plate, and the dielectric. The light passes through the transparent elements of the mosaic itself to liberate photoelectrons from the side on which the electron beam falls. In the past, two reasons for the poor performance of tubes of this type were (1) the strong absorption of light in the signal plate and (2) the very poor photoelectric efficiency of the mosaics then used. The former loss has been reduced from 60 per cent to about 30 per cent while the efficiency of the mosaic has been increased from about 3 μa per lumen to between 12 and 15 μa per lumen. It had been known for ten years that the most efficient photo surface for visible light was a layer of antimony treated with cesium which is in fact used in the conducting photocathode of the super-emitron. It had proved impossible, however, to make a mosaic of this material by the usual techniques. This problem was solved when my colleague, Holman, produced extremely fine metal meshes having 1,000 meshes per inch or one million apertures per square inch and a shadow ratio of as low as 30 per cent. Having been provided with such a mesh we are able to use it as a stencil to make the mosaic. It is held in close contact with the surface of the dielectric, while the basic antimony elements of the mosaic are evaporated through it. This gives a mosaic having about 1,750 elements in the line direction and 1,400 in the frame direction, and a total for the whole mosaic of 2.5 million elements. Thus, there are over ten mosaic elements for each picture point of a 405-line picture. Fig. 14 shows on the left a microphotograph of a small piece of this metal mesh and on the right a similar photograph of the mosaic elements formed by using it as a stencil.

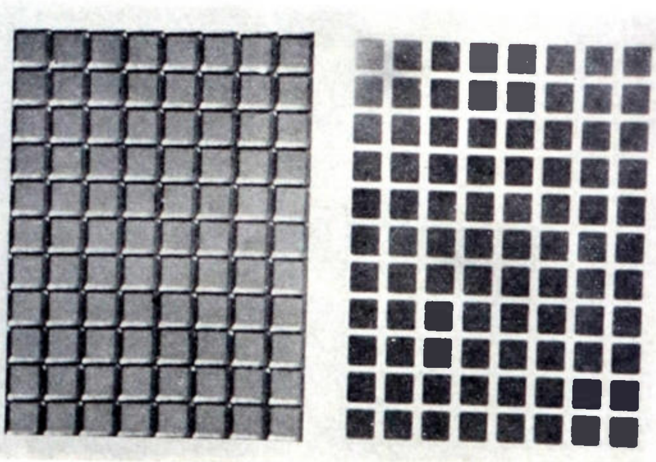

Fig. 14—Microphotographs of mesh (left) and mosaic (right) for the CPS emitron.

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[11] A. Rose and H. Iams, "Television pickup tubes using low-velocity electron-beam scanning," Proc. I.R.E., vol. 27, pp. 547–554; September, 1939.

It is interesting to look back to the first emitron mosaic made in our laboratories in 1932 which we made by evaporation through a mesh having 2,500 apertures per square inch.

I must now refer back to Fig. 13 to explain the focusing and scanning systems *4*. The tube is surrounded by a solenoid *11* which produces a fairly uniform axial magnetic field of about 50 gauss. As the electron beam is projected along the axis of the tube each electron traverses a spiral path and all return periodically to a focus in the well-known manner. The axial magnetic field and the electric accelerating fields are so adjusted as to bring the electrons to a focus on the surface of the mosaic. If the electron beam is not projected accurately parallel to the lines of force of the axial magnetic field, the whole beam will follow a spiral path and poor focus quality will result. It is impossible to guarantee mechanical accuracy of the electron gun, so two pairs of saddle coils are provided, one of which is shown at *12*, to give a steady transverse magnetic field where the beam leaves the gun. By suitable adjustment of the resultant field from these coils the beam can be deflected at the beginning of its path to bring it into alignment with the axial focusing field. Hence these coils have become known as the "alignment coils." On the long medium diameter section of the tube two more pairs of saddle coils *13* are mounted, producing transverse magnetic fields at right angles to one another. They are actuated with currents of sawtooth wave form, one at line frequency, the other at frame frequency. The maximum transverse field H_T produced by these coils is small (about 5 gauss) compared with the axial field H_A (50 gauss). So the instantaneous effect of this transverse field is to distort or bend the resultant field to make a small angle θ with the axis, where $\tan \theta = H_T/H_A$. Outside the region enclosed by these coils the magnetic lines of force again become parallel to the axis. Provided that the distortion in the magnetic lines of force is not too great, the spiral paths of the electrons will be so modified that they will follow the resultant lines of force through the deflecting fields, and after leaving the deflecting fields they continue to follow the lines of force which are now parallel to the tube axis. So the electron beam arrives normal to the mosaic.

If, having excluded all light from the mosaic and biased off the electron beam, we gradually apply volts to the electrodes of the tube, the mosaic will remain substantially at earth potential. If now we turn on the electron beam, the electrons will enter a retarding electric field as soon as they pass the mesh *4*. Between this mesh and the mosaic they will be slowed down to an energy of a few electron-volts and hence, on reaching the mosaic surface, each electron has a very small chance of liberating a secondary electron. It follows that the beam will drive the mosaic more and more negative in potential until it has reached the potential of the thermionic cathode from which the electrons originated. With the scanning fields operating, the whole mosaic is quickly established at cathode potential. Some corrections to this simple theory due to such effects as space charge, contact potential, and lateral energies of beam electrons, should be added here if time permitted, but they do not affect the final result much. When the beam electrons can no longer reach the mosaic they are reflected, accelerated towards the mesh *4*, retrace their outward paths very approximately, and are captured on the first anode.

If now a light image is formed on the mosaic, the photoelectrons liberated by the light are accelerated towards the mesh *4*. The electric field is sufficient to give full saturation of the photoelectric emission. Thus, a positive charge image is constantly being built up on the mosaic and at each passage of the beam the mosaic elements are driven back to cathode potential. The electron beam must be of sufficient intensity to discharge completely at each scan those mosaic elements exposed to the brightest parts of the image. Even in these peak white areas only about 10 per cent of the beam current is effective in discharging mosaic elements. In practice, a beam current of about 1 μa is used and the discharging current, which is the effective signal current, is about 0.1 μa. The residue of unused beam current returns to the electron gun. Obviously, in absolutely black areas of the image no beam electrons are accepted by the mosaic, all being returned to the first anode. That is to say, true black level in the picture signal corresponds to the state where the mosaic is receiving no electrons from the beam. It is obvious that this is the signal level produced during the return strokes of the scanning. Thus, there is provided automatically, when the beam is biased to zero, a signal which corresponds to true black, even though there may be no completely black area in the picture being transmitted. This gives a great advantage over the older tubes in that however extremely the light content of the picture may vary, the black level remains rock-steady. Consequently all the other gradations of picture brightness are truly reproduced.

A less desirable feature of this tube, which is common to all full-storage tubes, is the fact that movement of the image results in a blurring of the charge image and so of the reproduced picture. This loss of definition in individual pictures is equivalent to that observed in a photograph taken with 1/25th of a second exposure. However, just as in a cinematograph picture of moving objects, the eye is able to integrate the successive pictures to some extent and the total effect is not nearly so disturbing as would be expected from examining one single fame as a "still."

The capacity of the mosaic to the signal plate has an important bearing on this question of blurring of moving objects. If the dielectric is too thin, giving a capacity that is too great, the potential changes of the mosaic elements will be small. Hence a small proportion of the beam electrons will be able to reach the mosaic elements and the discharge will be inefficient. In these circumstances it is quite possible for charges to remain incompletely discharged for several frame periods which shows as serious blurring of moving objects and trailing after white objects. On the other hand, if the dielectric is too thick, the potentials of the mosaic elements may rise so high that the beam electrons have sufficient energy on reaching them to liberate almost one secondary electron per primary. Under these conditions the discharging is also inefficient with similar results. A compromise must be reached between these two extremes, and it is found possible to arrive at a fairly satisfactory result.

If a large amount of light falls on the mosaic it is possible for the photoemission to exceed the beam current. The mosaic elements will then rise rapidly in potential until the beam electrons can reach them with sufficient energy to liberate more than one secondary per primary. The beam then drives the mosaic elements still more positive, neighboring elements follow suit, and in a few seconds the mosaic will reach the potential of the wall anode. It then functions as a very bad emitron. This is corrected by first reducing the light to a reasonable level and then lowering the wall anode to earth for a few seconds and raising it again. Earlier tubes of this type were very unstable in this way and required very careful operation. This new tube is much more stable for the reason that the strips of highly insulating dielectric surface that separate the mosaic elements do not lose photoelectrons, and hence tend to remain at cathode potential. They therefore exert a stabilizing influence and may even reduce or stop altogether the escape of photoelectrons from those elements that have drifted dangerously positive.

In general, this control action of the "grid" of strips of insulating surface only comes into action when the potential of the mosaic element has risen well above the normal level for peak white. Over the normal working range of mosaic illumination it has little effect, and the charges built up are proportional to the image brightness; or the device has $\gamma = 1$. Since the normal cathode-ray tube increases the contrast of the signals which modulate its beam, this results in a picture with far too much contrast. The highlights are too bright and the darker tones are too dark. If the scene to be televised is very "flat" and lacking in contrast (imagine, for example, a grey winter's afternoon) this can be a good thing, since a much more interesting picture is presented to the viewer. But, on most occasions, a much more satisfactory result can be obtained by reducing the contrast of the picture. Thanks to the fact that this tube does give a definite black level to work on, my colleague White and his circuit experts have succeeded in doing this very satisfactorily. So now for the first time in the new CPS emitron cameras the contrast of the picture can be adjusted to give the most pleasing result.

You may have noticed a mesh *4* stretched across the path of the electron beam. This is to eliminate a white spot which may appear in the center of the picture which is produced by gas ions. The electrons pass along the tube, collide with molecules of residual gas, and produce positive ions. In the absence of this mesh *4*, these ions drift towards the mosaic under the influence of the electric field which retards electrons but accelerates positive ions. The convergence of this electric field also tends to concentrate these ions in a diffuse spot in the center of the mosaic. There they build up positive charges which are discharged by the beam, giving a white signal. The mesh *4* held at the same potential as the wall anode cuts off the electric field which collects these ions from some distance back along the tube axis. Any ions produced between the mesh and the mosaic are uniformly distributed over the mosaic area and hence do not appear as a disturbing signal in the picture. The mesh *4* need not be of such fine pitch as that used to make the mosaic, and it is placed at an antinode where the beam is spread to its greatest diameter and covers some hundreds of mesh apertures. Thus the modulation of the beam by interception of electrons on the mesh is negligible.

Fig. 15 is a picture of a CPS emitron tube. This shows the general appearance, but unfortunately the details of construction are not shown. Fig. 16 gives a half-front view of the CPS emitron camera. I can only briefly outline its design. The tube with its driving coils and head amplifier is mounted rigidly on a platform which can be racked back and forth on a very accurate "tramway," so the lenses are fixed and focusing is done by moving

Fig. 15—The CPS emitron tube.

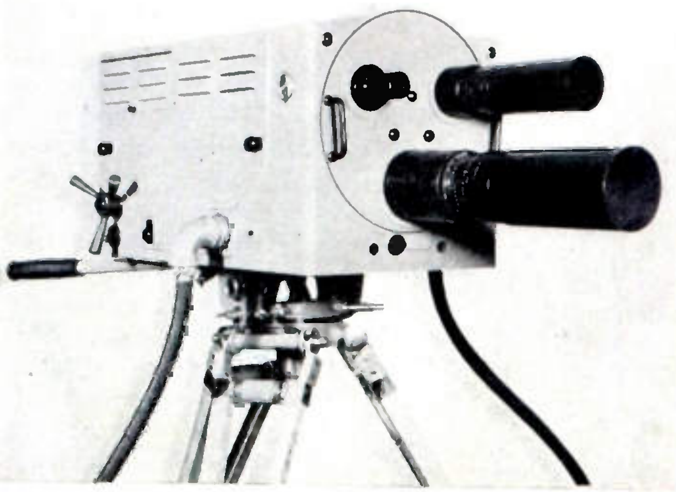

Fig. 16—CPS emitron camera (front view).

the tube. Two interchangeable lens turrets are provided, one having lenses 2.5-, 4-, and 6-inch focal length, all of maximum aperture $f/1.9$; the other having lenses of focal lengths 6-, 12-, and 17-inch and maximum aperture $f/4.5$.

Thus a range of angles of view from 5.5° to 40° is provided. A modified tube can now be provided with a smaller mosaic which gives an angle of view of 3 inches with the 17-inch telephoto lens. The iris diaphragms of the three lenses on each turret are ganged together and can be operated by a handle at the back of the camera. Thus all lenses are set at the same f number, so that when the turret is rotated the picture always comes up at the same signal strength. The turrets can be interchanged in a few minutes—if necessary, during a broadcast.

Fig. 17 shows a half-back view of the camera. At the side is the focusing handle and just below the viewing mask can be seen a large lever for rotating the turret and a smaller concentric lever for controlling the iris diaphragm. An electronic view finder is built into the camera and is observed by the operator through the mask which is seen on the back of the camera. This view finder is a small cathode-ray-tube television receiver on which the picture signals being generated by the camera are displayed. Thus it enables the operator to view-find and focus without an independent optical system.

Fig. 17—CPS emitron camera (back view).

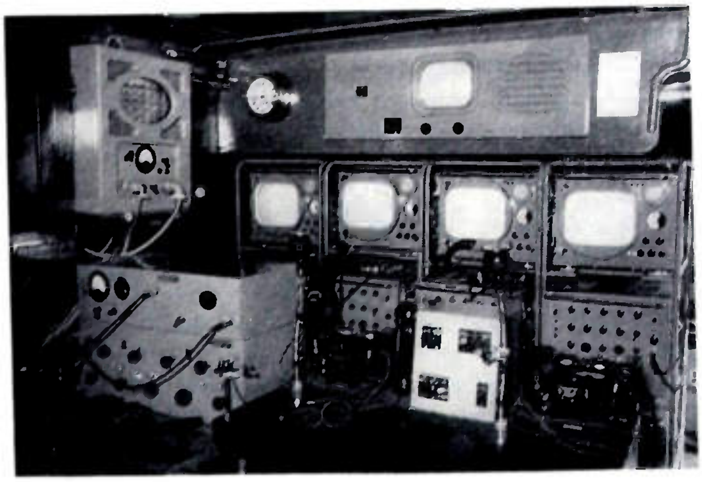

Fig. 18—Interior of CPS emitron control van.

In Fig. 18 the interior of the mobile control van is illustrated. Four similar units can be seen. Of the three on the right, each controls one of the three cameras that can be operated simultaneously. One operator controls each camera channel and is responsible for maintaining a satisfactory picture on his monitor. The fourth unit on the extreme left is the fading and mixing unit on which the picture from any camera can be selected for transmission. Above the four main units is the radio monitor where the picture as seen by the home viewer is displayed.

CONCLUSION

I would now like to summarize as far as I am able the relative merits of the television pickup tubes that have been described.

The quality of a television service depends on a large number of more or less independent features of the pickup tube that is used. The ten most important features are as follows:

1. Sensitivity and the closely related depth of focus.
2. Picture definition.
3. Picture geometry.
4. Prevalence of background blemishes.
5. Color response.
6. Spurious signals.
7. Stability to variations in light.
8. Picture tone gradation.
9. Effect on picture of movement.
10. Life of the tube.

I will now try to give an estimate of the performance of the three tubes, emitron, super-emitron, and CPS emitron in respect of each of these ten characteristics.

1. *Sensitivity*

The emitron has low sensitivity requiring about 200 foot-candles of illumination incident on the scene, and even then has very small depth of focus. The super-emitron has moderate sensitivity requiring about 50 foot-candles but has good depth of focus even with the lens at full aperture. The CPS emitron has still better sensitivity, requiring about 10 foot-candles, and also has good depth of focus at full aperture. Under reasonable illumination the lens can be stopped down and great depth of focus obtained. Even at much lower illumination a reasonable picture can be obtained, for though the signal-to-noise ratio becomes worse, the picture is not vitiated by uncontrollable and variable shading. It makes possible direct transmissions from locations where extra lighting is impracticable or impossible, e.g., theatres, public functions, and the like.

2. *Picture Definition*

All three tubes are very good for 405-line pictures, but the super-emitron is somewhat better than the emitron, and the CPS emitron is the best of the three. The limiting resolutions obtainable are as follows:

Emitron	800 lines
Super-emitron	1,000 lines
CPS emitron	1,250 lines.

The definition of the latter is more uniform over the whole picture than either of the other tubes.

3. *Picture Geometry*

The emitron is excellent, the super-emitron is the worst, being prone to spiral distortion, while the CPS emitron is about as good as the emitron.

4. *Background*

All tubes can be made with very clean backgrounds and there is little to choose between them in this respect, although extreme care is necessary in manufacture.

5. *Color Response*

The response of the emitron to colors is very close to the response of the human eye with a little more response to the red and infrared than is desirable. However, very light make-up is all that is necessary. The super-emitron and CPS emitron have the same response and are rather less sensitive to red than the eye. They have no response to the infrared. There is little to choose between the tubes on this score.

6. *Spurious Signals*

Strong and variable spurious signals are produced by the emitron and super-emitron which require continual correction. The bad effects due to them are seldom completely eliminated and they become serious at low light levels. The CPS emitron has no spurious signals that affect the viewer.

7. *Stability to Light Variations*

The emitron and super-emitron are completely stable to light, though to get the best results the amount of light falling on the photosurface must be adjusted. The CPS emitron is fairly stable and will cope with most light variations met with in practice. To get the best results, the amount of light falling on the mosaic must be kept within a range of 4 to 1 by adjustment of the iris diaphragm. Excessive light can destabilize the mosaic and put the tube out of operation for a few seconds.

8. *Picture Tone Gradation*

The emitron and super-emitron give low contrast pictures which are a good average compromise for most cases. However, their contrast is fixed and nothing can be done to alter it, since there is no fixed black level in the picture signal. For this reason also, the general level of a transmitted picture depends on the manual adjustment of an operator, who may not even see the actual scene being televised. It is therefore quite arbitrary. Furthermore, the uncorrected shading results in some fogging of the dark areas of the picture. The CPS emitron gives very true blacks and, whatever the distribution of light and shade in the picture, the reproduction is accurate. The gradation between black and white is excellent and is not vitiated by uncorrected shading. The signal output is proportional to light input which gives too high a contrast law for general use. However, since the black level is fixed, control can be applied successfully, and the contrast of the transmitted picture can be adjusted to optimum.

9. *Effect on Definition of Movement in the Picture*

Emitron pictures of normal movement remain very sharp, equivalent to an exposure time of 1/250 second, but very rapidly moving objects appear as a succession of separate images which sometimes leads to curious results. The super-emitron gives pictures which are not so sharp, some blurring being noticeable in cases of rapid movement equivalent to an exposure time of, say, 1/100 second. The CPS emitron gives pictures at best corresponding to the equivalent of 1/25 second exposure, and so some loss of definition of rapidly moving objects is noticeable.

10. *Life*

The emitron has a relatively short life of around 100 hours. The super-emitron has a long life—probably between 500 and 1,000 hours. The CPS Emitron has not yet had sufficient use to be certain on this point but some tubes have already done 200 to 300 hours of operation without any serious deterioration occurring, so I believe the prospect for a long life is good.

You will see from this brief summary that there is as yet no perfect television pickup tube. I leave it to you to give marks and to add up the score. In the end the question of what is, and what is not, a good television picture is to some extent subjective, and people are influenced by what they have become accustomed to. The only final test is a long series of broadcasts ranging over a wide variety of subjects. The virtues and vices of the emitron and super-emitron are well known. The CPS emitron has had, as yet, only one serious test—the Olympic events from the Empire Pool at Wembley. It is generally agreed, I think, that it survived this severe test with flying colors.

ACKNOWLEDGMENTS

I wish to thank Sir Ernest Fisk, I. Shoenberg and G. E. Condliffe, and the Directors of Electrical and Musical Industries, Ltd. Research Laboratories, Inc., for permission to present this paper to you; my colleagues of Electrical and Musical Industries, Ltd., who have done so much of the work, and last but not least my friend, J. Briton, who I am sure will have helped to make this cold meal considerably more palatable.

Who Is the True Inventor?*

I. E. MOUROMTSEFF†, FELLOW, IRE

> "Some men are born great,
> Some achieve greatness, and
> Some have greatness thrust upon them."
> Shakespeare ("Twelfth Night")

THE QUESTION, "who is the true inventor," is often asked about many modern inventions and, perhaps more often, about those cases which have already became past history. At first, this rather ideological question seems to be superfluous, if not ridiculous. Indeed, is it not the duty of the law, before a patent can be granted, to find out the whole truth about the invention? Or to restore the truth, if some error has crept into the original procedure? Large sums of money are sometimes spent on law suits in an effort to establish the "true" inventor.

And yet, one can hardly think of a single important invention for which the full credit would be unequivocally given by *vox populi*—and off the record, even by experts—to a single person, or to the same person. Moreover, the favored person is by far not always the official patentee. A good illustration of this apparently strange phenomenon is the case of radio.

Up to the most recent time, the average American and Western European, if asked, who invented radio, would unhesitatingly answer: "Of course, Guglielmo Marconi." However, to the confusion of many during the last few recent years, the idea was repeatedly advanced by the Soviet press and radio that the true inventor of radio was not Marconi but a Russian professor, Alexandre Popov.[1] The evidence given in support of this claim has apparently not been heeded by the Westerners, as to the great majority Marconi still remains the inventor of radio; yet some people begin to wonder if it was really so. This controversy was conspicuously reflected in the fact that the fiftieth anniversary of radio was solemnly observed in Moscow on May 7, 1945, while the Western world celebrated it by an International Marconi Radio Congress in the fall of 1947.

Those, who are acquainted with the history of scientific events preceding the advent of radio, know that Marconi and Popov are not the only names connected with it. Nearly a dozen other names of renown have been suggested by various authors, each one favoring his own hero as being the most important contributor in originating radio.

Such lack of consensus amounting to an ideological controversy is not new in the history of science and engineering. Thus, for example, in 1676, bitter arguments were exchanged between Leibnitz and Newton and their friends as to who was the true inventor of calculus, the basis of modern mathematics.[2] The arguments continued for many years and were accompanied by outright accusations of plagiarism. But even now the case remains unsettled.

About the middle of the nineteenth century, the discovery of the classical law of Conservation of Energy aroused a heated discussion among the supporters of James Prescott Joule, of Hermann Helmholtz and of Robert Mayer. Yet, even now, there is no full accord on the subject among scientific authors.[3]

Many more examples to the same point could be quoted from the history of science. It is enough to compare textbooks by different authors and books published in different countries in order to see a flagrant lack of agreement in evaluating the merits of various contributors to scientific advancement.

If, then, lack of harmonious judgment is a common occurrence in the domain of pure science, where untainted truth should be paramount, can one wonder that there is no general agreement in regard to those inventions which arouse the curiosity and admiration of the general public who do not know their true history. Steamboats, electromagnetic telegraph, telephone, electric light, movies, radio, radio broadcasting, television, radar, and others,—practically all inventions of the industrial age have a multiplicity of names attached to each of them. Many of them have the "true inventors" in several countries with memorial plaques adorning the walls of their birthplace. Some countries count two or more "true inventors" of the same device within their own boundaries. In fact, there is hardly a single invention with a single "true" inventor. What is the cause of this confusion?

The main fallacy of all such controversies, scientific or otherwise, undoubtedly lies in the attempt of the disputing parties to offer an *absolute* answer. Indeed, under the term "true" inventor or discoverer one tacitly understands a benefactor of humanity such that, had he not been born, the particular invention or discovery would have never seen daylight. Yet, the whole history of science, pure and applied, clearly shows that there are no such indispensable men. The very fact that several names are usually attached to many significant discoveries proves the incorrectness of such an assumption. Of course, in each case, the relative importance and merits of individual contributors may and usually do differ in value.

In an attempt to establish a criterion for ascertaining the true role of different contributors to human progress, we must make it clear to ourselves as to what factors are usually involved in making great inventions and discoveries. This subject is often touched upon in biographies of great men, in books on history and philosophy of science, etc., but usually no definite or uniformly general solution is given. First, one may ask a natural question: Why is it that from the multitude of students of science and engineering only an extremely few succeed in enriching humanity with really outstanding novel ideas and revolutionary inventions.

It seems that a satisfactory answer to this query can be found by postulating the following *working hypothesis*. Let us imagine, as Plato did, that all kinds of general ideas, when time is ripe, are found, so to say, "floating in the air." How they get there is immaterial to our present thesis. As general progress marches on, humanity reaches different levels of intellectual and spiritual maturity, at which it becomes capable of assimilating certain ideas heretofore unrevealed to it, or revealed and long forgotten. In this general evolutionary movement a few individuals, pioneers of the spirit, become attuned earlier than others to some of these ideas, more readily respond to them and perceive their manifestation in observable phenomena, while other people pass them by without noticing them. History shows that, in addition to some other necessary qualities, the *conditio sine qua non* (the condition without which the matter cannot be) for such an ability to perceive and to discover seems to be a spirit full of interest and devotion to the subject and the absence of prejudices. These people become channels through which humanity as a whole receives new ideas. They are those who are *born great*.

The postulated hypothesis is not more arbitrary than Newton's force of gravity *jumping over the space* separating material bodies;[4] or Huyghens' luminiferous ether;[5] or relativistic independence of velocity of light of the observer's motion. Similar to other practical hypotheses this one permits a convenient explanation of several observable facts about the subject—discoveries and inventions.

In the first place, it explains why many a discovery was often made by several individuals located in different quarters of the world and knowing nothing about each other's work. The postulated phenomenon is not unlike the reception of a broadcast

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* Decimal classification: R.015. Original manuscript received by the Institute, December 27, 1949.

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[1] A. S. Popov, 1859–1905. Professor of Physics in Navy Torpedo School, Cronstadt, 1884–1901, and in St. Petersbourg Electrical Institute (also, its elected President), 1901–1905.

[2] W. W. R. Ball, "A Short History of Mathematics," Macmillan and Co., New York, N. Y., pp. 362–367; 1893.

[3] I. E. Mouromtseff, "Of what value is history?" *Elec. Eng.*, vol. 68, pp. 945–948; November, 1949.

[4] Evidence is available that Newton himself viewed this idea absurd and incompatible with common sense.

[5] Huyghens' ether was endowed with contradictory properties of a perfect elastic solid and of a perfect fluid.

message by several independent, properly tuned receiving stations. Acceptance of this hypothesis immediately suggests the futility of looking for a single "true" inventor to the exclusion of all others. It should also suggest the fallacy, in many cases, of mutual accusation of plagiarism.

Indeed, even the bare fact of an earlier publication of a new idea by one of the discoverers, or an earlier patent application not always can be taken as the undisputable proof of the lack of authenticity with some other contributor. Even right now, with all modern means of exchange of information, very few individuals, if any, can claim that they are perfectly well informed about the recent achievements made in their fields in foreign countries. A lag of one or two, or even more years is quite common. Language barriers naturally augment the difficulties. As an interesting illustration, the case of Oliver Lodge can be given. As is known, this eminent British scientist was working in the same direction as Heinrich Hertz, that is, on the path to discovery of electromagnetic waves. In a public lecture in July, 1888, Oliver Lodge demonstrated to the audience standing electric waves on wires and in air, he thought, discovered by him. He was genuinely disappointed when, shortly after, he learned about the ingenious and more complete work of Hertz on the same subject, eight months before his lecture.[6]

The postulated hypothesis also makes comprehensible why in the achievements of pioneers—the original discoverers, explorers, great inventors—there always is a pronounced element of what may be called intuition or *inspiration*, or simply a mental jump into unknown and unexplored domains *not governed by laws of logic*. Logic comes into the picture later on after the discovery is made, when the discoverer (or the inventor) himself, or other persons begin to accumulate new facts and to arrange them in logical harmony with the previous knowledge into a new theory. One commonly calls this period of logical activity the *development* stage.

A new idea often comes to the discoverer like a flash of lightning, apparently unwarranted, often in the most inappropriate circumstances (e.g., Archimedes in the bath tub). At that moment, the discoverer may not even be in a position to prove the soundness of his idea, theoretically; he simply *knows* it is true; he *senses* it.

Many an author bears witness to such spontaneity and absence of logical assuredness in the truth of his discovery. Thus, it took Newton almost twenty years to prove in detail the soundness of his law of Universal Gravitation.[7] It took natural philosophers more than a hundred years to prove the correctness of Huyghens' inspired idea of undulatory theory of light. Robert Mayer in his diary plainly states that the idea of Conservation of Energy flashed through his head as lightning when he was performing a bleeding operation on a sailor at Java.[8]

The idea of inspirational discoveries versus logical development (Edison's "inspiration vs. perspiration") is strongly emphasized by the most outstanding French mathematician, Henri Poincaré, in his excellent essay, "Mathematical Creation."[9] It is even said that he often made public his mathematical discoveries without their logical proofs and had to supply these on his friends' insistence.[10]

One may remark that with different authors the term *inspiration* used in this article acquires different names, e.g., intuition, spontaneous revelation, subconscious mind, supersensory perception, hunch, accident, good luck, and what not. However, in every case it definitely means the same: a mental jump into the realm of new ideas, unaided by logic. The legal domain of logic is only the *past* with respect to the ever moving *present*.[11] In the distinctive role of inspiration versus logic lies also the assurance that no *inventing machine* can ever be designed to replace the human mind. Also, in defiance of the popular scientific illustration of the significance of probability, humanity will always need for creative work the Shakespeares, Goethes, and Tolstoys with their quills, not "the army of monkeys strumming on typewriters."[12]

The dominant role of inspiration in bringing about novel ideas explains the known fact that sometimes great contributions to science, pure and applied, were made by *amateurs*, men with no conventional schooling, or by men trained in an entirely different art. Enough to mention Benjamin Thomson of Boston (Count Rumford), Sir Humphry Davy, Faraday, Joule, Robert Mayer,[3] Pasteur, and pleiads of great inventors, such as Samuel Morse, Peter Cooper, Alexander Graham Bell, George Westinghouse, Thomas Edison, Marconi, etc. One cannot *teach* the art of great discoveries or great inventions.[13]

In this light, logic may even exert an impeding effect. Indeed, one may observe that a majority of revolutionary discoveries even by the greatest scientists were made preeminently in their prime, or at least during the first half of their manhood, when the ingrained habits of logical thinking were not yet restricting the flight of their inspiration. A shining example of this was Newton himself. Practically *all* his great ideas were conceived by him at the age of 23, when he retired to his farm, Woolsthorpe, during the great plague of 1665–1666.[14] The rest of his life he was busy with the logical elaboration of his ideas; often, he did it reluctantly, urged by his friends and admirers.[15]

The "odder" a novel idea, the greater may be its potential value. Thus, Faraday's revolutionary concept of "lines of force" was simply repulsive to the great mathematical physicists of his time. Many of them used all their mathematical skill to circumvent Faraday's explanation of electromagnetic induction as an effect of lines of force propagating through space.[16] It took another genius, James Clark Maxwell, to grasp and assimilate Faraday's ideas and to unfold them into a great logical system known as "Maxwell's theory" of electromagnetism. Again, when Marconi started his first experiments on wireless transmission across the ocean, his intent appeared odd, even ridiculous to those whose strictly logical minds could not conceive how radio waves—in *everything* but their length similar to light waves—could go around the earth's curvature. Therefore, the first news of Marconi's success (December 14, 1901) was met by many a good scientist with great scepticism.[17] Logical justification of this unpredicted phenomenon took almost a quarter of a century.

Admitting the fact of *illogical* inspiration by no means deprives logic of its great value in bringing discoveries and inventions down to earth and in interpreting them to mankind at large. Just the opposite, if not substantiated by a subsequent logical analysis and proper development, any novel idea, no matter how good, may be rendered futile; if unheeded, it may go back into oblivion. Such actually was the fate of the atomic theory of Lucretius. This also happened to the amazing ideas of Roger Bacon.[3] Again, lack of logical support postponed the advent of bacteriology and its medical implications by two centuries. Indeed, John Astruc, physician to Louis XIV, speaks with contempt and indignation of the "visionary imagination" of some of his contemporaries (he exposes their names) who believed that various contagious diseases were caused by "numerous schools of little, nimble, brisk invisible things of a very prolific nature which, once admitted (to the blood), increase and multiply in abundance . . . and occasion all symptoms of diseases."[18]

In order to emphasize the importance of logical development, we may conjecture, for

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[6] Oliver Lodge, "On the theory of lightning conductors," *Phil. Mag.*, vol. XXVI, pp. 217–230; August, 1888; especially the postscript.

[7] William Cecil Dampier Dampier-Whetham, "A History of Science," Macmillan Co., New York, N. Y., pp. 167–168; 1929.

[8] Wilhelm Ostwald, "Grosse Männer," *Akad Verlagsgesellschaft m.b.H.*, Third and Fourth Ed., p. 65; 1910.

[9] Poincaré says: "The ideas come to me with the same characteristics of brevity, suddenness and immediate certainty."

[10] W. W. R. Bell, "A Short History of Mathematics," Macmillan Co., pp. 536–537; 1893; Quotation from Poincaré's biography by Gaston Darboux, his fellow mathematician.

[11] "Heisenberg has recently pointed out that the principle of uncertainty implies that the past may be completely deduced from the present, but the future cannot." J. G. Crowther, "Men of Science," W. W. Norton and Co., Inc., New York, N. Y., p. 325; 1936.

[12] Emil Borel, "Le hasard," Librairie Félix Alcan, Paris, Second Ed., p. 164; 1914.
A. S. Eddington, "The Nature of the Physical World," Macmillan Co., New York, N.Y., p. 164; 1929.

[13] "*Rules of discovery.* The first rule of discovery is to have brains and good luck. The second rule of discovery is to sit tight and wait till you get a bright idea." (G. Polya, "How to Solve It?" Princeton University Press, p. 158; 1948.)

[14] In a memorandum in Newton's handwriting, preserved at the University of Cambridge, Newton speaks reminiscently of his past important work: ". . . All this was in the two plague years of 1665 and 1666, for in those days I was in the prime of my age for invention, and minded Mathematics and Philosophy more that any time since. . . ." (William Cecil Dampier Dampier-Whetham, "A History of Science," Macmillan Co., New York, N. Y., p. 165; 1929.)

[15] Wm. Cecil Dampier Dampier-Whetham, "A History of Science," The Macmillan Co., New York, N. Y., p. 176; 1929.

[16] James Clark Maxwell, "A Treatise on Electricity and Magnetism," Vol. II, Third Ed., Oxford Univ. Press, London: Humphry Milford, (First Ed.) 1872, (3rd Ed.), pp. 486–492; 1893.

[17] Oliver Lodge, "Talks about Radio," George H. Doran Co., New York, N. Y. p. 61; 1925.

[18] Howard W. Haggard, "Devils, Drugs and Doctors," Harper and Brothers, New York, N. Y., p. 247; 1929.

example, that without Maxwell's supreme mathematical logic, Faraday's inspirational ideas of gradual propagation of electromagnetic disturbances through space and time could have lapsed into oblivion, or their full development would have been postponed by some period of time.

The great logical contributors are those who *achieve greatness* through their endeavor, knowledge, and skill.

It may be remarked that inspiration is necessary not only for discovery of novel basic ideas but also is needed at various stages of their logical development. It is needed each time the trodden path of logic comes to a dead end in solving even minor problems. The apparently simple idea of drilling a hole may come through inspiration. Thus, rows of holes along the edges of a movie film conceived by LePrince in 1890,[19] actually saved movies from being a failure at their very inception.

The outlined hypothesis and its inferences, once accepted, surely can help one more easily to untangle the existing ambiguities, discrepancies, and outright errors in historical information as often given in textbooks on science, historical essays and in so-called historical movies. However, even with the most sincere desire to give full justice to various contributors to human progress, it often proves difficult to escape the influence of certain biases. One of them is the "legal" bias.

The legal viewpoint is of course involved through the patents pertaining to the basic idea under consideration. As already stated, the legal criteria of the true inventor do not always coincide with the common understanding of justice. To begin with, patent laws are different in different countries: this immediately renders the legal truth multifaced. Furthermore, from the legal viewpoint a bare idea—no matter how brilliant—cannot be subject to a patent; it should be embodied in a useful device. Thus, for example, the law would refuse to view Sir William Crookes as the originator of radio, although he was actually the first to go on record in 1892[20] with the idea of utilizing the then recently discovered electric waves for practical communication. On the other hand, if a "legal" inventor, inspired by some one else's idea, made it *serviceable* to mankind through his skill and ingenuity, one must certainly admit that he contributed to human progress more than he who had conceived a wondrous idea but "hid it under a bushel."[21,22] The relative merits of all persons involved should be judged in each individual case according to the fruits of their deeds, either directly or by inspiring other people. No general rule can be given for this.

In addition to the legal bias, the correct pronouncement about a certain invention or a certain discovery is not infrequently influenced by what one may call, "human relation" factors. These are: personal friendship, business connections, commercial and industrial interests of organizations to which the inventor belongs or belonged even in the remote past—all of these tend to exaggerate his merits unduly and to minimize the virtues of competitors. In evaluation of individual inventors and scientists, such one-sided publicity is to be discounted.

Undoubtedly, the most damaging to truth in general picture of human progress seem to be *national* prejudices. Because of these, young generations are educated in the habit of disregarding contribution to progress by other nationals and of greatly exaggerating the merits of their own landsmen.[3] Why, for example, is the well-known Boyle's law of isothermal expansion of gases not called Boyle-Mariotte's law? Mariotte, a Frenchman, discovered the same law quite independently of Boyle, although he did it several years later. Yet, it was Mariotte who first realized the prime importance of this law and immediately applied it to the calculation of the distribution of atmospheric pressure, while the same law, probably considered unimportant, was not even included in the first publication of Boyle's work.[23]

Again, why should the author of a textbook on physics in the chapter on light convey to the students the idea that velocity of light was *first* measured by Michelson? The names of the previous workers in this realm, Roemer, Bradley, Fizeau, and Foucault, are not even mentioned. Michelson contributed enough to science not to live on borrowed glory.

Such "why's" could be extended over many pages. There are indeed numberless cases in which *greatness is thrust* upon certain men while it is deliberately taken away from the others. It is not infrequently thrust upon those who themselves have already achieved greatness or were born great. (Edison and Einstein are good examples.) Why should not science be completely impartial in the matter of recognition of the merits of its great men without regard to their nationalities? Only mutual respect and justice may some day become a solid basis of *pax per cultura* (peace through culture), the dream of many scientists. Why should not Alexandre Popov and Guglielmo Marconi be given justice and honor, each according to his specific merits, without trying to disparage the other?

The factual material and official records of the dawn of the "wireless," historical literature, also the vast evidence produced by friends and admirers on both sides—all corroborate that Marconi and Popov arrived at the solution of practical wireless communication independently and, broadly speaking, simultaneously. Both inventors started their work in this direction during 1895. In that year Marconi developed his first apparatus and experimented at his father's farm near Bologna, Italy. Popov demonstrated to Russian scientists his thunderstorm recording station; this later on became his first radio receiving station. After further development, both inventors were able to demonstrate, in 1897, practical communication over distances of several miles, Marconi in England and Italy, Popov in Russia. It is hardly possible to establish with certainty on which particular day either inventor conceived his final idea of wireless communication. Is it, however, necessary? Is it necessary in a matter of such great cultural importance to apply horse-race logic in which a split-second may mean a total gain or a total loss? From all evidence it is clear that Russia learned about wireless and obtained it in practical form through the knowledge and ingenuity of the sedate scientist professor Popov; the Western world unquestionably received all that through the energy and ingenuity of the young Marconi, and his unabatable faith in the great future of radio.

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[19] Roger Burlington, "Inventors Behind the Inventions," Harcourt, Brace and Co., New York, N. Y., p. 147; 1947.

[20] Sir William Crookes, "Some possibilities of electricity," *Fortnightly Rev.*, vol. 57, pp. 172–181; February, 1892.

[21] A good example of this is represented by the discovery of "Edison Effect." After a feeble attempt to commercialize this phenomenon, Th. Edison completely forgot about it. It took O. W. Richardson, A. Wehnelt, J. A. Fleming, and Lee DeForest to investigate the phenomenon scientifically and to put it to work.[22]

[22] Clay H. Sharp, "Edison effect and its modern applications," *Proc. AIEE*, vol. 41, p. 69; 1922.

[23] Philip Lenard, "Great Men of Science," Macmillan Co., New York, N. Y., pp. 63–64; 1925.

A Source of Error in Radio Phase Measuring Systems*

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ASSOCIATE, IRE, AND A. TAIT†

Summary—**A source of error in radio navigation or distance-measuring systems involving the measurement of phase is discussed. Experimental evidence is given showing the effect of certain reradiating structures on the accuracy of such phase measurements.**

A SOURCE of error in radio-navigation and distance-measuring systems involving measurements of phase has been observed during the course of field calibration of an experimental system designed to measure the phase difference of radio-frequency fields as received at two widely spaced antennas. As far as is known, this type of error has not previously been reported in the literature.

Fig. 1 shows a simplified block diagram of the experimental system which consisted of two radio receivers spaced about 2.4 miles apart and a "local oscillator" and phase-measuring unit located approximately midway between the two receivers. The audio-frequency beat voltages obtained at the receivers by the combination of the signals from the local oscillator and a remote mobile transmitter were relayed to the phase measuring unit over vhf FM links. A measurement of the variation of the phase difference of the fields produced at the two receivers by the distant mobile transmitter could then be made conveniently by comparing the phase between the two audio-frequency beat signals as received over the vhf link at the central phase-measuring unit.

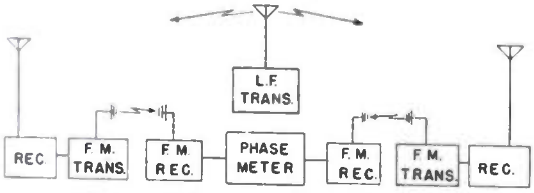

Fig. 1—Block diagram of low-frequency phase-measuring system.

Referring to Fig. 2, if the mobile transmitter is carried halfway around the system from P_1 to P_2 via any path, the total change in the indicated phase difference should be $4\pi D/\lambda$ radians, where D = distance between receivers, λ = wavelength in same units as D.

If the mobile transmitter travels completely around the system returning to position P_1, the total change in phase difference should be zero.

During several tests on a frequency of 300 kc, it was found that when the mobile transmitter traversed certain particular paths in the vicinity of re-radiating structures, such as power lines or telephone lines, the total indicated phase change differed from the expected result by $2\pi N$ radians, where N is an integer. Fig. 2(b) shows a plot of the phase change measured by the system as the mobile transmitter traversed the path shown in Fig. 2(a). For comparison, the calculated change in phase differential is also plotted for the path shown in Fig. 2(a). Inspection of these results shows that the measured differential phase change over the path P_1 to P_2 was approximately 2,390 degrees, while the calculated value was 2,784 degrees; furthermore, a net phase loss of 720 degrees was encountered in making a complete circuit around the system. It was noted during the tests that the sections along the path where rapid changes in phase were observed usually occurred in the neighborhood of power and telephone lines. Further tests resulted in the localizing of a relatively small region in which the mobile transmitter could be carried around a closed path, with the result that the indicated phase at the completion of the trip differed by 2π radians from the value indicated at the start.

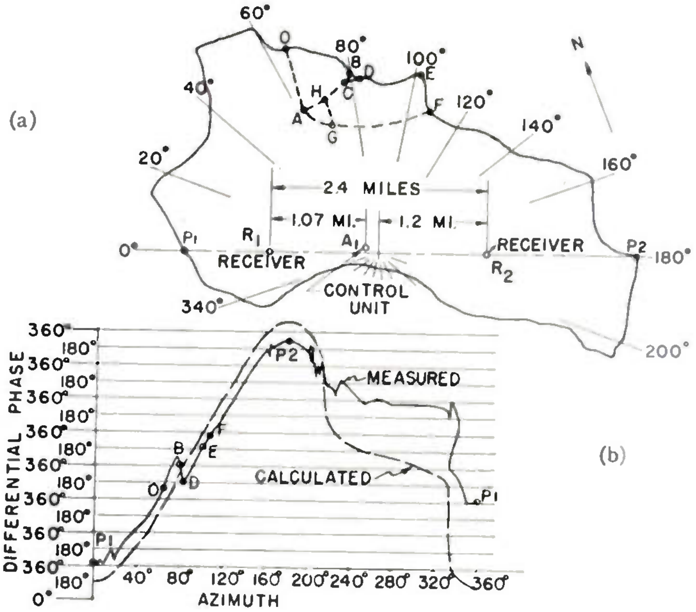

Fig. 2—(a) Plan of low-frequency phase-differential measuring system and path traversed by mobile transmitter around the system. (b) Phase-differential versus position of mobile transmitter.

Fig. 3 shows a section of the path shown in Fig. 2(a) and also a plot of the phase differential measured as the mobile transmitter was carried around this section (C-B-D). These results show a change of 360 degrees of phase differential for one traverse of the mobile transmitter around the closed path.

An analysis indicated that this type of error could occur under certain conditions when the field produced at

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* Decimal classification: R246. Original manuscript received by the Institute, June 30, 1949; revised manuscript received, November 18, 1949.

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one of the receivers by re-radiation from a parasite is of the same order of magnitude as that produced by the mobile transmitting antenna. In particular, it was found, both theoretically and experimentally, that the phase of the field at a distant receiver would change by

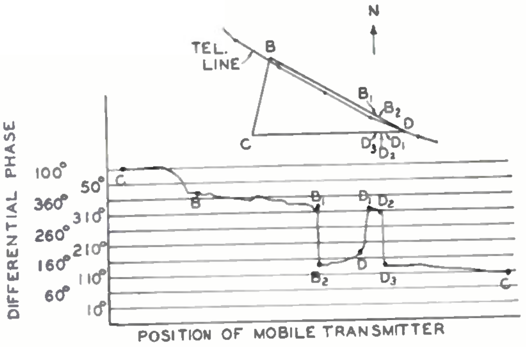

Fig. 3—Plot of phase-differential measured by system, versus position of mobile transmitter over path *C-B-D*.

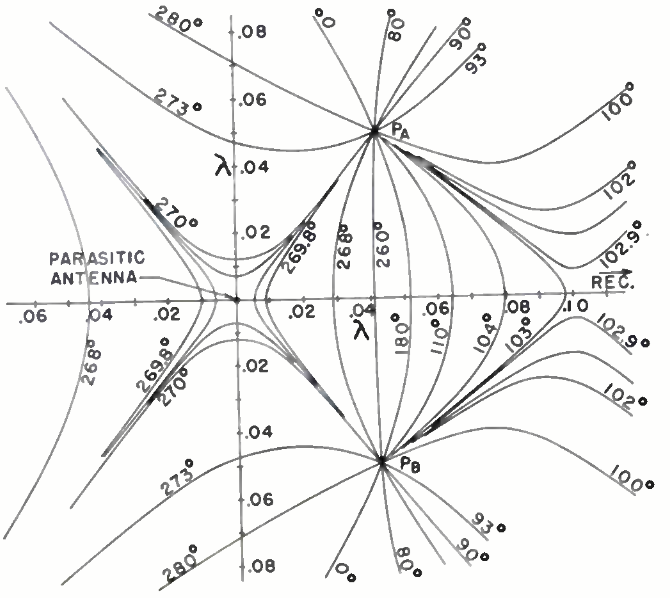

Fig. 4—Loci of driven-antenna position in the region of a parasite for constant phase of the resultant field at a distant receiver. *Note*: The driven antenna and the parasite antenna are each ¼ λ.

360 degrees if a mobile transmitter employing a ¼λ vertical antenna was carried around certain closed paths in the vicinity of a grounded ¼λ vertical parasite with its self-reactance approximately tuned out. Fig. 4 presents curves showing calculated variations in the phase of the field at a remote receiver produced by a mobile transmitting antenna and a tuned parasite at a fixed position relative to the receiver. In determining the effects of the tuned parasite, the computations of mutual impedance for the various spacings between the transmitting and parasite antennas were based on the assumption that infinitely thin conductors were employed.[1,2]

The calculated curves of Fig. 4 were based on the following derived equations:

$$\overline{E}_R = \left[|E_P| \angle \beta + |E_D| \angle \frac{360d}{\lambda} \cos\{\alpha - \Delta\} \right] \angle -\frac{360s}{\lambda} \tag{1}$$

where,

$\overline{E}$ = Resultant field intensity at the distant receiver, assuming the parasitic antenna to be fixed at the origin.

$|E_P|$ = Field intensity amplitude at distant receiver, produced by radiation from the parasitic antenna.

$|E_D|$ = Field intensity amplitude at distant receiver, produced by radiation from the mobile (driven) antenna.

$\beta = 180° + \theta_{DP} - \tau$.

θ_{DP} = Phase of mutual impedance between "driven" antenna and parasitic antenna,
$= \tan^{-1} X_{DP}/R_{DP}$.

τ = Phase of self-impedance of parasitic antenna, which is zero in this case, since tuned conditions are assumed for the parasite.

d/λ = Distance from parasitic antenna to driven antenna, wavelengths.

s/λ = Distance from parasitic antenna to distant receiver, wavelengths.

α = Angle between x axis and a line from the origin to the driven antenna.

Δ = Angle between the x axis and a line from the origin to the distant receiver. This angle was assumed to be zero for the calculations of curves in Fig. 4.

Equation (1) is of the form:

$$\overline{E}_R = |E_R| \angle \phi - \psi \tag{2}$$

where,

$$\psi = \frac{360s}{\lambda}$$

and

$$\phi = \tan^{-1} \left[\frac{|E_P| \sin\{180° + \theta_{DP} - \tau\} + |E_D| \sin\left\{\frac{360d}{\lambda} \cos(\alpha - \Delta)\right\}}{|E_P| \cos\{180° + \theta_{DP} - \tau\} + |E_D| \cos\left\{\frac{360d}{\lambda} \cos(\alpha - \Delta)\right\}} \right]. \tag{3}$$

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[1] G. H. Brown, "Directional antennas," Proc. I.R.E., vol. 25, pp. 78–145; January, 1937.

[2] C. Russell Cox, "Mutual impedance between vertical antennas of unequal heights," Proc. I.R.E., vol. 35, pp. 1367–1370; November, 1947.

For a given (fixed) position of the distant receiver, the angle ϕ is the phase of the resultant field intensity at the receiver.

From equation (3):

$$\alpha = \cos^{-1}\left[\frac{\phi - \sin^{-1}\left\{\frac{|E_P|}{|E_D|}\cos\phi\,(\tan\phi\cos\theta_{DP} - \sin\theta_{DP})\right\}}{\frac{360d}{\lambda}}\right]. \quad (4)$$

Since the parasitic antenna and the driven antenna were each $\lambda/4$ and tuned, it follows that:

$$\frac{|E_P|}{|E_D|} = \frac{|Z_{DP}|}{R_P} \quad (5)$$

where

$|Z_{DP}|$ = mutual impedance between the parasitic antenna and the driven antenna.

R_P = resistance of parasitic antenna.

Values of ϕ and d/λ were assumed and the corresponding values of α were calculated from (4), (5) and the work cited in footnote reference 2.

The set of curves shown in Fig. 4 indicate two points of singularity, P_A and P_B. When the transmitting antenna is at either of these singular points, the field intensity at the remote receiver is zero and the phase is indeterminate. If the transmitting antenna is carried around a closed path enclosing either one of these singular points, the resultant phase of the field at the distant receiver will change 360 degrees for each traverse of the closed path.

Furthermore, it was determined experimentally that other combinations of lengths of a mobile transmitter antenna and a re-radiating parasitic antenna resulted in a phase discrepancy of 360 degrees, when the mobile transmitter traversed particular closed paths in the vicinity of the parasite. It appears that this type of error would be encountered in certain types of radio navigation or radio distance-measuring systems involving measurements of radio-frequency phase if the mobile transmitter or receiver passed near structures, objects, or wire systems which act as efficient parasites.

A phase-error effect similar to that discussed above would be encountered in the case where the reflected signal from an airplane to a radio receiver in a ground-based vhf phase-measuring system became greater than the ground-wave signal to the receiver. In this case the reflected signal would take over or "capture" the phase, with the result that the indicated phase change could be very large and caused solely by the movement of the airplane. This spurious phase change could either add or subtract from normal phase reading, depending upon the movement of the airplane relative to the phase-measuring system. In general, the net spurious phase change produced by the movement of the airplane would not be zero and would depend on the difference in path lengths for the signal reflected from the aircraft at the aircraft positions where "capture" began and ended.

Further experimental and theoretical studies are under way for determining the nature of the conditions under which errors of this type are encountered, as well as methods of operation or instrumentation techniques for avoiding such errors in practical systems.

An Analysis of Some Anomalous Properties of Equiphase Contours*

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Summary—If one thinks of the voltage of a radio signal as a function of the location of some target point, then the phase of this voltage is also such a function. One might imagine that since the voltage exhibits no peculiar properties, neither does the phase; but this is not true. There are many cases for which the phase becomes a multivalued function. This fact is very important in the theory of the phase measurement type of radio surveying or navigational systems; for the phase difference between two points depends not only on the location of the points, but also on the path over which the target is carried.

INTRODUCTION

ONE OF THE more popular types of radio surveying systems is the phase measurement type in which the phase of a radio-frequency signal is used to determine the location of a "target point." This

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* Decimal classification: R246. Original manuscript received by the Institute, August 9, 1949; revised manuscript received, March 8, 1950.

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can be done in any of a great many ways; but essentially the technique consists of erecting an antenna at the target point and another at some point whose location is known, and then transmitting a continuous wave between the two. If the phase of the received voltage can be measured, an equiphase contour will be determined along which the target point must lie. With two such arrangements, two contours will be determined, and the target point must be located at their intersection.

Unfortunately, measuring the phase of a voltage is not a simple matter, largely because the term "phase" is, as it stands, ambiguous. Phase must always be measured relative to the phase of some reference voltage of the circuit, and even then it can be defined only to within an integral number of whole cycles.

In order to resolve these two difficulties, it has been found necessary to change to some extent the elementary concept we have just outlined. The first difficulty is generally disposed of by speaking only in terms of the phase difference between voltages at the two ends of some established base line; while the second is resolved if we are content to carry the target antenna from some point whose location is known to the point whose location we wish to find, meanwhile making sure that we count out the number of whole cycles over which the phase changes.

Whelpton and Redgment[1] have described a large number of such systems that were developed during the war. This number is so great and the applications so wide that there seems to be little room for doubt that the concept of using phase measurements as a position indicator is fundamentally sound.

Recently, however, Bateman, Florman, and Tait[2] have reported an experiment which raises some new questions as to this soundness. Using an experimental phase measuring system, they found that by carrying the target antenna between two points by several different paths they got, contrary to all expectations, different phase measurements. These differences seemed always to be equal to an integral number of whole cycles.

According to these authors, this phenomenon is caused by the presence of parasitic reradiators. They found that such parasites would often set up interference patterns which contained points where the received voltage vanished. Encircling such a point in a closed path gave a total phase change not of zero, but of precisely one whole cycle.

And as a matter of fact, using their hypothesis to guide us, it is quite easy to construct examples in which this anomaly appears. Bateman, Florman, and Tait in their paper have given one such example. Sandeman[3] has put the principle to practical use by setting up spiral-phase fields for azimuth determination.

In Fig. 1 is still another example. Here we have drawn a set of equiphase contours which appear at the ground when there is a dipole at R and a second at S, four wavelengths away. The dipole at S carries half the current that the one at R does, and the two are 180° out of phase. Notice how at the five points A, B, C, D, and E the contours seem to fold back on each other. If we measure the phase difference between, say, points R and P, the result depends very much on how the path we take interlaces these five points. Three such paths are shown in Fig. 1 which give this phase difference variously as 3,060°, 2,340°, and 2,700°.

It thus appears that the phase of a radio signal is not as has been previously thought, a single-valued function of position, but must be considered multiple-valued. This paper is written as an introduction to this multiple-valuedness. In it will be shown when and why equiphase contours become ambiguous.

The Analysis in Two Dimensions

First of all, let us suppose that we are surveying on some plane surface. A target antenna is placed at a point P which we can describe by the rectangular co-ordinates (x, y). Then at the receiver terminals, whether they are attached to the target antenna or to some stationary antenna, we shall find a voltage $E(x, y)$ which will be a function of the point P.

It is the phase of this voltage that we must measure. Therefore, if we separate $E(x, y)$ into its real and imaginary parts,

$$E(x, y) = u(x, y) + iv(x, y), \tag{1}$$

the phase $\phi(x, y)$ will be defined as

$$\phi(x, y) = \tan^{-1}\frac{v(x, y)}{u(x, y)}\,. \tag{2}$$

We cannot, as we have already seen, measure this phase directly. But if we proceed from the point (x, y) to the point $(x+dx, y+dy)$ an infinitesimal distance away, we can measure the resultant change in phase,

$$\begin{aligned} d\phi &= \frac{\partial\phi}{\partial x}dx + \frac{\partial\phi}{\partial y}dy \\ &= \frac{u\dfrac{\partial v}{\partial x} - v\dfrac{\partial u}{\partial x}}{u^2+v^2}dx + \frac{u\dfrac{\partial v}{\partial y} - v\dfrac{\partial u}{\partial y}}{u^2+v^2}dy. \end{aligned} \tag{3}$$

Continuing along some curve C, our phase measuring device will add together all the infinitesimal changes of phase, and the net phase change will be

$$\Phi = \int_C d\phi(x, y). \tag{4}$$

Now if $d\phi$ is a continuous differential throughout the entire plane, then, of course, this integral is equal sim-

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[1] R. V. Whelpton and P. G. Redgment, "The development of C. W. radio navigation aids, with particular reference to long-range operation," *Jour. IEE*, vol. 24, pt. III A, pp. 244–254; 1947.

[2] Ross Bateman, E. F. Florman, and A. Tait, "A source of error in radio phase measuring systems," Proc. I.R.E., this issue.

[3] E. K. Sandeman, "Spiral-phase fields," *Wireless Eng.*, vol. 26, pp. 96–105; March, 1949.

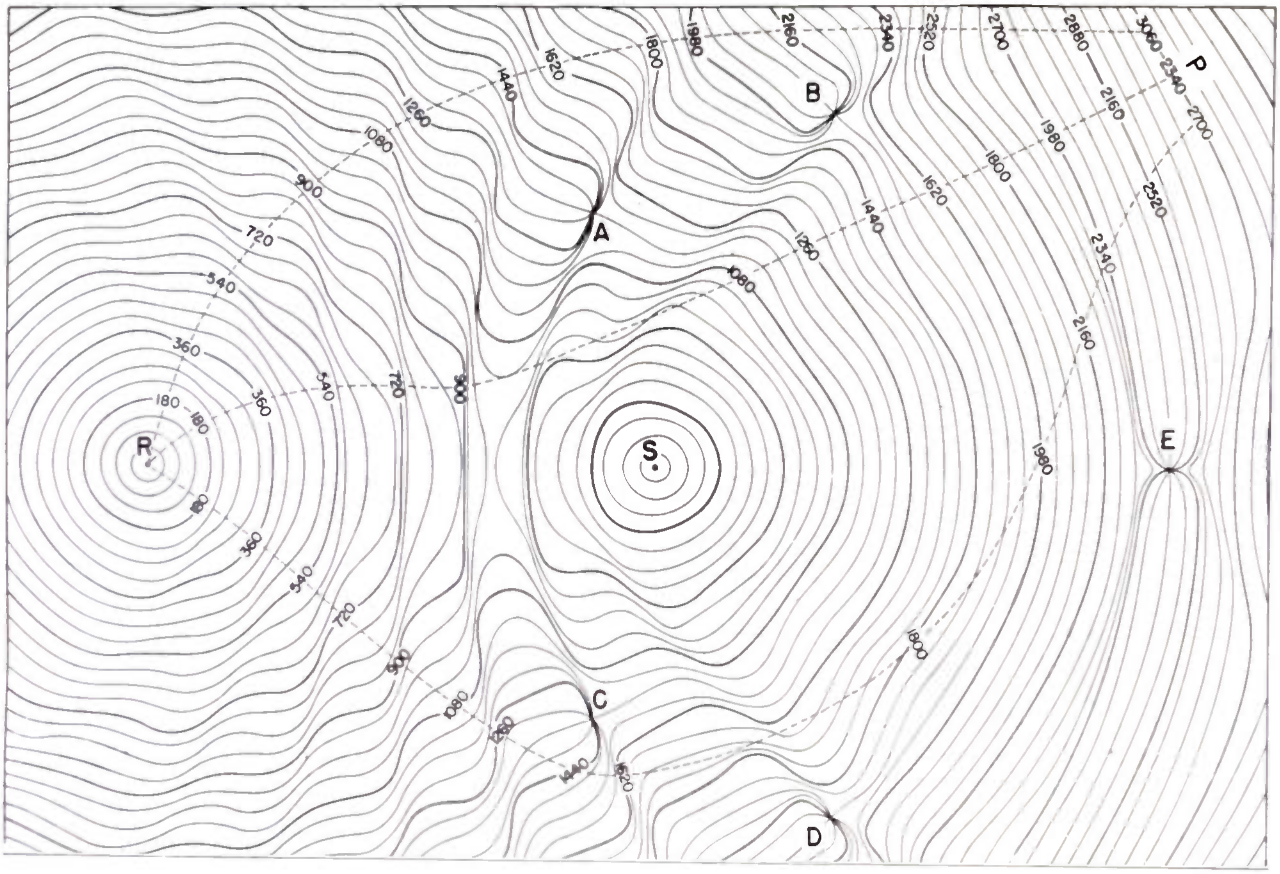

Fig. 1—Equiphase contours with a contour interval of 45° for an antenna system consisting of a dipole at R and a second dipole at S four wavelengths away. This latter dipole carries half as much current as the former, and the two are 180° out of phase.

ply to the difference of the values of ϕ at the two end-points of the curve C, and we run into no difficulties at all. However, by examining (3) we see that $d\phi$ will not in general be a continuous differential. It will have singular points wherever both u and v vanish. In the presence of such points, therefore, we can expect anomalies.

Although the existence of singularities somewhere in the plane means that we can no longer evaluate immediately the integral of (4), it is still possible to make several simplifying statements. First, it is shown in many of the text-books[4] that if the curve C is closed and $d\phi$ is continuous at all points within and on C, then the integral vanishes. From this it can in turn be shown that if K is some simply connected region within which $d\phi$ is always continuous, then the integral in (4) is independent of the path taken provided only that this path together with its end points lies wholly within the region K.

Thus just as in the integration of an analytic function of a complex variable, we can determine what effect the path has on the net phase change if only we know what values the integral assumes when the path is a small closed loop around each singularity.

Suppose, then, that at some point Q both u and v vanish. We shall transfer the origin of our co-ordinate system to the point Q. Then from Taylor's theorem we know that at any point in the neighborhood of Q

$$\begin{aligned} u &= u_x x + u_y y + O(r^2) \\ v &= v_x x + v_y y + O(r^2), \end{aligned} \tag{5}$$

where the subscripts indicate partial derivatives evaluated at Q, and r is the radial distance $(x^2+y^2)^{1/2}$ from Q.

Consider the co-ordinate transformation

$$\begin{aligned} x &= (u_y^2 + v_y^2)^{1/2}\rho \cos \theta \\ y &= (u_x^2 + v_x^2)^{1/2}\rho \sin \theta. \end{aligned} \tag{6}$$

The curves of constant ρ are evidently concentric ellipses encircling the origin Q. Since it makes no difference what particular form we use for the path C so long as this path encloses the singularity at Q and no other singularities, we shall define C to be one of these ellipses. Substituting (3), (5), and (6) into (4), we find the total phase change

$$\Phi = \frac{u_x v_y - v_x u_y}{(u_x^2 + v_x^2)^{1/2}(u_y^2 + v_y^2)^{1/2}} \cdot \int_{-\pi}^{\pi} \frac{[1 + O(\rho)]d\theta}{1 + \dfrac{(u_x u_y - v_x v_y)\sin 2\theta}{(u_x^2 + v_x^2)^{1/2}(u_y^2 + v_y^2)^{1/2}}}. \tag{7}$$

Note that if $u_x v_y - v_x u_y = 0$ this expression vanishes; that is, we still get zero phase change. To evaluate Φ in the

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[4] W. F. Osgood, "Advanced Calculus," Macmillan and Co., New York, N. Y., pp. 220–231; 1929.

case where $u_x v_y - v_x u_y \neq 0$ we remark that it must be independent of ρ. This means that the term in (7) which contains the function $O(\rho)$ must be zero when integrated. For otherwise we could by taking ρ small enough always effect an impossible variation in Φ. Discarding this term and using the formula

$$\int \frac{d\theta}{1 + k \sin \theta} = \frac{2}{(1 - k^2)^{1/2}} \tan^{-1} \left[\frac{\tan \theta/2 + k}{(1 - k^2)^{1/2}} \right] \qquad (8)$$

we can integrate (7), getting finally

$$\Phi = 2\pi \frac{u_x v_y - v_x u_y}{[(u_x v_y - v_x u_y)^2]^{1/2}} = \pm 2\pi, \qquad (9)$$

where the sign to be taken is the same as that of the quantity $u_x v_y - v_x u_y$.

In other words, we have shown that if we encircle an isolated zero of the voltage $E = u + iv$, we shall find a net phase change of $+2\pi$, -2π or 0 when the quantity $u_x v_y - v_x u_y$ is positive, negative, or zero, respectively.

This quantity $u_x v_y - v_x u_y$ which plays such an important part in the preceding analysis, is nothing more than the Jacobian

$$J = \begin{vmatrix} \dfrac{\partial u(Q)}{\partial x} & \dfrac{\partial u(Q)}{\partial y} \\ \dfrac{\partial v(Q)}{\partial x} & \dfrac{\partial v(Q)}{\partial y} \end{vmatrix}, \qquad (10)$$

evaluated at the point Q. If we write E^* for the complex conjugate of E, it is not difficult to see that J may be expressed directly in terms of E by the relation

$$J = Im \left\{ \frac{\partial E^*}{\partial x} \frac{\partial E}{\partial y} \right\}. \qquad (11)$$

Let us see how this may be applied to the example in Fig. 1. We are given the voltage

$$E(P) = \frac{e^{i2\pi r_1/\lambda}}{r_1} - \frac{e^{i2\pi r_2/\lambda}}{2r_2} \qquad (12)$$

where r_1 is the radial distance PR and r_2 the distance PS. At the five points A, B, C, D, and E, defined by the relations

$$\begin{aligned} r_1 &= 2n\lambda \\ r_2 &= n\lambda \qquad n = 2, 3, 4 \end{aligned} \qquad (13)$$

the voltage will be zero. If we place the origin of our co-ordinate system at R with the x axis directed along the line RS, a straightforward differentiation at these five points will give

$$J = Im \left\{ \frac{\partial E^*}{\partial x} \frac{\partial E}{\partial y} \right\} = \frac{-y\pi}{2n^6\lambda^5} \qquad n = 2, 3, 4. \qquad (14)$$

At the points A and B, y is positive; (14) shows that therefore J is negative. As we circle either of these points in a positive, counterclockwise direction, we should find that the phase has decreased by one whole cycle. Fig. 1 shows that this is indeed true.

Meanwhile at the points C and D where y is negative, J will be positive. Circling either of these points will increase the phase by one whole cycle.

But at E we have a different sort of result. For this time y is zero, so that according to (14) J is also zero. As we have seen, this means that the integral (7) vanishes and there is no net phase change. This, too, is borne out by Fig. 1.

THE ANALYSIS IN THREE DIMENSIONS

Perhaps this whole question of equiphase contours is better pictured in a full three dimensions. To do this it is merely necessary to define all our quantities as functions of the point $P(x, y, z)$. Thus we define the voltage

$$E(x, y, z) = u(x, y, z) + iv(x, y, z) \qquad (15)$$

and the phase

$$\phi(x, y, z) = \tan^{-1} \frac{v(x, y, z)}{u(x, y, z)}. \qquad (16)$$

Then if C is a three-dimensional curve, we have for the total phase change

$$\Phi = \int_C \frac{\partial \phi}{\partial x} dx + \frac{\partial \phi}{\partial y} dy + \frac{\partial \phi}{\partial z} dz = \int_C \nabla\phi \cdot \boldsymbol{ds} \qquad (17)$$

where $\boldsymbol{ds}$ is the incremental vector directed along C.

Suppose that C is a closed curve bounding some surface S; and suppose that throughout S, $\nabla\phi$ together with its derivatives is a continuous vector field. Then we can transform (17) by means of Stokes' theorem, finding, since the curl of a gradient is always zero, that

$$\Phi = \int_S \nabla \times (\nabla\phi) \cdot \boldsymbol{n} da = 0 \qquad (18)$$

where N is the unit vector normal to the positive side (which is determined by the direction of $\boldsymbol{ds}$ on the boundary C) of the element of area da.

On the other hand, if there is some point on S for which E vanishes, then $\nabla\phi$ is not continuous. And, in fact, by reasoning similar to that we have already used for a plane surface, it is easily shown that

$$\Phi = \pm 2\pi \qquad (19)$$

where now the sign is determined by the sign of the quantity

$$\boldsymbol{n} \cdot \nabla u \times \nabla v = Im\{\boldsymbol{n} \cdot \nabla E^* \times \nabla E\}. \qquad (20)$$

It must be remembered that the gradients are to be evaluated at the point where E vanishes, while the vector $\boldsymbol{n}$ is determined by the usual right-hand rule as applied to the direction in which the target antenna is carried along the curve C.

Let us examine now the analytic behavior of the set of points for which E is zero. At such a point we know that the two equations

$$\begin{aligned} u(x, y, z) &= 0 \\ v(x, y, z) &= 0 \end{aligned} \tag{21}$$

must be satisfied. But we may also look upon these equations as a set of two equations in the three unknowns x, y, and z. Such a set will define a curve Γ, if and only if the Jacobian matrix

$$\left\| \begin{matrix} \dfrac{\partial u}{\partial x} & \dfrac{\partial u}{\partial y} & \dfrac{\partial u}{\partial z} \\ \dfrac{\partial v}{\partial x} & \dfrac{\partial v}{\partial y} & \dfrac{\partial v}{\partial z} \end{matrix} \right\|$$

is of rank two. Saying that this matrix is of rank two is, moreover, equivalent to saying that

$$\nabla u \times \nabla v \neq 0. \tag{22}$$

Furthermore, this is precisely the restriction that we must place on (19) and (20).

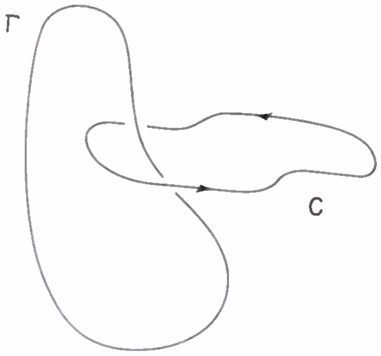

Fig. 2—An example of a path of integration C interlinking a curve Γ of zero voltage. The net phase change along C will be one whole cycle.

We can interpret these results in the following way. The net phase change around a closed curve C is one whole cycle, if and only if there exists within this curve a curve Γ along which the voltage vanishes. This can only happen if the curves C and Γ are related as in Fig. 2. If C, instead of interlinking Γ, goes completely around this curve then we shall no longer find a net phase change.

CONCLUSION

Returning to Fig. 1, we recall that the points A and C are determined by the relations

$$\begin{aligned} r_1 &= 4\lambda \\ r_2 &= 2\lambda. \end{aligned}$$

Actually, of course, these two relations in three dimensions determine an entire circle. The points A and C are only intersections of this circle with the ground plane. If the circle is interlinked with any closed curve, whether it be on the ground or not, we will observe a net phase change of 2π radians.

Similarly the points B and D of Fig. 1 both belong to a second circle. Interlinking this circle will also produce a net phase change. But the point E represents a degenerate case. For here a closed curve has been collapsed down to the single point E. Since we can no longer interlink this degenerate curve, it is not possible to obtain a net phase change.

Another type of degeneracy is present in the case of two dipoles of equal amplitude but opposite phase. In this case, the voltage E vanishes at all points of the plane of perpendicular bisectors to the line joining the two dipoles. Still, since (21) will not define a curve but only a plane, the vector $\nabla u \times \nabla v$ is zero, and the phase remains unambiguously defined.

Finally, there is one case of zero voltage which we have purposely avoided up to now. This is the case of a single vertical dipole; for the field directly above a dipole is zero. Now this set of zeros defines a curve; and we have said that circling any curve of zero voltage will produce a net phase change. Obviously, in this case this is not so. The answer to the paradox lies in the fact that along this line the gradient ∇E does not exist. In short, we cannot write down Taylor's series in (5), and all our conclusions from that point on are invalid. We must, instead, consider this as a special case which has no effect upon our phase measuring device.

And so we have concluded an examination of what seems to be all forms of the voltage E which will arise in practice. We have seen that, except for certain special cases, the existence of points where the voltage vanishes implies a multiple-valued phase. If we include within our list of degenerate cases curves of the type considered in the last paragraph, we can state this fact more explicitly: A phase measuring device will provide results which are independent of the path over which it is carried, if and only if there do not exist nondegenerate curves on which the received voltage vanishes.

A Microwave Propagation Test*

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Summary—A description is given of a microwave propagation test which was conducted over a period of a year with simultaneous transmission on wavelengths of 16.2, 7.2, 4.7, and 3.1 cm over an unobstructed 42-mile overland path. Comparative charts depict variations in daily fading range, illustrate diurnal and seasonal characteristics of fading, and reveal the marked disparagement between winter and summer fading. Curves are presented showing relative field-strength distribution for both winter and summer periods, and also the distribution of hourly minima. These curves may be useful in considerations bearing on continuity of service that may be expected with relation to wavelength and to time of day, winter or summer.

INTRODUCTION

THE OBJECT of this paper is to describe a microwave propagation test, recently completed, and to present a summary of the results obtained.

The term microwave is generally considered to designate frequencies of 1,000 Mc and higher (wavelengths of 300 cm and shorter).

DISCUSSION

Some of the more important advantages which obtain for radio communication systems operated at microwave frequencies are virtual freedom from external noise created by electrical and magnetic disturbances, considerable power gain from relatively small antenna systems, highly directional radiation at low elevation angles, and complete absence of ionospheric reflections.

The successful application of microwave techniques to communication systems requiring a high degree of circuit continuity and stability involves a reasonable knowledge and understanding of the behavior of microwave signals under the influence of varying atmospheric conditions, especially under unfavorable conditions, and particularly with respect to their relative effects at different wavelengths. The microwave propagation test described in this paper was undertaken for the purpose of investigating such behavior.

To insure that an adequate volume of suitable data would be obtained, four widely separated frequencies were selected for the test and the test itself was continued for a period of well over a year, beginning in December, 1946. No attempt was made to provide an elaborate system for recording meteorological data, since the primary purpose of the test was to study the effects of, rather than their relation to, variable weather conditions. It was thought, however, that a qualitative analysis of a general nature could be made using the United States Weather Bureau weather map and data which are published in the New York Times newspaper.

Observations were made on unmodulated carrier signals transmitted from a specially constructed tower at Neshanic, N. J., and received atop the Western Union building at 60 Hudson Street, New York, N. Y., a distance of 41.7 miles, entirely over land. The intervening terrain is characterized by a series of hills, as indicated by the profile in Fig. 1, most of which are densely wooded. The first Fresnel zone for 3.1 cm and for 16.2 cm, representing the shortest and longest wavelength

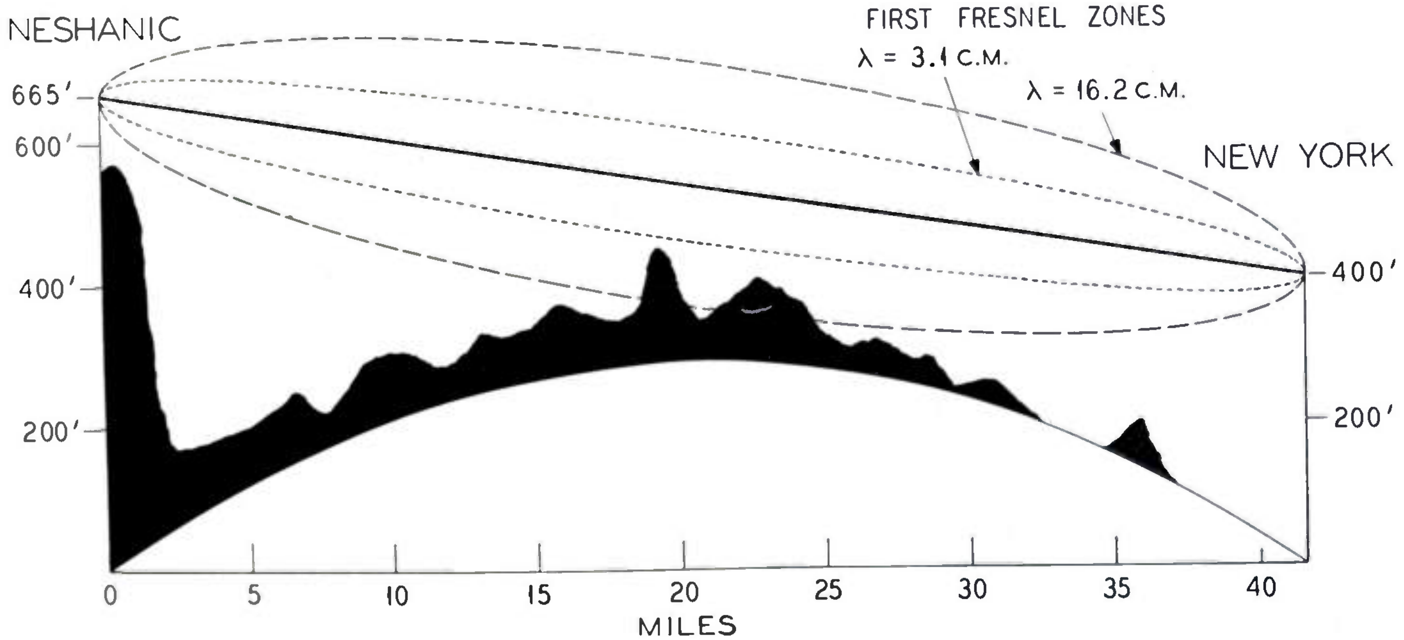

Fig. 1—Profile of propagation test path (true earth radius).

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* Decimal classification: R423.16. Original manuscript received by the Institute, August 9, 1949; revised manuscript received, February 13, 1950. Presented, 1950 National IRE Convention, New York, N. Y., March 9, 1950.

† Western Union Telegraph Company, New York, N. Y.

tested, are also shown. The picture of the 100-foot Neshanic Tower, with the four transmitting antennas facing New York, is shown in Fig. 2. In Fig. 3 (the propagation tower may be seen in the background, with the main radio beam telegraph system tower appearing in the foreground.

For comparative purposes, as explained, four individual circuits or channels were set up and operated as indicated in Table I.

Fig. 2—Neshanic propagation tower.

Fig. 3—Neshanic main radio beam tower (left), and propagation tower (right).

Each of the four transmitters consisted essentially of a reflex klystron oscillator tube for generating radio-frequency power, its associated cavity, a matching section and wave meter. These components, comprising the "head-end" unit, were inclosed in a temperature controlled transmitter box. To minimize frequency drift, particular attention was given to the design and construction of the regulated power supply and to the thermostatically controlled oven temperature regulation system. Pictures of the various transmitters are shown in Fig. 4.

TABLE I

Channel No.	Frequency Mc	Wavelength cm	Transmitter Power milliwatts	Beam Width Measured at Half-Power Points degrees
1	1,850	16.2	137	9.0
2	4,150	7.2	159	4.7
3	6,325	4.7	13.6	2.6
4	9,550	3.1	7.4	1.9

The antenna reflectors were of the parabolic type, 48 inches in diameter and physically identical at both the transmitting and receiving terminals.

In order that a continuous record of the relative transmitter power would be obtained, a small portion of the transmitted energy was rectified in a crystal and the output fed to a four-point electronic recorder. The recorder is shown in Fig. 5 mounted below the transmitter control unit. This control unit permitted adjustment of the radio-frequency power output, frequency, cavity current, and repeller voltage of the transmitting tube without opening the ovens.

Each of the four individual receiving installations consisted essentially of an antenna system, a temperature-controlled "head-end" receiving unit, mounted in the immediate vicinity of the antenna, and an intermediate-frequency amplifier unit and recording equipment located about three hundred feet distant. The recording equipment consisted of four Brown Instrument Company strip recorders, one for each channel, which were calibrated so as to provide a continuous graphical record of variations in signal strength at the input terminals of the receivers. These recorders were operated at a strip speed of six inches per hour.

Although the strip record proved useful in examining specific cases of abnormal fading, this record was not in a form suitable for analyzing fading effects on a statistical basis. Due to the high recording speed, a 24-hour interval involved a strip 12 feet in length. A more convenient and useful representation of the essential data was obtained by transferring the maximum, minimum, and average value of signal strength for each hourly interval from the strip record to weekly charts. Samples of these charts have been included to illustrate unusual fading conditions which will be discussed in a later para-

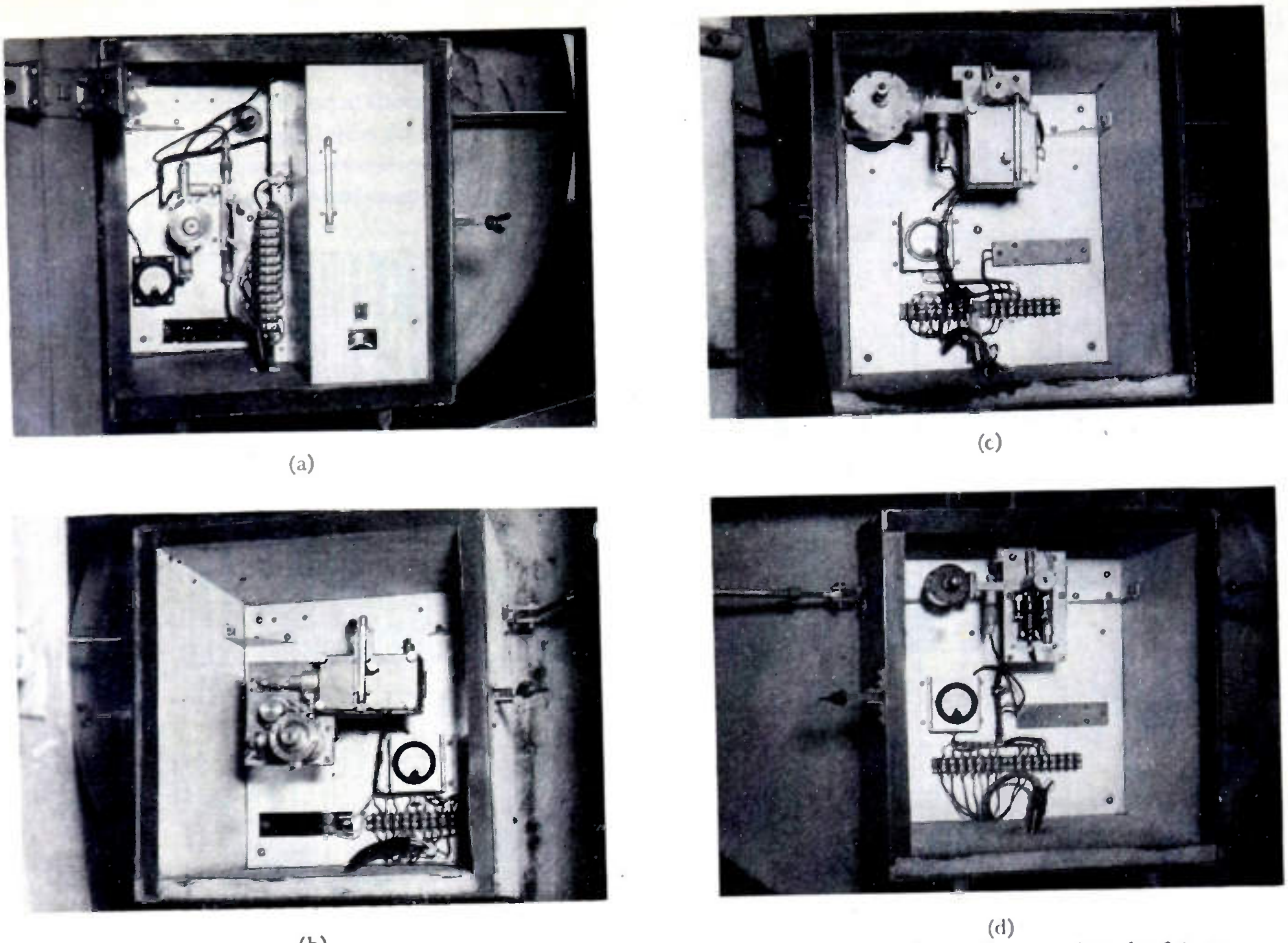
(a) (b) (c) (d)

Fig. 4—(a) Transmitter for 16.2 cm; (b) transmitter for 7.2 cm; (c) transmitter for 4.7 cm; (d) transmitter for 3.1 cm.

graph. It should be noted, however, that signal strength is expressed in terms of decibels above or below a fixed reference level, designated as the normal signal level. The normal signal level, determined separately for each channel, represents the average signal strength for periods during which fading was absent or negligible, i.e., periods when transmission was highly stable. Once established, the value of the normal signal level was considered fixed, except that adjustments to this reference level were made whenever it was found necessary or desirable to recalibrate the equipment in order to compensate for changes, such as tube replacements, affecting either the transmitter power level or receiver performance.

TABLE II

Channel No.	Frequency Mc	Wave-length cm	Free-Space Attenuation db	Net Attenuation db	Received Carrier-to-noise Ratio db
1	1,850	16.2	134	83	26
2	4,150	7.2	141	76	34
3	6,325	4.7	145	73	27
4	9,550	3.1	149	69	28

Information relating to path attenuation and received carrier-to-noise ratio, is shown separately for each channel in Table II.

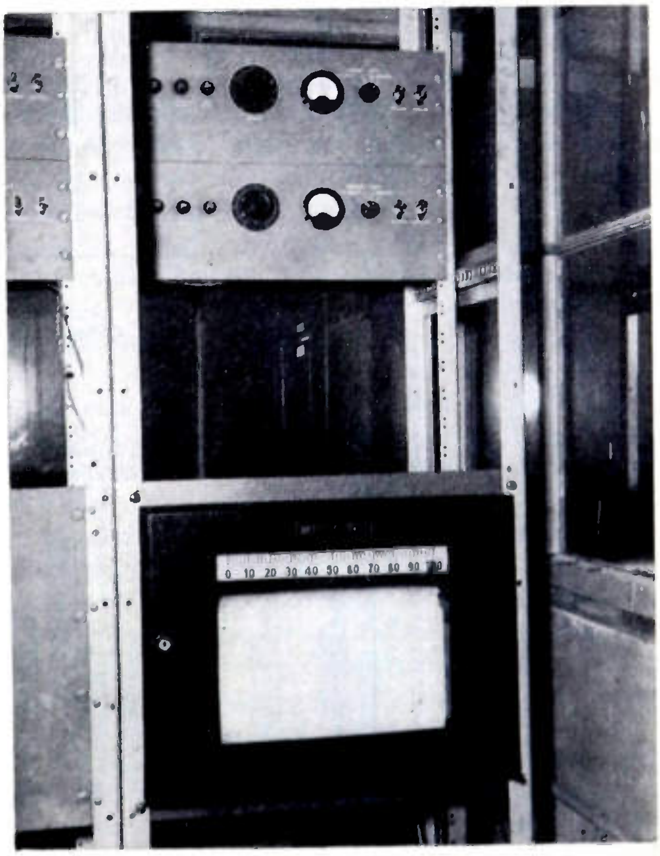

Fig. 5—Control unit and recorder.

In each case, the difference between the value shown for free-space attenuation and that for net attenuation corresponds to the power gain of the antenna system. An efficiency of 65 per cent was assumed for each antenna. Carrier-to-noise ratios were calculated on the basis of a receiver bandwidth of 4 Mc and receiver noise power of approximately 17 db above thermal.

During periods when conditions for reception were favorable, i.e., periods when fading was entirely absent or negligible, the received signal approached within a few db of the calculated value on each channel. The extent of variations in signal strength is reflected in Fig. 6,

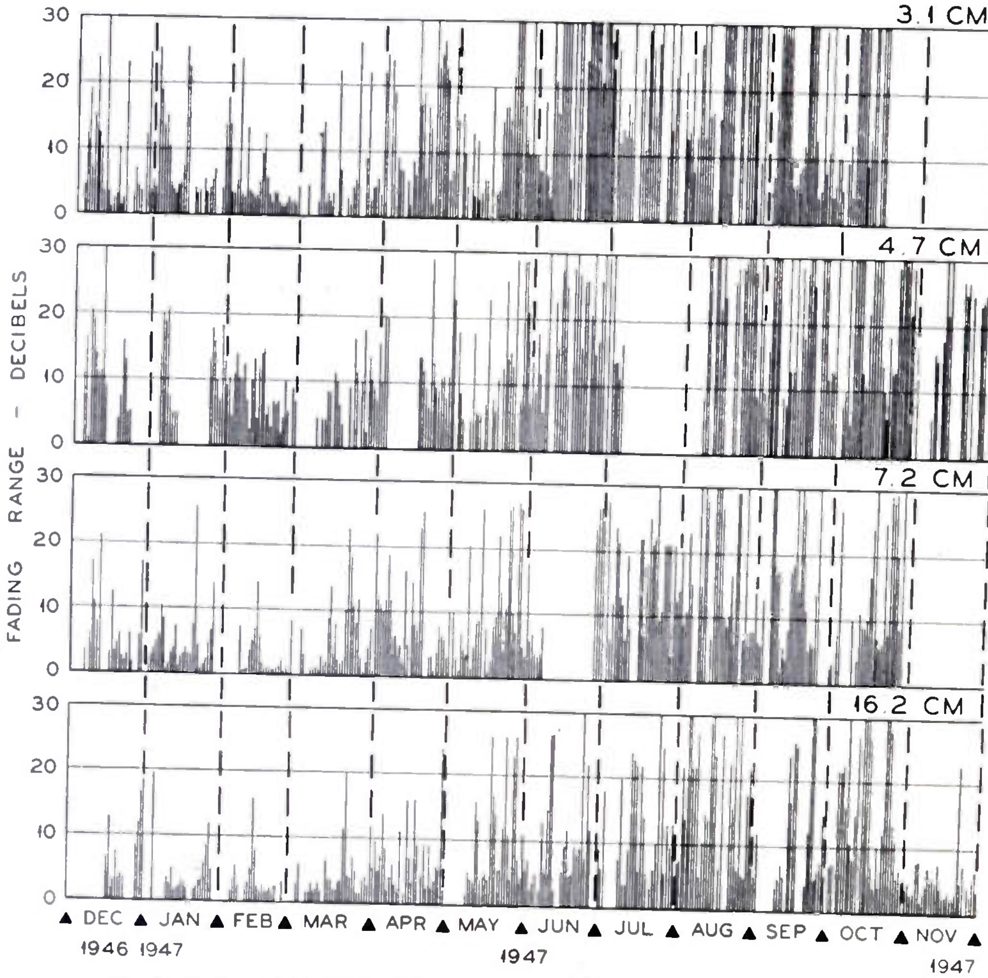

Fig. 6—Daily fading ranges from December, 1946, through November, 1947.

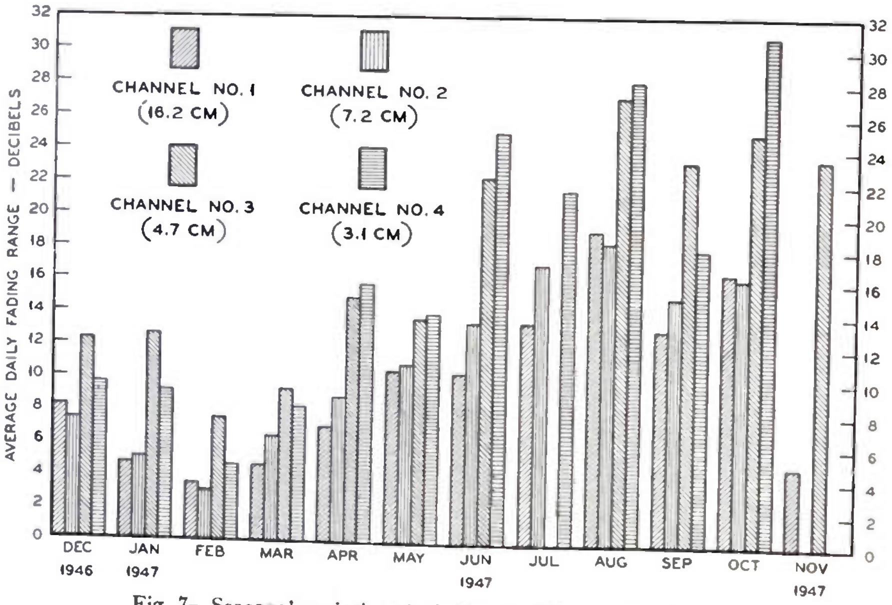

Fig. 7—Seasonal variations in fading at different wavelengths.

which shows the daily fading range separately for each channel from December, 1946, through November, 1947. The daily fading range, as used here, is the ratio of the maximum signal occurring during a 24-hour period to the minimum signal for the same period, expressed in decibels. As may be observed from the chart, fading generally was more pronounced during the summer months and more severe on the higher frequencies.

To provide a comparison of fading ranges at different wavelengths, and in order to depict inherent seasonal characteristics of fading, the daily fading ranges were averaged by month for each channel and the results represented graphically in Fig. 7. The general trend of increasing fading range with frequency is evident. Variations in fading range may also be observed to follow a rough seasonal pattern. In order to accentuate the seasonal trend, the average daily fading range is shown separately for 7.2 cm in Fig. 8.

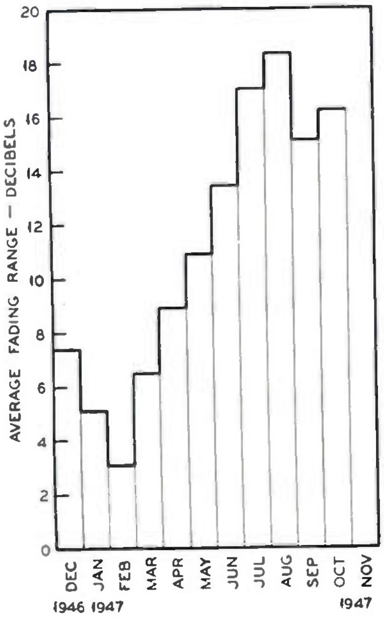

Fig. 8—Seasonal variation in fading at 7.2 cm.

Fading exhibits not only seasonal characteristics but diurnal as well. Fig. 9 illustrates diurnal variations in fading which occurred during the months of February and July, 1947, on 7.2-cm transmission. Here, the average fading range, in decibels, is shown graphically for each hour of the day. The fading range for any one hour represents the ratio of maximum to minimum signal strength for that hour, expressed in decibels. Fig. 9 reveals a marked difference in fading with respect to different periods of the day, i.e., that maximum fading generally occurred during the night and early morning hours while minimum fading occurred in the late morning and afternoon, usually between 10 A.M. and 6 P.M., local standard time. Although diurnal effects are illustrated for only one summer and one winter month and for only one wavelength, a similar definite diurnal tendency was found to exist throughout the year on all four wavelengths examined. The pronounced difference between winter and summer fading is again evident. Diurnal effects are also reflected in Fig. 12, which will be explained in a later paragraph.

Average fading range data are useful and essential in dealing with system design problems bearing on fading tolerance considerations, although such data obviously are not sufficient. These data, for example, furnish no information with respect to the relative frequency of occurrence of abnormal fades or to the duration of such fades, both of which are important considerations since they are related to circuit continuity which in turn determines circuit performance.

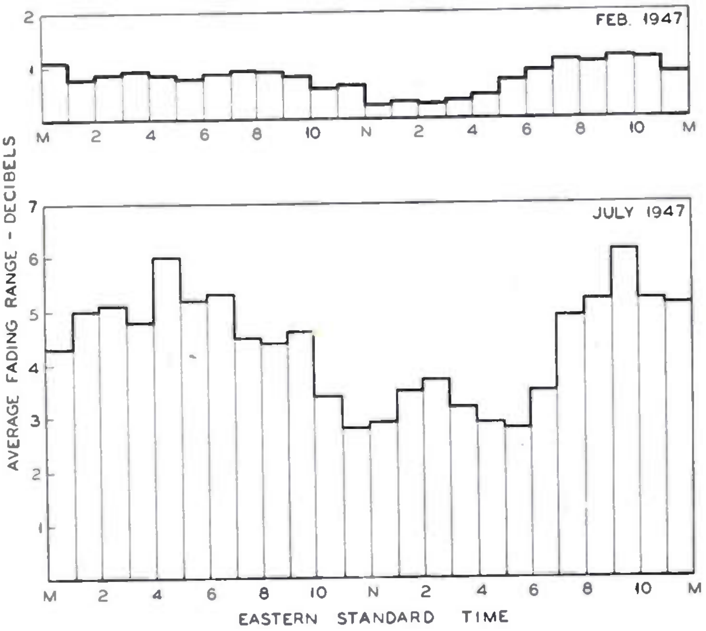

Fig. 9—Diurnal variation in fading at 7.2 cm during the months of February and July, 1947.

Relative circuit performance may be gauged fairly well from distribution curves of instantaneous received signal levels. For this purpose, distribution curves have been prepared for the wavelengths employed in the test and are shown in Figs. 10 and 11. The curves shown in Fig. 10 represent a period of five months, from June 1, 1947, through October, 1947; the curves shown in Fig. 11 represent a period of three months, from January 1, 1947, through March, 1947. Data on which these curves are based were obtained by selecting a number of signal levels and determining the relative time the received signal was below each level during the periods shown. The month of October was included in the summer period, since fading during that month corresponded very closely with typical summer fading, as may be seen from Fig. 7.

In using these curves, one should remember that the reference level represents normal signal strength, as explained earlier, and that this reference is roughly 2 or 3 db below the free-space value. Also, high accuracy is not claimed, particularly for those portions of the curves appearing below the 0.1 per cent level which may be in error by as much as 20 per cent.

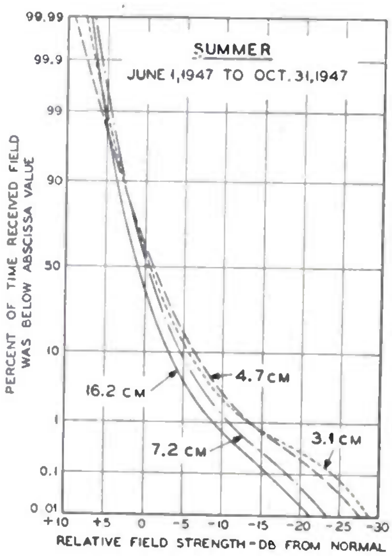

Fig. 10—Distributions of relative field strength in summer.

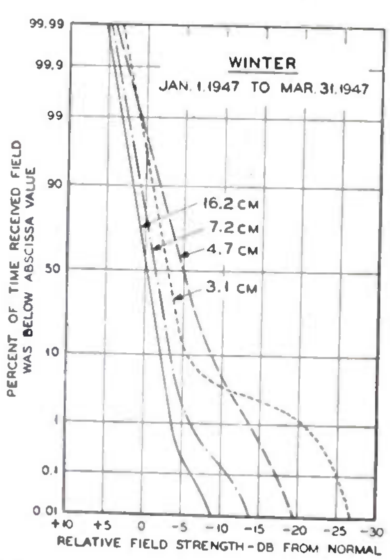

Fig. 11—Distributions of relative field strength in winter.

The superior performance of the longer wavelength circuits is evident in both summer and in winter. A comparison of the curves in Fig. 10 with those of Fig. 11 reveals several interesting characteristics with respect to performance at the different wavelengths tested. The summer curves are generally similar and the spread is not very great. This similarity is particularly evident for the two lower frequency curves, exhibiting a well-defined mutual relationship which does not appear to exist for the two higher frequencies. The winter curves will be seen to indicate a definite normal probability distribution, at least down to a given signal level. The spread for the winter curves is more pronounced. Again, a well-defined relationship occurs between the two lower-frequency curves. Greater variability in signal strength is evident on all four wavelengths during the summer period. Summer signal strengths were often more than 5 db above normal for several hours at a time, occasionally as high as 10 db for shorter periods on the two higher frequency channels, whereas during winter months signal levels greater than 3 or 4 db above normal were rare and of very short duration. As will be observed, upward swinging of the signal during the summer season was more pronounced at the shorter wavelengths. On the other hand, it was not unusual for the received signal to drop into the noise region during the summer months whereas such occasions were comparatively rare during the winter period, except that for a few days in January the 3.1-cm signal was abnormally low. The extreme fading indicated for 3.1 cm is attributed to heavy fog which occurred in January, 1947 and continued over a period of several days. Most of the time during this period the signal was in or near the noise level.

Although the curves in Figs. 10 and 11 furnish a fairly representative picture as to relative circuit performance at different wavelengths, in certain applications of microwave radio transmission, such as commercial telegraph operation, information bearing on the frequency of interruptions in service due to fading conditions, however short they may be, and the approximate time of the day when such interruptions are most likely to occur, is particularly instructive. Interruptions in service will result whenever fading conditions cause the received signal to vanish or to fall below some nominal critical signal level. This fact suggested that an examination into minimum signal levels, particularly those which fall below what might be considered as a safe operating level, should reveal a rough guide for use in determining the hourly periods during which interruptions in service are most likely. Accordingly, four arbitrary signal levels, −6, −12, −18, and −24 db, were selected and the number of hourly minima falling below each of these levels was recorded for each of the four wavelengths, according to the hour of the day and month, for the entire year of 1947. By number of hourly minima is

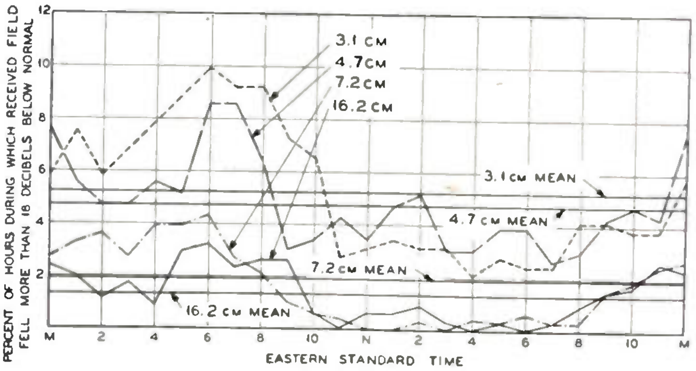

Fig. 12—Diurnal variations in fading, from January 1, 1947, to December 31, 1947, as illustrated by the per cent of hourly periods during which the received field fell more than 18 decibels below normal.

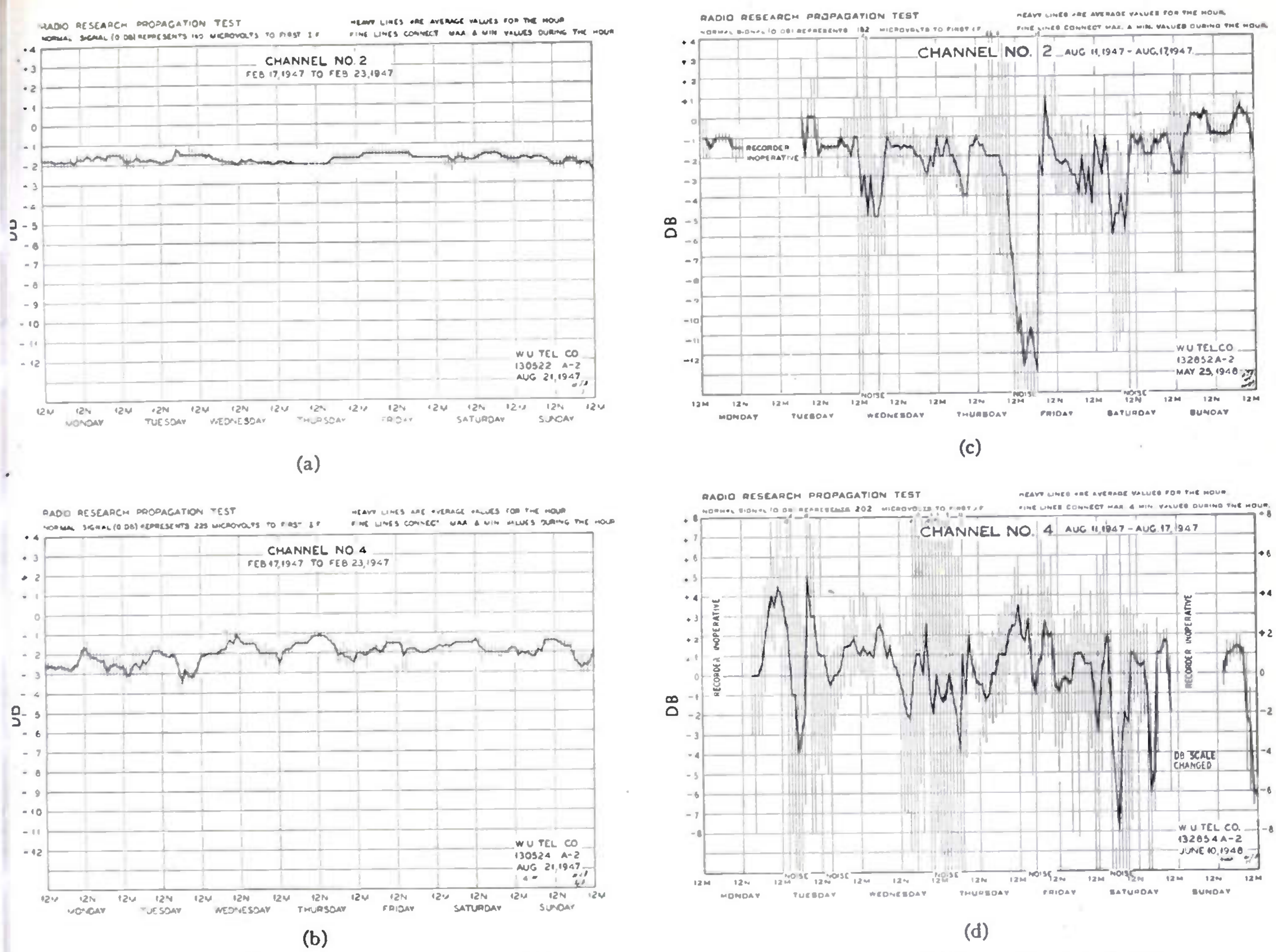

Fig. 13—(a) Hourly fading data on 7.2 cm, February 17–23, 1947; (b) data on 3.1 cm, February 17–23, 1947; (c) data on 7.2 cm, August 10–17, 1947; (d) data on 3.1 cm, August 10–17, 1947.

meant the number of hourly periods during any portion of which the signal fell below the level considered. Since these data were arranged in tabular form and are extremely voluminous, they are not included in this paper. Fig. 12, however, shows the manner in which the hourly minima were distributed with respect to the −18-db level for each hour of the day and for each of the four wavelengths. These curves illustrate the superior performance of the longer wavelengths and again reflects diurnal effects on all four circuits.

Data were also obtained as to the number of hourly occasions in which the received signal fell into the noise region and these data suggest that the number of such occasions varied roughly as the square of the frequency during summer months and nearly linear with frequency during winter months, again excluding the abnormal period mentioned earlier. Usually, the received signal was in the noise region for very short intervals of time, generally a matter of seconds, and when integrated represented virtually negligible percentage of the total time considered.

The four charts of Fig. 13 provide an interesting comparison of fading effects on transmission at 7.2 and 3.1 cm. These charts are intended to reveal the marked contrast between unusually favorable conditions which occurred during an entire week in February and the unusually variable conditions encountered during a week in August.

Conclusion

Microwave radio communication systems operating on wavelengths down to about 5 cm may be expected to perform in a highly satisfactory manner throughout the year under properly selected conditions of path length, path clearance, and transmitter power. Circuit performance will deteriorate appreciably and progressively at shorter wavelengths and at wavelengths of about 3 cm and below, shorter path lengths together with nominal increases in transmitter power are indicated if a high degree of circuit continuity is required. Fading is observed with greater prevalence during the summer months, and at nights, as a result of changing meteorological conditions. Diversity reception will prove beneficial in reducing the effects of multipath transmission.

Acknowledgement

This test was made possible through the combined efforts of engineers from the RCA Victor Division of RCA and the Radio Research Division of Western Union. The RCA engineers provided valuable assistance in formulating the test program, designing the test equipment, and in placing the test into operation. To these men and to members of our Radio Research Division who participated in operation of the test, the authors express their sincere appreciation.

Magnetic Triggers*

AN WANG†, ASSOCIATE, IRE

Summary—**Magnetic cores of fairly rectangular hysteresis loop material are used as a trigger device. Magnetic fluxes are used instead of electrical currents to indicate the two stable positions of the trigger. This paper shows how the magnetic flux level may be detected without a mechanical motion. The construction and functioning of several types of such magnetic triggers are discussed.**

I. Introduction

TRIGGER CIRCUITS have been used extensively during recent years in various electronic devices. They usually consist of a pair of vacuum tubes connected in such a way that two stable states of the system exist. The magnetic triggers to be described below are a result of research in utilizing the fundamental hysteretic properties of a ferromagnetic material as a trigger device.

Any magnetic material, easily saturated and having fairly large retentivity properties, can be considered as a trigger device by itself. Consider the hysteresis loop shown in Fig. 1. When the material has been under the influence of a large positive magnetizing force, positive residual magnetism is retained. The material will remain at the point *1*. If it was last subjected to a negative magnetizing force, the material will remain in the state of negative magnetization represented by the point *0*. These two states, *1* and *0*, represent the two possible stable conditions of the magnetic material. They can be easily reversed by the application of a magnetizing force of sufficient amplitude in the opposing direction. In contrast to the vacuum-tube trigger pair, the magnetic flux polarity, rather than the voltage level, determines the two stable states. While dc voltages are necessary to maintain the dc currents flowing in the vacuum-tube trigger pair, it is not necessary to have a magnetizing force to maintain a flux in the magnetic core material. Thus, a magnetic trigger should be able to maintain the triggered position without the need of any power. The magnetizing pulse then takes the place of the triggering pulse. It should be powerful enough to

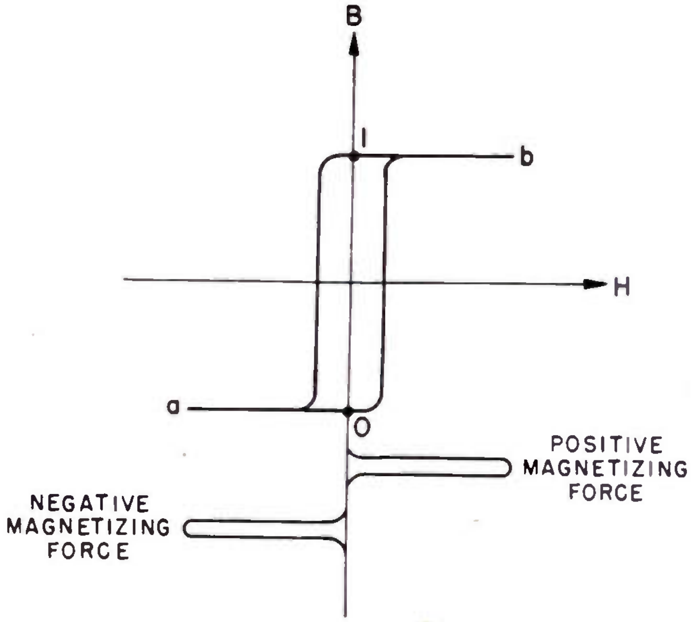

Fig. 1—Hysteresis loop curve of a highly saturated magnetic material.

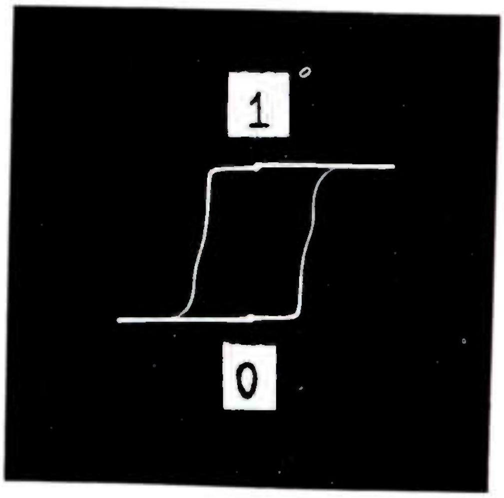

Fig. 2—Operating path of the magnetic trigger in *B-H* diagram.

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* Decimal classification: R282.3×538×R136. Original manuscript received by the Institute, August 25, 1949; revised manuscript received, January 18, 1950. The work was undertaken under Contract W 19-122-AC-24 between Harvard University and the United States Air Force.

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drive the magnetic core to saturation. Alternate magnetizing pulses of 10 microseconds duration are applied to a magnetic material. The flux of the material is flip-flopped between two saturation values. The material travels around the hysteresis loop while being triggered and remains at one of the two stable points *1* and *0* when not being triggered. (See Fig. 2.) Notice that the points *1* and *0* are stationary when no magnetizing pulses are applied.

While it is very easy to detect the difference of electric current, it is very difficult to sense the magnetic flux polarity unless it is changing. Thus, the simple magnetic trigger will not work statically until means of detecting the flux polarity are found.

II. MAGNETIC TRIGGER PAIR

By the nature of the saturation phenomena of magnetic materials, it is possible to determine the polarity of the residual magnetism by applying a powerful magnetizing force. Assume that this magnetizing force is in the negative direction (Fig. 1). Then, if the residual magnetism is at the point *0*, very little flux change results. When the magnetizing force is discontinued, the magnetization returns to the point *0*. If the residual magnetism is at point *1*, a large change of flux results along the trace of the hysteresis loop dropping from *1* to *a* and then returning from *a* to *0*. During the sharp decrease of flux, a large voltage is induced across the coils linking the magnetic core. Then, no matter what polarity the residual magnetism, eventually the magnetic core returns to point *0*. A large induced voltage across the secondary coil indicates that the original state was at point *1*, while a very small induced voltage indicates that the original state was at point *0*.

A similar magnetic core is now used in addition. Let us assume that it is at position *0*. If the large induced voltage from the first core is able to drive the second core from its *0* position to *1*, the flux condition of the first core is transferred to the second core. This has been demonstrated experimentally.[1] The connections are as shown in Fig. 3. The hysteresis curve below each core represents its magnetic state. When both cores are at *0* position, the first magnetizing pulse, being negative, changes the flux of core number 1 very little, so that core number 1 produces very little linking current i_{12} to change the flux of core number 2. Similarly, the application of the second pulse will do the same. Cores number 1 and number 2 experience small loops *oac* and *o'a'c'*, constituting a stable limit cycle for each core. When a *1* is stored in core number 1, the application of the first magnetizing force changes the flux from *1* to *a*, from where it drops back to *0*. This large change of flux induces sufficient voltage in its coil N_1 to produce a current i_{12} which should be able to drive core number 2 from *0'* to *b'* to *1'*. The differential equation of the link circuit is

$$-N_1\frac{d\phi_1}{dt} - N_2\frac{d\phi_2}{dt} = Ri_{12} + L\frac{di_{12}}{dt}. \qquad (1)$$

Integrating this,

$$-N_1\Delta\phi_1 - N_2\Delta\phi_2 = Ri_{12}dt + L\Delta i_{12}. \qquad (2)$$

As the flux of core number 1 changes from position 1 to position *a*, the flux of core number 2 changes from *0'* to *b'*. N_1 and N_2 should be equal if they are symmetrical. The current i_{12} is always positive, and so is Δi_{12} during this change. It is necessary that

$$\Delta\phi_1 + \Delta\phi_2 = \text{a negative value}. \qquad (3)$$

This means that the change of flux from 1 to *a* should always be greater than the change of flux from *0* to *b*.

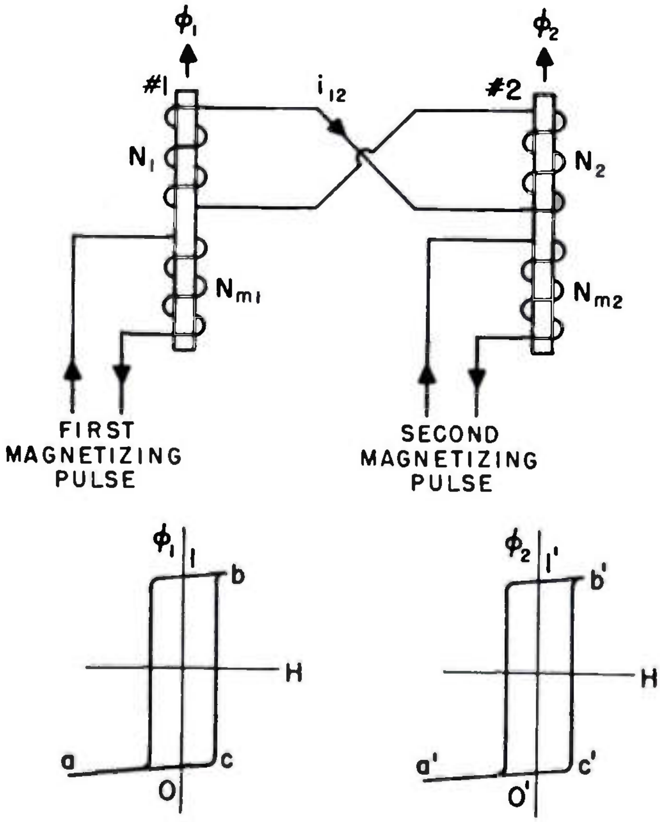

Fig. 3—Magnetic trigger pair and its operation.

If it is possible to make core number 1 return from *0* to *b* to *1* by the application of the second magnetizing pulse, the stored digit is kept there. This requires the existence of a stable major hysteresis loop with slight unsymmetry, which has been available. The operation of such a magnetic trigger pair is then stable.

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[1] Air Force Contract W19-122-AC-24 Progress Report No. 2, Harvard Computation Laboratory, sect. IV, 8; Nov. 10, 1948. Progress Report No. 4, Harvard Computation Laboratory, sect. V, 3; May 10, 1948.

III. Modified Magnetic Trigger Pair

The above-mentioned unsymmetry in hysteresis loops can be stable only in a limited region. If $\Delta\phi_2$ is smaller than $\Delta\phi_1$ by a certain amount, the next time $\Delta\phi_1$ is less than $\Delta\phi_2$, and so on. Gradually, in several cycles, both cores will end up a position *0*, and the stored information will be lost. This is the case experimentally when too much resistance or leakage inductance is present in the linking circuit. Also, to produce change of flux as the frequency of operation is made higher, i_{12} must necessarily be larger, due to the presence of eddy current. This makes the loop more unsymmetrical and consequently more unstable. The trouble can be eliminated by a modification of the basic circuit. Note that (2) can be satisfied by using a value of N_1 greater than N_2, while $\Delta\phi_1$ and $\Delta\phi_2$ are exactly equal and opposite. Under this condition, stable operation of the trigger pair is possible without the necessity of having stable unsymmetrical hysteresis loops. The circuit of Fig. 4 is used. Rectifiers are necessary in the linking circuit so that when the first core is driven, the upper link is operative, and when the second core is driven, the lower link is operative.

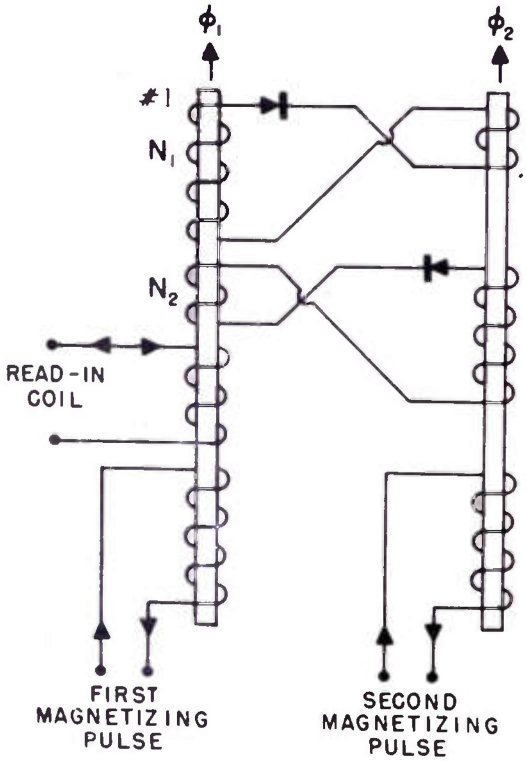

Fig. 4—Modified magnetic trigger pair.

In this case, stability of operation can be obtained more easily at the expense of two extra rectifiers. This form of trigger pair has been tested to hold information for an indefinitely long time, and can be triggered back and forth at a repetition frequency up to 50 kilocycles per second. The rectifiers used are of selenium type. Germanium diodes of course can also be used. There are many ways of introducing signals into the trigger pair. One way is to apply the positive or negative read-in pulse to the first core at the same instant as the second magnetizing pulse is applied to the second core.

Fig. 5 shows the flux variation of the first core as the two magnetizing pulses are alternately applied. The upper portion represents the condition when a *0* is stored in the trigger pair, while the lower portion represents the condition when a *1* is stored. The oscillogram shows the operation at a repetition frequency of 5 kilocycles per second.

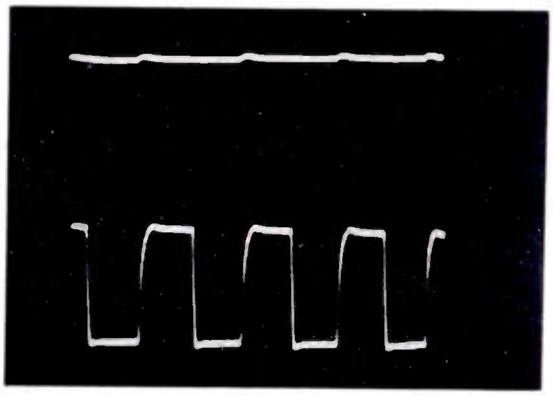

Fig. 5—Flux variation of the magnetic trigger pair.

IV. Single Stroke Trigger

The trigger pairs described in the above two sections use two magnetic cores. The magnetic flux condition of the first core is transferred to the second core for temporary storage by the first magnetizing pulse; then the state is transferred back by the second magnetizing pulse. Since there is no other way of telling the flux polarity statically, two transfers must be made. There is a great advantage in the use of a single core and a single magnetizing pulse to determine the polarity of the magnetic flux, while preserving that flux.

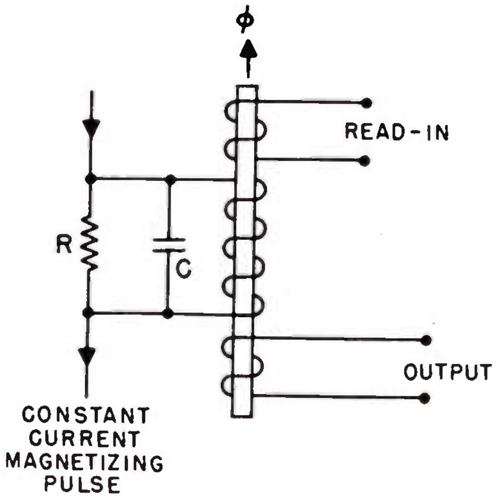

Fig. 6—Single-stroke magnetic trigger.

The procedure is as follows (Fig. 6): Assume first that the core is magnetized in the positive direction, or at the position *1*. The magnetizing pulse changes the position from *1* to *a*; this causes a large voltage across the coil and charges the capacity C. Then the magnetizing pulse ceases to flow. C is discharged through the coil which offers very low inductance because it is already saturated in the direction of the current. During the discharge of the capacity C, the circuit operation is very like a parallel resonance of L, C, and R, where R is high and L very low. The Q of this parallel circuit is equal to $R\sqrt{C/L}$, which is high. The charge on the condenser C is easily reversed without much damping. However, when the condenser discharges again, the discharging current flowing in the winding is in such a direction as to cause the flux of the core to change from *0* to *1* again. The coil offers a high inductance, the discharging rate is much slower, and the damping high. If the values of R, C, and this inductance are such that damping is closed to critical, the condenser is completely discharged while the flux returns from *0* to *1* again. This situation is very similar to a single-stroke electrical trigger pair. The polarity of the flux of the core returns to the original condition automatically a certain time after the first triggering. This interval is determined by the discharging time of the condenser. In Fig. 7 are shown the change of flux ϕ with time as the magnetizing pulse is applied, the voltage across the condenser V_c, the current i_L through the coil, and the magnetizing current i_M. If originally the core is in the state *0*, the magnetizing pulse does not change the flux by any appreciable amount. The coil essentially short-circuits the CR circuit. The condenser is not charged and everything remains the same as the magnetizing pulse subsides. The output is very small. The flux change for this case is shown as the last curve in Fig. 7.

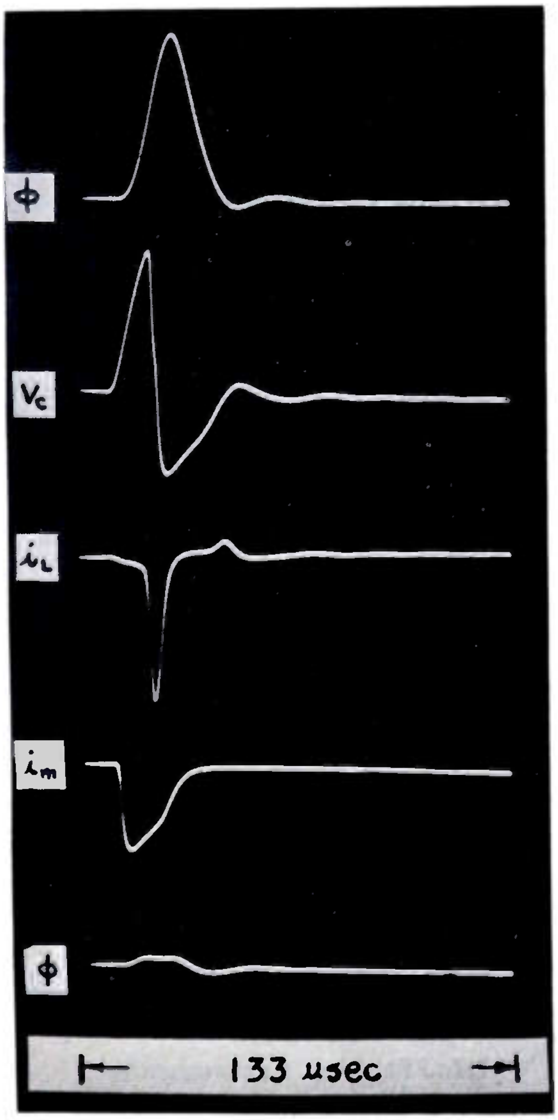

Fig. 7—Flux, voltage, and current shape of the single-stroke magnetic trigger during triggering.

V. CONCLUSION

The above discussions show the general feasibility of using magnetic cores as triggers. There is every possibility that such a magnetic trigger can take the place of vacuum-tube trigger pairs for some of their applications. Binary digits can be stored in such units. A stored digit can be delivered out or be transferred to another core. The possibility of transferring binary digits from core to core directly makes it possible to construct an information delay line in which a series of binary digits can be pushed along a series of such cores by magnetizing pulses at any rate from a very low speed up to about 30 kilocycles per second. This has been described in another paper.[2] Exact mathematical treatment of the subject is still difficult in view of the highly nonlinear characteristic of the hysteresis loop of the core material.

VI. ACKNOWLEDGMENT

The author wishes to express his sincere gratitude to H. H. Aiken for his original suggestion of the problem, and to W. D. Woo for his valuable suggestions.

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[2] A. Wang and W. D. Woo, "Static magnetic storage and delay line," *Jour. Appl. Phys.*, vol. 21, pp. 49–54; January, 1950.

Feedback in Very-High-Frequency and Ultra-High-Frequency Oscillators*

F. J. KAMPHOEFNER†, MEMBER, IRE

Summary—This paper covers a study of feedback considerations in low-power negative-grid triode oscillators for the frequency range of 100 to 1,000 Mc. Discussion is mainly confined to oscillators using a single tuned circuit between grid and plate in the modified Colpitts circuit wherein the feedback is provided by the internal tube capacitances. The optimum feedback conditions are derived, and the analysis is applied to several typical oscillators.

I. INTRODUCTION

THE COLPITTS circuit, having a single tuned circuit between grid and plate, and using the internal tube capacitances for feedback, is of practical interest for several reasons. The cathode lead inductance is not a part of the radio-frequency circuit, and therefore does not affect the feedback. Second, the plate and grid lead inductances can be lumped with the external tuned circuit. Third, there is a minimum of parts, since there is no external radio-frequency path between plate and cathode or grid and cathode. Finally, it can be shown that the Colpitts circuit is less subject to initial transit-time effects than the Hartley or tuned-plate-tuned-grid types.

A paper by Bell, Gavin, James, and Warren,[1] in 1946, pointed out the basic requirements for oscillation in a triode, and also the necessary feedback conditions. The present study was initiated in order to extend these results.

II. CONDITIONS FOR OSCILLATION

Fig. 1(a) illustrates the general case of the negative-grid triode oscillator. Applying linear circuit theory, the net dynamic grid-plate admittance is:

$$\begin{aligned} G_{gp}'' = G_{gp} + \{ [&G_{gk}G_{pk}(G_{gk} + G_{pk} + G_m) \\ &+ B_{pk}G_{gk}(B_{pk} + B_m) + B_{gk}(B_{gk}G_{pk} \\ &+ B_mG_{pk} - B_{pk}G_m)] \div [(G_{gk} + G_{pk} + G_m)^2 \\ &+ (B_{gk} + B_{pk} + B_m)^2]\} \end{aligned} \tag{1}$$

$$\begin{aligned} B_{gp}'' = B_{gp} + \{ [&B_{gk}B_{pk}(B_{gk} + B_{pk} + B_m) \\ &+ B_{gk}G_{pk}(G_{pk} + G_m) + G_{gk}(G_{gk}B_{pk} \\ &+ B_{pk}G_m - B_mG_{pk})] \div [(G_{pk} + G_{gk} + G_m)^2 \\ &+ (B_{gk} + B_{pk} + B_m)^2]\}. \end{aligned} \tag{2}$$

In the case of the Colpitts circuit with small losses and short transit time, the admittance presented to the tuned circuit becomes:

$$G_{gp}'' = \frac{\omega^2 C_{gk}^2 - \mu\omega^2 C_{gk}C_{pk}}{\dfrac{(1+\mu)^2}{r_p} + r_p\omega^2(C_{gk} + C_{pk})^2} \tag{3}$$

and

$$B_{gp}'' = \omega C_{gp} + \frac{\omega^3 C_{gk}C_{pk}r_p^2(C_{gk}+C_{pk}) + \omega C_{gk}(1+\mu)}{(1+\mu)^2 + r_p^2\omega^2(C_{gk}+C_{pk})^2}. \tag{4}$$

The necessary conditions for oscillation are that the magnitude of the negative conductance presented by the tube be greater than the positive conductance of the load, and the susceptance presented by the tube be equal in magnitude but opposite in sign to that of the load at the frequency of oscillation.

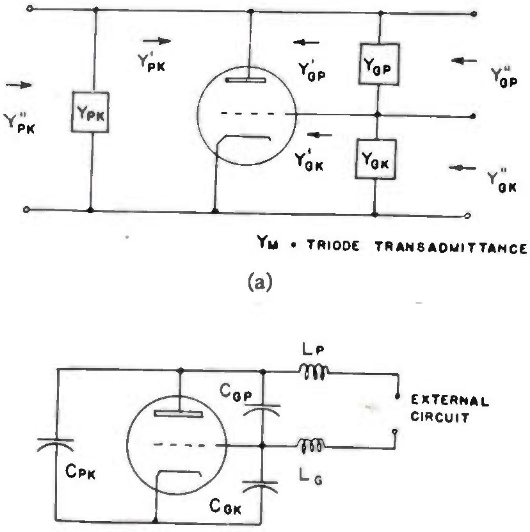

Fig. 1—(a) Basic diagram of a negative-grid triode oscillator; (b) special case of Colpitts oscillator using internal feedback.

III. FEEDBACK RELATIONSHIPS

Considering (3), if the grid-plate conductance, G_{gp}'', is to be negative, the term $(\mu\omega^2C_{gk}C_{pk})$ must be greater than $\omega^2C_{gk}^2$, or $(C_{gk}/C_{pk}) < \mu$; this verifies the familiar low-frequency relationship for the Colpitts circuit. If the conductance is to be most negative, however, it is not necessarily true that the ratio C_{gk}/C_{pk} should be as small as possible; on the contrary, it will have a certain optimum value. In order to establish this optimum ratio, examine the more general equation obtained from (1) by assuming only that transit time is small:

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* Decimal classification: R355.912. Original manuscript received by the Institute, June 5, 1949; revised manuscript received, February, 21, 1950.

† Stanford Research Institute, Stanford, Calif.

[1] J. Bell, M. R. Gavin, E. G. James, and G. W. Warren, "Triodes for very short waves—oscillators," *Jour. IEE* (London), part IIIA, vol. 93, pp. 833–846; 1946.

$$G_{gp}'' = G_{gp} + \frac{G_{gk}G_{pk}(G_{gk} + G_{pk} + G_m) + (G_{pk}B_{gk}^2 + G_{gk}B_{pk}^2 - G_m B_{gk} B_{pk})}{(G_{gk} + G_{pk} + G_m)^2 + (B_{gk} + B_{pk})^2}. \tag{5}$$

It has been shown[1] that G_{gp}'' can be computed as a function of B_{gk} and B_{pk} for a particular tube, and Fig. 2 shows such computations as a family of curves for the 6F4, without losses. It is apparent from the curves that:

1. If $(B_{gk}/B_{pk}) < 18$, then the grid-plate conductance becomes negative and oscillation is possible. Since $B_{gk}/B_{pk} = C_{gk}/C_{pk}$ at any one frequency, this checks the previously mentioned condition for oscillation in a Colpitts oscillator, namely $(C_{gk}/C_{pk}) < \mu$ $(\mu = G_m/G_p = 18)$.

2. At a given frequency, the most negative grid-plate conductance is obtained by making both C_{gk} and C_{pk} large and equal to each other.

3. If either B_{pk} or B_{gk} must be limited to a small value, then an optimum ratio of B_{gk}/B_{pk} (i.e., C_{gk}/C_{pk}) can be defined.

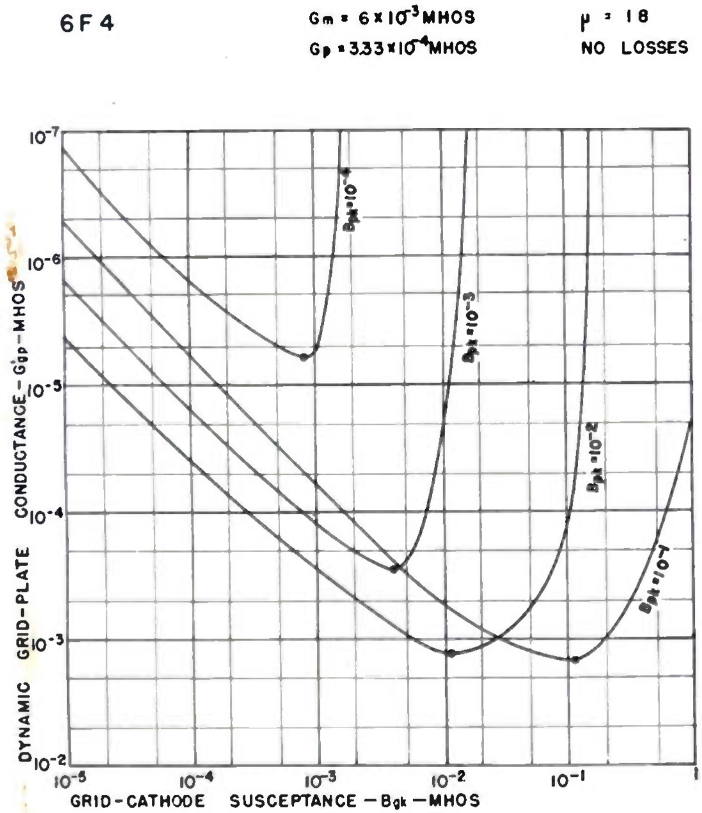

Fig. 2—Dynamic grid-plate conductance of 6F4 as a function of feedback susceptances.

4. A locus of points can be drawn through the minimum points of the family of constant B_{pk} curves; this defines the optimum grid-cathode susceptance for a given plate-cathode susceptance. Similarly, an envelope of the family of constant B_{pk} curves can be drawn; this defines the optimum plate-cathode susceptance for a given grid-cathode susceptance.

Now in an ultra-high-frequency oscillator, it is not feasible to increase the feedback capacitances indefinitely; too large a set of feedback capacitances will be reflected in a large capacitance at the remaining pair of tube elements, resulting in a low-impedance tank circuit, and a low self-resonant frequency for the tube; as a result, a compromise may be necessary.

To compare the 6F4 acorn triode with a fictitious 6F4 having "optimum" feedback capacitances, plot from (3) the grid-plate conductance as a function of frequency for a 6F4 using interelectrode capacitances for feedback, and add to the same illustration the optimum curves as extracted from Fig. 2. This has been done in Fig. 3. It can be seen that at low frequencies the internal feedback capacitances of the 6F4 are smaller than those which would produce the most negative output conductance.

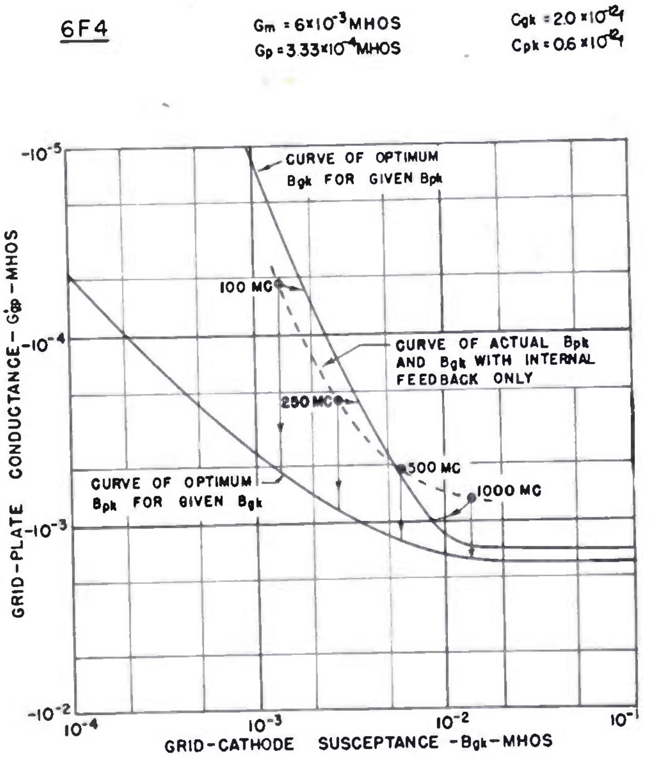

Fig. 3—A comparison of actual and optimum feedback ratios for 6F4.

At 250 Mc, for example, the conductance can be made more negative by increasing either C_{gk} or C_{pk}, the change being much greater if it is C_{pk} that is altered. (Notice that the locus of optimum B_{gk} for a given value of B_{pk} must be approached along a curve of constant plate-cathode susceptance—*not* grid-plate conductance.) Increasing both capacitances would make the conductance still more negative, but would excessively increase the effective grid-plate capacitance presented to the tuned circuit. At higher frequencies the situation is a little different. At 1,000 Mc, for example, Fig. 3 shows that the grid-plate conductance is again made more negative if both feedback capacitances are increased or if the plate-cathode capacitance alone is increased. If the grid-cathode capacitance is increased without also increasing the plate-cathode capacitance, however, the grid-plate conductance then becomes less negative. In any case, the internal capacitances are large enough in the 6F4, and of such ratio, that the conditions for oscillation are favorable for the entire ultra-high-frequency range.

In order to compare this with a tube that would not oscillate without external feedback, consider the 2C37 co-planar triode with these characteristics:

$\mu = 25$ $\quad C_{gk} = 1.40 \times 10^{-12} f$

$r_p = 5{,}000$ ohms $\quad C_{pk} = 0.02 \times 10^{-12} f.$

Applying (3), the dynamic conductance presented to the load is plotted in Fig. 4, and is seen to be positive for the entire frequency range. The tube will therefore not oscillate in an ultra-high-frequency Colpitts circuit unless some form of feedback is added. Since it is often difficult to obtain satisfactory external feedback, a second tube type has been made commercially available which is basically a 2C37 with an artificially increased plate-cathode capacitance; this is the 2C36, which has the characteristics:

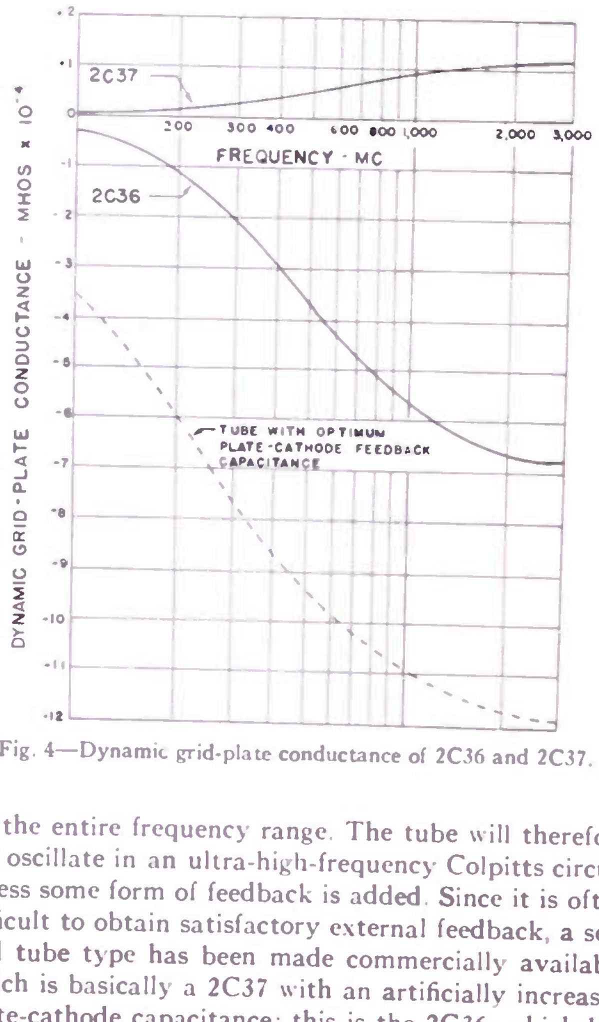

Fig. 4—Dynamic grid-plate conductance of 2C36 and 2C37.

$\mu = 25$ $\quad C_{gk} = 1.40 \times 10^{-12} f$

$r_p = 5{,}000$ ohms $\quad C_{pk} = 0.36 \times 10^{-12} f.$

The grid-plate conductance of the 2C36 is also plotted in Fig. 4, and it can be seen that it requires no additional external feedback for oscillation.

The dotted curve, showing optimum feedback, is based on the assumption that the internal plate-cathode capacitance is adjusted for optimum feedback. (This is the practical case, since $C_{gk} > C_{pk}$.) This curve could be obtained from another family of curves, just as Fig. 3 was taken from Fig. 2, but a less laborious method is to recall that it is the envelope of the first family of curves that is desired, and this envelope can be found by setting $(\partial G_{gp}''/\partial B_{pk}) \equiv 0$ and solving for B_{pk}. If this is done, using (5) rather than (3) so that the expression is kept more general by the inclusion of loss terms, the result is:

$$\begin{aligned} &B_{gk}(2G_{gk} + G_m)B_{pk}^2 + \{2G_{gk}[(G_{gk} + G_{pk} + G_m)^2 \\ &\quad - G_{pk}(G_{gk} + G_{pk} + G_m)] + 2B_{gk}^2(G_{gk} - G_{pk})\}B_{pk} \\ &\quad - \{B_{gk}[G_m(G_{gk} + G_{pk} + G_m)^2 \\ &\quad + 2G_{gk}G_{pk}(G_{gk} + G_{pk} + G_m)] \\ &\quad + B_{gk}^3(2G_{pk} + G_m)\} \equiv 0. \qquad (6) \end{aligned}$$

This is a quadratic equation of the form

$$aB_{pk}^2 + bB_{pk} + c = 0$$

so that,

$$B_{pk} = \frac{-b \pm \sqrt{b^2 - 4ac}}{2a}. \qquad (7)$$

(When practical values are inserted in the solution, the sign of the second term of the quadratic solution is seen to be positive.) The computation by this method is still tedious, but it is much easier than plotting the entire family of curves.

IV. LEAD INDUCTANCE

The equations of this paper are independent of the influence of lead inductance as long as the feedback is internal and the grid and plate lead inductances are considered as part of the external tuned circuit. If it is desired to lump the grid-plate lead inductances with the tube, this can be done by converting the computed dynamic admittance of the tube to an impedance form and adding to it the lead reactance and loss resistance. The requirement of the external tuned circuit for oscillation at a given frequency can then be predicted by the relationship

$$R_{\text{tube}} + R_{ckt} \leq 0 \qquad (8)$$

$$X_{\text{tube}} + X_{ckt} = 0, \qquad (9)$$

provided that $(G_{gp}'')^2 \ll (B_{gp}'')^2$.

Despite the limitations of electron transit time and the fact that it is based on a small-signal approach, the analysis can be of value when applied to the practical design of low-power very-high-frequency and ultra-high-frequency oscillators. An example and extension of this line of attack can be found in a companion paper by Pettit,[2] whose help in the formulation of the present paper is gratefully recognized.

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[2] J. M. Pettit, "Very-high-frequency triode oscillator using a series tuned circuit," this issue, PROC. I.R.E., pp. 633–635.

Ultra-High-Frequency Triode Oscillator Using a Series-Tuned Circuit*

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Summary—**As the conventional triode oscillator is applied to higher and higher frequencies, one or both of two phenomena finally prevent oscillation. These phenomena are transit-time effects and self-resonance of the tube elements. The analysis described in this paper refers these effects to the resistance and reactance requirements of the tube upon the external circuit. In those instances where a given triode is limited primarily by the reactance effects, due to large lead inductance, it is shown that the self-resonant frequency of the tube is not the upper limit of oscillation if one departs from the conventional parallel-tuned circuit and uses a series-tuned circuit instead.**

THERE IS ALWAYS an interest in extending the frequency range of any given triode tube to higher frequencies. This is especially true in receiver applications, where, in addition, it is often desired to have wide tuning range and single control. The customary oscillator used in this application is the standard Hartley or Colpitts circuit, but at higher frequencies the modified Colpitts circuit shown in Fig. 1 is employed. In this latter circuit the feedback is accomplished through the internal tube capacitances. It has been shown by previous authors[1] that this circuit is one of a class that is most favorable in reducing transit-time effects. With conventional small triodes, it provides a useful oscillator for receiver and signal generator applications up to frequencies of several hundred megacycles, often using the "butterfly" circuit for the tuned element, as indicated in Fig. 1.

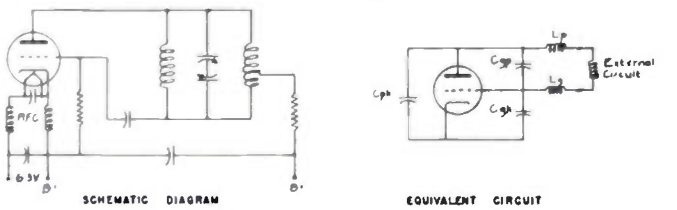

Fig. 1—Ultra-high-frequency oscillator—modified Colpitts circuit.

In order to analyze the requirements of the external tuned circuit for a given triode tube, it is expedient to investigate the nature of the two-terminal impedance looking into the grid-plate terminals, then to compute the series resistive and reactive components as a function of frequency.[2] When this is done, results such as those shown in Fig. 2 are obtained (6F4 triode, connections to only *one pair* of grid and plate terminals). Calculated data are presented, together with measurements made upon a scaled model operating at one tenth of the frequencies shown. Simple linear theory is utilized, employing a real and constant transconductance. This provides, of course, no information as to the magnitude of the oscillation nor the available power output, but does define a set of minimum conditions required for oscillation. The lower curve shows the negative resistance provided by the tube, and it follows that the external circuit must provide a positive resistance equal to or less than the appropriate negative value at the frequency of interest. Likewise, the reactance of the external circuit must be equal in magnitude to and opposite in sign from the reactance provided by the tube.

This paper is restricted to those applications where the reactance of the external circuit is the limiting factor, especially operation in the region near and above the self-resonant frequency of the tube, which occurs at about 730 Mc for the example in Fig. 2. It is thus assumed that transit-time effects have not impaired the magnitude of the negative resistance presented by the tube to such an extent that usable external circuits are too lossy to operate. This situation might be expected

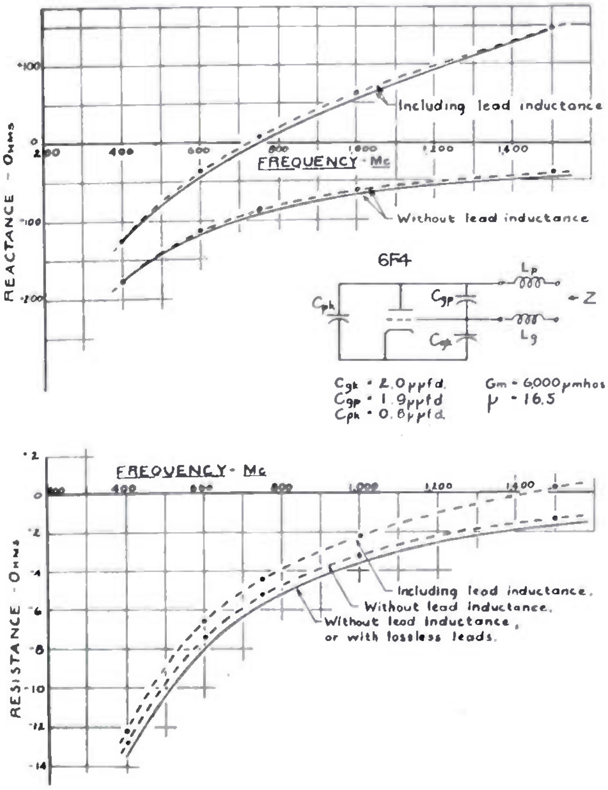

Fig. 2—Dynamic impedance presented to tuned circuit by 6F4 Colpitts oscillator. (Note: Solid curves are computed. The dotted curves and encircled points are measured, using the frequency-scaled model.)

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* Decimal classification: R355.912. Original manuscript received by the Institute, September 12, 1949; revised manuscript received, January 4, 1950.

† Stanford University, Stanford, Calif.

[1] J. Bell, M. R. Gavin, E. G. James, and G. W. Warren, "Triodes for very short waves—oscillators," *Jour. IEE* (London), part III A, vol. 93, pp. 833–846; 1946.

[2] Further details on this approach will be found in a companion paper, F. J. Kamphoefner, "Feedback in vhf and uhf oscillators," in this issue, PROC. I.R.E., pp. 630–632.

to prevail in close-spaced triodes mounted in miniature envelopes, for instance the triode-connected 6AK5, where lead inductance is large and it, instead of transit time, will tend to be the limiting factor.

The manner in which the external circuit operates in conjunction with the tube to provide the proper reactance and thus to determine the oscillating frequency is shown in the calculated curves of Fig. 3. Here the customary parallel-tuned circuit is indicated, although the reactive component plotted is the equivalent series value. The point of intersection of the reactance required by the tube and the reactance available from the circuit determines the operating frequency, shown to be 570 Mc in the example under consideration.[3] It will be noted that the resonant frequency of the circuit alone is actually 667 Mc. The tube acts like a capacitance, very closely equal to the grid-plate capacitance, and thus requires an inductive reactance to produce resonance. Obtaining this from a tuned circuit requires that the circuit be tuned to a higher frequency than the desired frequency of oscillation of the combined circuit. It can be plainly seen from Fig. 3 that, no matter how high the resonant frequency of the external circuit may be, oscillations cannot occur at a frequency higher than the self-resonant frequency of the tube, namely 730 Mc. At this frequency a short circuit across the grid-plate terminals would suffice equally well.

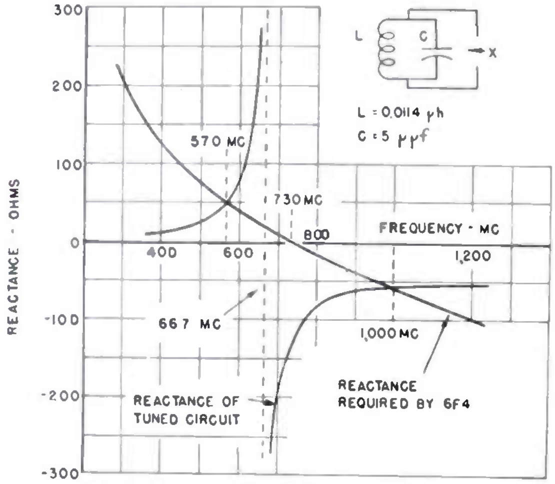

Fig. 3—Reactance of parallel-tuned circuit.

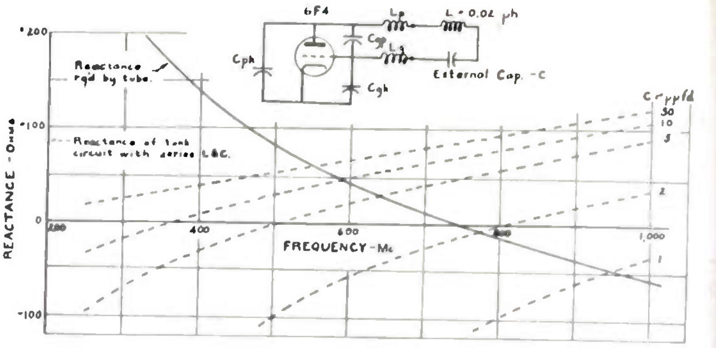

Fig. 4—Series-tuned circuit for tuning through the self-resonant frequency of the 6F4.

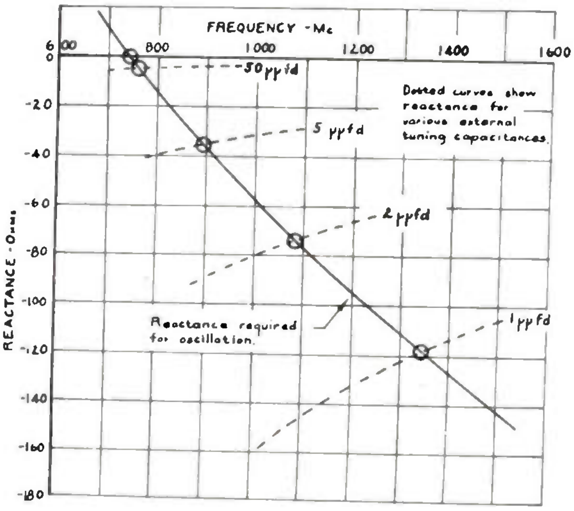

Fig. 5—Capacitance required to operate a 6F4 triode above its self-resonant frequency.

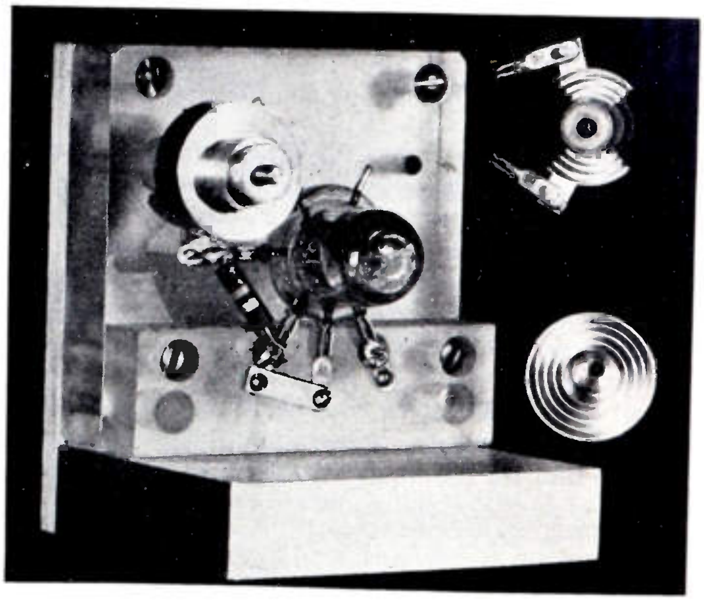

Fig. 6—Series-tuned ultra-high-frequency oscillator.

It is noted that the behavior of the reactance required by the triode in the vicinity of self-resonance resembles that of a series-tuned circuit. Again considering the examples shown in the illustration, if a small inductance and a series capacitance are connected to the grid-plate terminals as in Fig. 4, the dotted curves in this figure show the calculated reactance presented by the series circuit for different values of tuning capacitance. Once more the intersections of these dotted lines with the solid lines, representing the reactance required by the tube, determine the operating frequencies of the oscillator circuit. It is seen that with the capacitance variation shown, a substantial tuning range is covered, extending from well below to considerably above the self-resonant frequency of the tube.

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[3] There is also an intersection at approximately 1,000 Mc, but the lower frequency provides the more favorable resistance conditions.

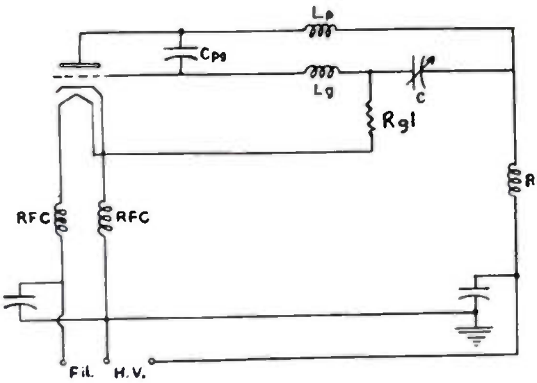

Fig. 7—Schematic diagram of series-tuned ultra-high-frequency oscillator.

It is indeed feasible to operate such a circuit by adding only a capacitor as the external circuit, using the inductance of the leads as the inductive portion of the circuit. This is doubtless familiar to a few readers. The tuning ranges calculated in this circumstance are shown in Fig. 5. Such an oscillator, using the triode for which the previous calculations have been made, is pictured in Fig. 6, with its detailed schematic in Fig. 7. The capacitor used was a concentric air trimmer, modified to provide a split stator and an isolated rotor. The frequency range is approximately 630 to 900 Mc, with a corresponding capacitance variation of 8.1 to 1.5 micromicrofarads.

For applications in which a large tuning range is desired, both capacitance and inductance variation can be employed simultaneously. The chart of Fig. 8 shows the range of possibilities for the 6F4 example.

It will be noted that, from the viewpoint of oscillation theory, the active part of the circuit, the electron stream, continues to see a parallel reactance. The external series circuit merely serves to adjust the magnitude of the inductive branch paralleled by the grid-plate capacitance.

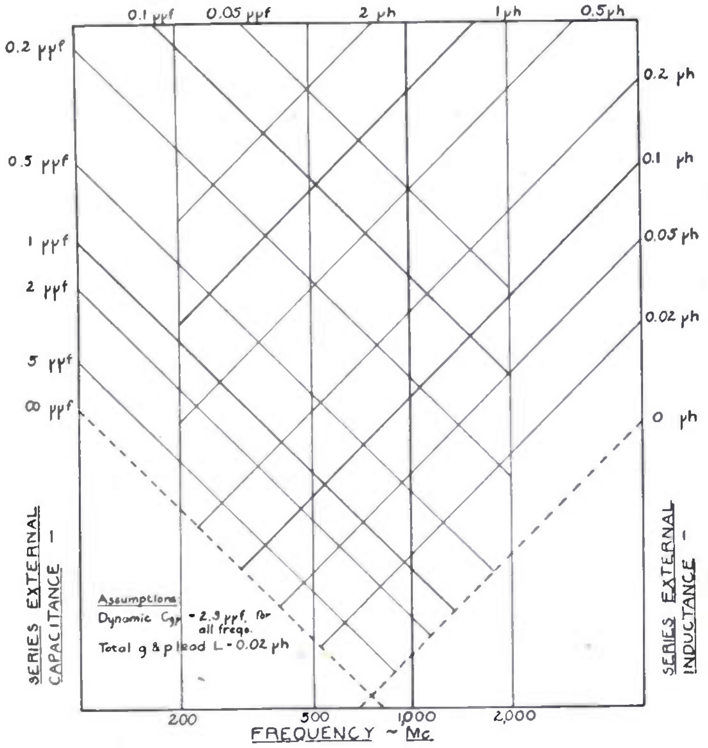

Fig. 8—Frequency range of a 6F4 Colpitts oscillator for single series-tuned circuit with both inductance and capacitance variable.

In conclusion it should be emphasized that the so-called self-resonant frequency of the tube used in a triode oscillator need not be a limit on the frequency coverage if proper choice of the external circuit is made.

ACKNOWLEDGMENT

The assistance of F. J. Kamphoefner in both the analytical and the experimental work is gratefully acknowledged.

The Theory of a Three-Terminal Capacitor*

ROBERT E. CORBY†

Summary—**Theoretical equations are derived for the insertion loss of a three-terminal capacitor which check the experimental curves within their precision of measure. The condenser is assumed to behave like some equivalent transmission line and the line constants of this line determined as a function of frequency. Skin effect and proximity effect are taken into consideration, and engineering curves are plotted to make the prediction of the behavior of such a condenser easy to determine.**

INTRODUCTION

THE THEORY OF a three-terminal capacitor described by Allison and Beverly[1] can be developed from Maxwell's equations and transmission line theory. The insertion loss[2] of this filter is given as

$$P = 20 \log (R_{10}/Z_x),$$

where R_{10} is the impedance level of the line (usually chosen as 10 ohms), and Z_x is the impedance of the device whose loss is desired. From transmission line theory, the sending end impedance of an open-circuited line is $Z_0 \coth \gamma l$ where the characteristic impedance Z_0 is to a close approximation,

$$Z_0 = (L/C)^{1/2}((\omega^2 + k^2)/\omega^2)^{1/4},$$

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* Decimal classification: R215. Original manuscript received by the Institute, August 22, 1949; revised manuscript received, January 13, 1950.

† University of Arizona, Tucson, Ariz.

[1] W. M. Allison and N. E. Beverly, "New high-frequency capacitor," *Trans. AIEE*, vol. 63, p. 915; December, 1944.

[2] Alan Watton, Jr., "The duct capacitor," PROC. I.R.E., vol. 36, pp. 550–554; April, 1948.

where k is the ratio of R/L and the other symbols have their usual significance. The propagation constant γ can be divided into its real and imaginary parts α and $j\beta$, where α and β are given as

$$\alpha^2 = (\omega LC/2)((\omega^2 + k^2)^{1/2} - \omega)$$
$$\beta^2 = (\omega LC/2)((\omega^2 + k^2)^{1/2} + \omega).$$

Substituting these equations back into the expression for insertion loss, we get

$$P = 20 - 10 \log L/C - 5 \log (\omega^2 + k^2)/\omega^2 + 10 \log (\sinh^2 \alpha l + \sin^2 \beta l)/(\sinh^2 \alpha l + \cos^2 \beta l). \quad (1)$$

In using the above equation, it is convenient to use frequencies associated with the multiples of 1/8 wavelength. Frequency and wavelength λ are related from elementary theory by

$$f^2 = 16\pi^2/\lambda^2(\lambda^2C^2R^2 + 16\pi^2LC).$$

DETERMINATION OF THE LINE CONSTANTS

It can easily be shown that the magnetic field is to a close approximation the same as that which would exist if the condenser were unrolled. The general skin effect equation $\nabla^2 i = j\omega\mu\sigma i$ can be solved in rectangular coordinates from which the expression for the impedance per unit of length can be determined.[3] Separating this impedance into its real and imaginary parts, we can write for R and L

$$R = R_0(2m)^{1/2}(\sinh (2m)^{1/2} + \sin (2m)^{1/2}) /(\cosh (2m)^{1/2} - \cos (2m)^{1/2})$$

$$L = \frac{2\pi a 10^{-9}}{w(2m)^{1/2}}\left[\frac{\sinh (2m)^{1/2} - \sin (2m)^{1/2}}{\cosh (2m)^{1/2} - \cos (2m)^{1/2}}\right] + \frac{2\pi c 10^{-9}}{w},$$

where

$m \equiv \sigma\omega\mu a^2/4$
$\sigma \equiv$ conductivity of foil
$\omega \equiv$ angular frequency
$R_0 \equiv$ direct current resistance per unit length
$\mu \equiv$ conductor permeability
$c \equiv$ dielectric thickness
$a \equiv$ conductor thickness
$w \equiv$ narrow conductor width.

Both of the above equations can be simplified by defining two function F_1 and F_2 which contain all the hyperbolic and trigonometric terms. These can be plotted as a function of m alone. The final expressions for R and L written in more useful units are

$$R = R_0F_1$$
$$L = (2\pi/w)(aF_2 + c),$$

where R_0 is in ohms/cm, and L is in micromicrohenrys/cm, a is in mils, w is in inches, and c is in mils.

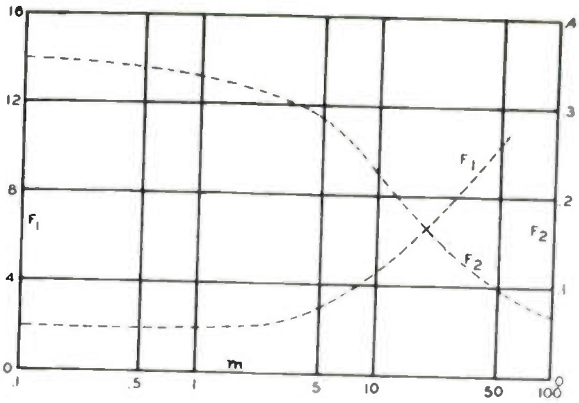

Fig. 1—Useful curves in predicting the behavior of a condenser.

$$F_1 = \frac{(2m)^{1/2}[\sinh (2m)^{1/2} - \sin (2m)^{1/2}]}{\cosh (2m)^{1/2} - \cos (2m)^{1/2}} \quad m = \frac{\sigma\omega\mu a^2}{4}$$

$$F_2 = \frac{\sinh (2m)^{1/2} - \sin (2m)^{1/2}}{(2m)^{1/2} \cosh (2m)^{1/2} - \cos (2m)^{1/2}}$$

CONCLUSION

Fig. 1 is a curve showing the functions F_1 and F_2. From these curves, R and L can be determined, from which α and β can be determined, from which the insertion loss can be determined. An examination of (1) shows that as the frequency approaches infinity, the third and fourth terms approach zero and the insertion loss becomes

$$P = 34.47 + 10 \log K + 20 \log w/c$$

where K is the dielectric constant, w is in inches, and c is in mils.

Measurements were made on many condensers with different dimensions, conductivities, and dielectrics, and the above equations were found to be valid up to frequencies of the order of 50 Mc. At frequencies above this, the loss in general dropped rapidly with frequency due in large part to the effect of the external inductance inherent in the measurement apparatus. Fig. 2 shows a typical check between theory and experiment.

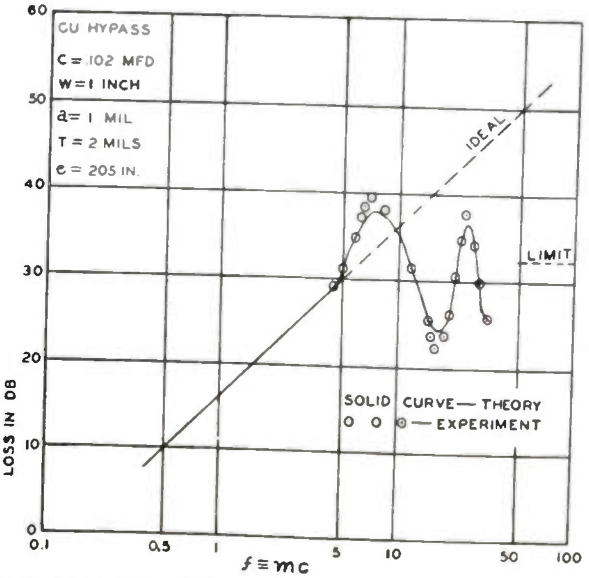

Fig. 2—Correlation between theory and experimental results.

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[3] S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, Inc. New York, N. Y., chap. 6; 1944.

Maximum Tank Voltage in Class-C Amplifiers*

LEO E. DWORK†, MEMBER, IRE

Summary—**Theoretical considerations are presented to justify the frequent appearance in class-C amplifiers of radio-frequency plate voltages which are greater than the dc plate voltage. A method is developed for predicting the magnitude of this rf voltage under any given set of conditions. This method is then experimentally verified for a given case.**

INTRODUCTION

FREQUENTLY, when preliminary adjustment of the tank circuit of a class-C amplifier is attempted, or when the specifications on a transmitter require that the tank be tuned with load disconnected, there is a voltage flashover across the tank condenser even though it was designed with what was believed to be a large safety factor. This voltage breakdown is due to a radio-frequency plate voltage which can be many times the dc plate supply voltage.

The usual procedure in the design of class-C amplifiers has been to assume a radio-frequency peak plate voltage which is slightly less than the dc voltage on the plate of the tube. From this assumed rf voltage the required plate tank circuit elements are then chosen. Specifically, the tank circuit condenser is chosen to withstand the combined voltage stress of the dc and the assumed rf plate voltages. However, as mentioned above, this has under certain conditions resulted in a voltage breakdown of the plate tank condenser.

The purpose of this investigation was as follows:

(a) To determine what causes this high rf tank voltage.
(b) To determine what factors influence its magnitude.
(c) To devise a method for predicting the value of this voltage under any given set of conditions.
(d) To experimentally check this prediction for a given case.

The basic circuit of the class-C power amplifier is shown in Fig. 1. The grid is biased beyond cutoff and driven by a sine wave generator. The resulting plate current flow is a pulse which, under steady-state conditions, can be Fourier analyzed into components consisting of a dc term and fundamental and harmonics of the grid driving voltage.

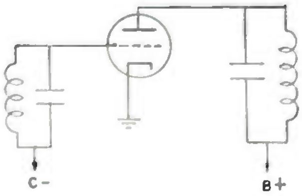

Fig. 1—Basic circuit of the class-C amplifier.

Under normal conditions in a class-C amplifier the plate circuit consists of a parallel coil and condenser which is tuned to the same frequency as the grid driving voltage. The Q of this tank circuit is such that the resulting rf plate voltage is to a very close approximation a pure sine wave. The magnitude of this rf plate voltage is equal to the product of the fundamental component of the plate current pulse by the impedance of the plate tank circuit at this frequency. The use of the constant current curves of the tubes is desirable when analyzing the class-C amplifier since the ac plate and grid voltages are both sinusoidal.[1] The path of operations on these constant current curves becomes an ellipse, or of course a straight line or circle.

Symbol Definitions

e_b = instantaneous plate voltage
E_{bb} = dc plate voltage
E_{pm} = peak value of the rf plate voltage
e_c = instantaneous grid voltage
E_{cc} = dc grid voltage
E_{gm} = peak value of the rf grid voltage
i_b = instantaneous plate current
i_{p1} = instantaneous value of the fundamental component of plate current
I_{pm1} = peak value of i_{p1}
$Z_p = |Z_p| \underline{/\psi}$ = plate tank impedance at fundamental frequency
Z_r = impedance of the same plate tank circuit when it is adjusted to antiresonance
X_c = reactance of the plate tank condenser at fundamental frequency
r_p = plate resistance of the tube
μ = amplification factor of the tube
θ = supplement of the phase angle between the rf grid and plate voltages
T = period of the rf grid voltage
α_1 = the peak value of the fundamental sine term resulting from the Fourier analysis of i_b
β_1 = the peak value of the fundamental cosine term resulting from the Fourier analysis of i_b.

TANK CIRCUIT ADJUSTED TO ANTIRESONANCE

When the plate tank circuit is tuned to antiresonance—unity power factor—the fundamental component of the plate current pulse must be in phase with the ac tank voltage.

Consider the elliptical path of operation shown in Fig. 2. Zero time is taken as the instant of maximum

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* Decimal classification: R363.15. Original manuscript received by the Institute, August 27, 1948; revised manuscript received, December 22, 1949.

† RCA Institutes, Inc., New York, N. Y.

[1] I. E. Mouromtseff and H. N. Kozanowski, "Analysis of the operation of vacuum tubes as class-C amplifiers," PROC. I.R.E., vol. 23, pp. 752–779; July, 1935.

tank voltage (minimum plate voltage). The ac plate voltage varies with time as a cosine wave. After equal time intervals on the positive and negative side of zero time the instantaneous plate voltages are equal, but, as can be seen from Fig. 2, the instantaneous plate currents are not equal. Since the fundamental sine term of the Fourier analysis of the plate current pulse is given by

$$\frac{2}{T}\int_{-T/2}^{+T/2} i_b \sin \omega t dt$$

it cannot be zero, and therefore the fundamental plate current cannot be in phase with the tank voltage. However for the straight line path of operation, line $A-C$ in Fig. 2, the plate current pulse has zero axis symmetry, and therefore the fundamental component of the plate current is in phase with the tank voltage. Thus when the tank circuit is tuned to antiresonance, the path of operation becomes a straight line and the sinusoidal plate and grid voltages are 180° apart.

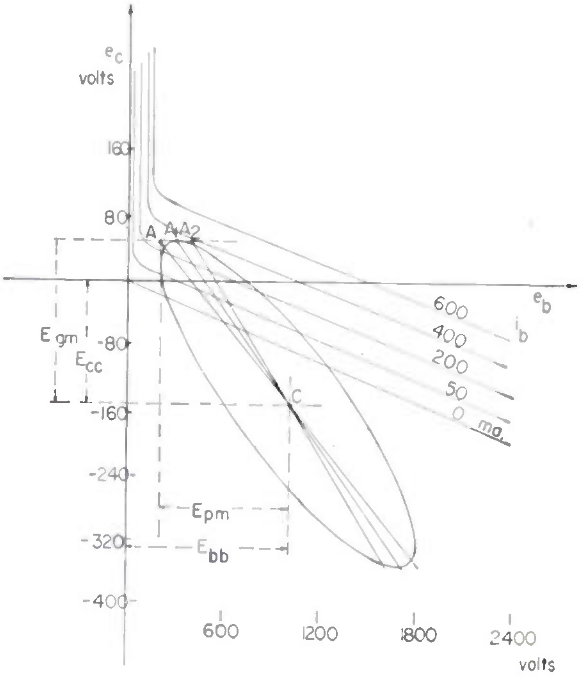

Fig. 2—Paths of operation with E_{pm} less than E_{bb} (Type 211 tube).

For a given dc plate voltage, dc grid voltage, and rf grid drive, plate tank impedance at antiresonance and the tank tuned to antiresonance, there is a fairly simple trial procedure which will locate the path of operation. With reference to Fig. 2, the dc plate and grid voltages locate the point C. The peak rf grid drive then locates the horizontal line $e_c = E_{gm} + E_{cc}$, on which point A must lie. Point A is one extreme point, while point C is the midpoint of the path of operation. A position is chosen on the line $e_c = E_{gm} + E_{cc}$ for point A. Fig. 2 shows three such choices. The resulting straight line path of operation permits an accurate determination of the plate current pulse whose fundamental component can be determined by a Fourier analysis. The resulting fundamental component of plate current multiplied by the antiresonant tank impedance should equal the rf tank voltage which is the horizontal projection of the line $A-C$. If the current impedance product does not equal the horizontal distance between points A and C, then a new point A should be chosen either to the right or left of the old point A, depending on whether the current impedance product is less or greater than the horizontal distance between points A and C.

A few trial investigations in the manner outlined permits a very close approximation of the actual path of operation. That the equilibrium position of point A found by this method represents a stable position, and is the only possible position which will satisfy the given conditions, can be seen from the fact that as point A moves to the left both the angle of conduction of plate current and the value of the peak plate current decrease. Both of these factors cause a reduction in the fundamental component of plate current impedance product. But, as point A moves to the left, the increased horizontal projection of the line $A-C$ requires a larger tank voltage which means a larger current impedance product. Exactly the reverse of the above occurs as point A moves to the right. This means that the one position of point A for which the current impedance product equals the horizontal projection of the line $A-C$ is a stable equilibrium point, and the line $A-C$ thus found must be the path of operation.

The Fourier analysis of the plate current pulse, referred to above, can be performed either by a point-by-point graphical analysis using the actual constant current curves of the tube, or by an analytical analysis which assumes that the constant current curves are equispaced parallel straight lines.[2] The second method is much simpler, since under these conditions the plate current pulse reduces to a clipped sine wave. The error introduced by this assumption is generally small. Special tools such as the "Sarabacher Calculator" greatly simplify the work when a graphical analysis is made.[3] However, when the plate current pulse is narrow, as was frequently the case in this study, the "Sarbacher Calculator" gives very inaccurate results. Under these conditions an accurate graphical analysis becomes very tedious. The analytical method was therefore used in this study.

Still referring only to an antiresonant plate tank circuit, when the dc plate and grid voltages are specified then point C is specified. The rf grid drive specifies the projection on a vertical of the line $A-C$. The projection on a horizontal of the line $A-C$ equals the rf tank voltage and this must equal the current impedance product resulting from the Fourier analysis of the plate current pulse which in turn is specified by the path of operation —line $A-C$. These are the only conditions which must be satisfied for stable equilibrium of the system. Note

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[2] "Applied Electronics," E. E. Staff of Massachusetts Institute of Technology.

[3] R. I. Sarbacher, "A mechanical device for calculation of class D and C amplifier performance," *Electronics*, vol. 1; December, 1942.

that none of the above conditions restrict the horizontal projection of the line A–C to a value smaller than the dc plate voltage. When the point A is in the region of negative plate voltage the plate current is in the form of a double pulse, for each cycle, as shown in Fig. 3(b).

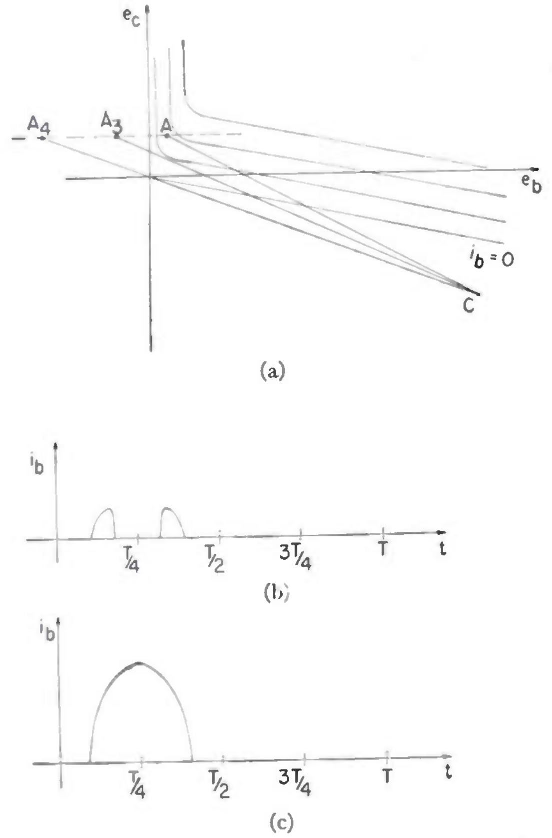


Fig. 3—(a) Paths of operation with the plate tank circuit tuned to antiresonance. (b) Plate current pulse when the path of operation is the line A_3C. (c) Plate current pulse when the path of operation is the line AC.

The corresponding path of operation is the line A_3C in Fig. 3(a). Due to the zero axis symmetry of this double pulse the fundamental component of the plate current is still in phase with the tank voltage.

With point C fixed and with the projection on the vertical of the line A–C fixed, as point A moves to the left the current decreases while the tank voltage increases; hence, the tank impedance must increase. The very large drop in fundamental plate current which results when point A just moves to the left of the zero plate voltage line results in a required large increase in tank impedance to support this condition. Under most conditions in practice, these large tank impedances are not encountered. For example, with two type 211 tubes in parallel, with a dc plate voltage of 1,000 volts, a grid bias of 150 volts, and a grid drive of 200 volts, it was necessary to use a tank impedance of 304,000 ohms in order to obtain a peak tank voltage of 1,250 volts. The normal tank impedance under these conditions of bias and drive would be about 5,000 ohms, and this would result in a peak tank voltage of about 950 volts. Fig. 3(a) shows the load lines under these conditions. Line A_3–C is for the 304,000-ohm tank, and line A–C is for the 5,000-ohm tank.

The maximum possible tank voltage under the conditions of an antiresonant tank may be found readily. As the tank impedance increases, the plate current decreases, the point A moves to the left on the constant current curves, and the rf tank voltage increases. The limiting condition occurs when the path of operation goes through the origin because under this condition there would be no plate current flow. This limiting path of operation is given by the line A_4–C in Fig. 3(a). The resulting value of tank voltage is given by:

$$E_{pm\,\max} = E_{bb}\frac{E_{gm}}{|E_{cc}|} \tag{1}$$

where $E_{pm\,\max}$ equals the maximum possible peak value of the rf tank voltage under antiresonant conditions. An infinite tank impedance is necessary to support this condition since the plate current is zero.

TANK CIRCUIT DETUNED

The condition cited above must be modified somewhat when the plate tank is detuned from antiresonance. When the ac plate and grid voltages are 180° out of phase, the path of operation becomes a straight line, and therefore the plate current pulse has zero axis symmetry which forces the fundamental component of plate current to be in phase with the tank voltage. This condition is impossible when the tank circuit is detuned from antiresonance. Therefore, with the detuned plate tank, the ac plate and grid voltages cannot be 180° out of phase and thus the path of operation on the constant current curves becomes an ellipse. The projection on the horizontal of this ellipse equals twice the peak plate voltage, while the projection on the vertical equals twice the peak rf grid drive voltage. The fundamental component of the plate current pulse multiplied by the tank circuit impedance must equal the rf plate voltage. This last condition of course, involves two conditions—magnitude and phase.

The determination of the path of operation for a given plate tank circuit and grid drive becomes much more involved when the tank is detuned than when the tank is tuned to antiresonance. This is due to the fact that now two trial guesses must be made before the path of operation can be located: (1) Magnitude of the peak rf plate voltage, and (2) the phase angle between the rf plate and grid voltages. From the assumed path of operation we can determine the plate current pulse, which can be Fourier analyzed to give the magnitude of the fundamental plate current component and its phase relationship with the rf tank voltage. The ratio of the rf tank voltage to the fundamental current component is the tank impedance value required by the assumed path of operation. If this calculated impedance is not equal to the given tank impedance, a new path of operation must be chosen. However, it is difficult to determine whether the new path of operation should differ from the original path by virtue of a different magnitude of rf plate voltage, or a different angle between plate and grid voltages, or a combination of the two. This is because both of these factors influence both the magnitude and angle of the calculated impedance. No simple relationship exists between the angle of the tank impedance and the angle between the rf grid and plate voltages.

The inverse problem of finding the tank impedance for a given path of operation may be solved as readily as was the problem of the antiresonant tank condition. This problem is discussed further in the next section.

As in the case for the antiresonant tank condition, it is quite possible to satisfy the conditions for equilibrium of the elliptical path of operation (tank detuned) with the rf tank voltage greater than the dc plate voltage. The paths of operation, and the corresponding current pulses, in two such cases, are shown in Fig. 4. Also shown in Fig. 4 is the path of operation for an antiresonant tank with the same value of tank voltage as for the detuned tank. It can be seen from Fig. 4 that, for the same rf tank voltage, an elliptical path of operation makes possible a much larger plate current pulse than that given by the straight line path of operation. This increase in plate current reduces the magnitude of tank impedance necessary to make the rf tank voltage larger than the dc plate voltage. Further calculations will demonstrate that this decrease in tank impedance is larger than the normal decrease in impedance of a tank circuit as it is detuned. This means that the required impedance of the tank at antiresonance is, in order to produce a certain tank voltage with the tank detuned, less than the impedance of a tank required to produce the same tank voltage at antiresonance. This conclusion is made obvious by Fig. 6(a), where the straight line path of operation requires an infinite tank impedance while a finite tank impedance can produce the elliptical path of operation shown.

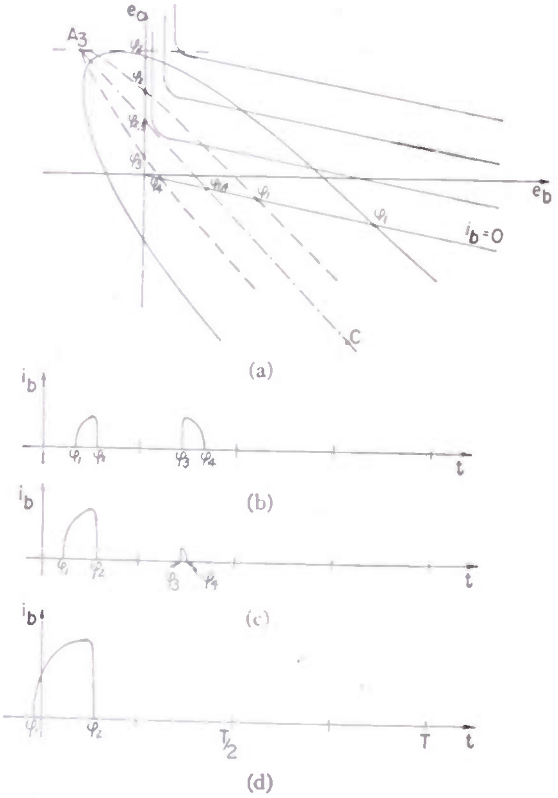


Fig. 4—Paths of operation and plate current wave shapes with $E_{pm} > E_{bb}$ for different values of phase shift between the rf plate and grid voltages. (a) Solid line is for $\theta = 30°$; dashed line is for $\theta = 10°$; dash-dot line is for $\theta = 0°$. (b) $\theta = 0°$. (c) $\theta = 10°$. (d) $\theta = 30°$.

CALCULATION OF TANK IMPEDANCE

For assumed values of dc plate voltage, dc grid voltage, rf grid drive, rf plate voltage, and phase angle between the rf plate and grid voltages, the calculation of the tank impedance proceeds as follows:

If we take time, $t = 0$, at the point where the sine wave rf tank voltage goes through zero with positive slope then:

$$e_b = E_{bb} - E_{pm} \sin \omega t \tag{2}$$

and

$$e_c = E_{cc} + E_{gm} \sin (\omega t + \theta). \tag{3}$$

For the tube:

$$i_b = \frac{e_b + \mu e_c}{r_p} \tag{4}$$

$$\therefore i_b = \frac{1}{r_p} \left\{ [E_{bb} + \mu E_{cc}] - \sqrt{A^2 + B^2} \sin (\omega t - \xi) \right\} \tag{5}$$

where (Fig. 5)

$$A = E_{pm} - \mu E_{gm} \cos \theta \tag{6}$$

$$B = \mu E_{gm} \sin \theta \tag{7}$$

$$\xi = \tan^{-1} \frac{B}{A}. \tag{8}$$

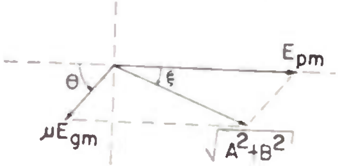


Fig. 5

This expression for i_b only holds as long as both i_b and e_b are positive. Under any other conditions, $i_b = 0$.

The assumption of a constant μ and r_p is identical with the assumption that the constant current curves are equispaced straight lines. This assumption somewhat modifies the shape of the plate current pulse mostly in that it assumes that the plate current suddenly falls to zero as the plate voltage goes negative.

In order to evaluate the fundamental sine and cosine terms of i_b, we must find the limits of the angle of conduction of plate current. For example, in considering the ellipse in Fig. 4 for which $\theta = 30°$, plate current flows from $\omega t = \phi_1$ to $\omega t = \phi_2$. ϕ_1 is given by the intersection of the path of operation with the $i_b = 0$ line for the tube. Therefore in equation (5) set $\omega t = \phi_1$, $i_b = 0$ and solve for ϕ_1

$$\therefore \quad \phi_1 = \xi + \sin^{-1} \frac{E_{bb} + \mu E_{cc}}{\sqrt{A^2 + B^2}} \cdot \tag{9}$$

ϕ_2 is given by the intersection lf the path of operation with the $e_b = 0$ line. Therefore in (2) set $\omega t = \phi_2$, $e_b = 0$ and solve for ϕ_2

$$\therefore \quad \phi_2 = \sin^{-1} \frac{E_{bb}}{E_{pm}} \cdot \tag{10}$$

Thus:

$$\alpha_1 = \frac{1}{\pi} \int_{\phi_1}^{\phi_2} i_b \sin \omega t d(\omega t) \tag{11}$$

$$\alpha_1 = \frac{1}{\pi r_p} \Bigg[-(E_{bb} + \mu E_{cc})(\cos \phi_2 - \cos \phi_1)$$

$$+ \frac{\sqrt{A^2 + B^2} \cos \xi}{2} \left(-\phi_2 + \phi_1 + \frac{\sin 2\phi_2 - \sin 2\phi_1}{2} \right)$$

$$+ \frac{\sqrt{A^2 + B^2} \sin \xi}{2} (\sin^2 \phi_2 - \sin^2 \phi_1) \Bigg] \cdot \tag{12}$$

Similarly:

$$\beta_1 = \frac{1}{\pi r_p} \Bigg[(E_{bb} + \mu E_{cc})(\sin \phi_2 - \sin \phi_1)$$

$$+ \frac{\sqrt{A^2 + B^2} \sin \xi}{2} \left(\phi_2 - \phi_1 + \frac{\sin 2\phi_2 - \sin 2\phi_1}{2} \right)$$

$$- \frac{\sqrt{A^2 + B^2} \cos \xi}{2} (\sin^2 \phi_2 - \sin^2 \phi_1) \Bigg] \cdot \tag{13}$$

If, as in the case of the ellipse in Fig. 4 for which $\theta = 10°$, the plate current is a double pulse per cycle, then it is also necessary to evaluate ϕ_3 and ϕ_4 and then the integrals for α_1 and β_1, reduce to the above plus an identical set of integrals between the limits ϕ_3 to ϕ_4. In any case,

$$i_{p_1} = \alpha_1 \sin \omega t + \beta_1 \cos \omega t$$

$$= \sqrt{\alpha_1^2 + \beta_1^2} \sin \left(\omega t + \tan^{-1} \frac{\beta_1}{\alpha_1} \right) \tag{14}$$

$$I_{pm_1} = \sqrt{\alpha_1^2 + \beta_1^2} \Big/ \tan^{-1} \frac{\beta_1}{\alpha_1} \tag{15}$$

$$Z_p = \frac{E_{pm}}{I_{pm_1}} = \frac{E_{pm}}{\sqrt{\alpha_1^2 + \beta_1^2}} \Big/ -\tan^{-1} \frac{\beta_1}{\alpha_1} \cdot \tag{16}$$

This is the required tank circuit impedance for equilibrium of the assumed path of operation.

For given values of E_{pm}, E_{bb}, E_{cc}, and E_{gm} all values of θ are not generally possible. Consider first only positive values of θ. If

$$E_{pm} > E_{bb} \frac{E_{gm}}{|E_{cc}|},$$

then $\theta = 0$ is clearly impossible. This is the case of the straight line path of operation shown in Fig. 6(a). As θ is increased the ellipse "opens up." The minimum possible value of θ is that for which the ellipse intersects the $i_b = 0$ curve and thus causes a plate current pulse. As θ is increased further, the magnitude and angle of conduction of the plate current pulse increases. A typical possible operating condition is shown by the ellipse in Fig. 6(a) for which $\theta = 20°$. The resulting plate current pulse is shown in Fig. 6(c). The Fourier analysis of this pulse will show that the fundamental component of plate current leads the tank voltage, which means that the tank impedance is capacitive. As can be seen from Fig. 6, further increases in the value of θ will shift the plate current pulse, shown in Fig. 6(c), to the left. This causes an increase in the capacitive angle of the tank impedance. The limiting value of θ occurs when the α_1 term of the Fourier analysis becomes zero. For this condition the tank becomes a pure capacitive reactance. A value of θ larger than this will cause the α_1 term to become negative, which means that the tank impedance must have a negative resistance component, which is of course impossible.

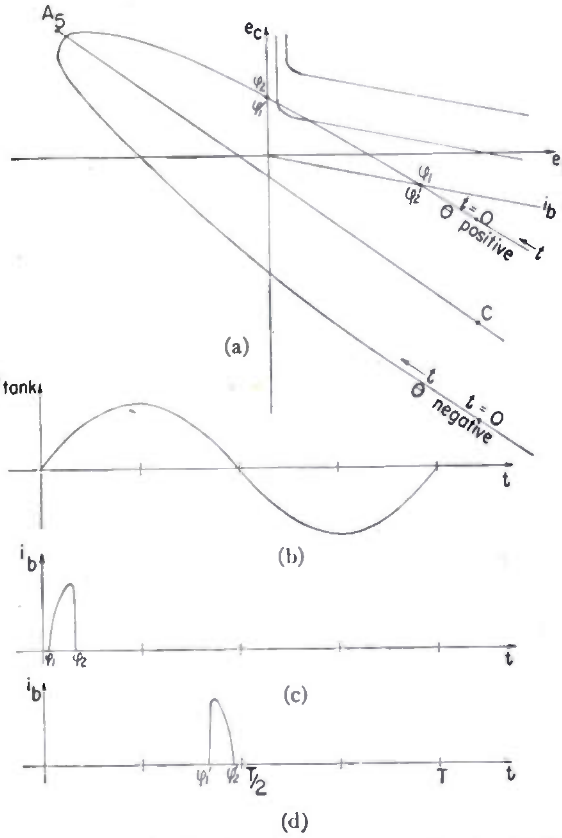


Fig. 6—(a) Path of operation with $E_{pm} > E_{bb}$ and $\theta = \pm 20°$. (b) Rf plate voltage for $\theta = \pm 20°$. (c) Plate current wave shape for $\theta = +20°$. (d) Plate current wave shape for $\theta = -20°$.

For a negative value of θ, the elliptical path of operation is identical with that given for the same positive value of θ, except that for negative values of θ the operating point moves clockwise around the path of operation instead of counterclockwise as for positive values of θ. Hence the ellipse shown in Fig. 6(a) is also for $\theta = -20°$. The resulting plate current pulse is shown in Fig. 6(d). The plate current pulse is now identical to that in Fig. 6(c), except that positive time is in the opposite direction.

It can be shown that the magnitude of the tank impedance and the magnitude of the phase angle of the tank impedance is only a function of the magnitude of θ and not its polarity. Only the sign of the phase angle of the tank impedance changes with the sign of θ. Positive values of θ result in capacitive reactance .tanks, while negative values of θ result in inductive reactance tanks. The minimum and maximum physically possible values of θ for negative θ are thus identical with those for positive θ.

DETERMINATION OF MAXIMUM TANK VOLTAGE

For given values of E_{bb}, E_{cc}, E_{gm} and for a given plate tank circuit, the method for evaluating the rf plate voltage for different conditions of tank circuit tuning is somewhat analogous to the method of evaluating the rf plate voltage when the tank is adjusted to antiresonance.

Assume a value of E_{pm} and θ. This determines the path of operation. Calculate the fundamental sine and cosine terms of the plate current pulse. Calculate the required tank circuit impedance for this assumed value of E_{pm} and θ. From the magnitude and phase angle of this impedance, calculate what the impedance of this tank would be if it were adjusted to antiresonance. The following expression for Z_r can be used, for this last calculation, when the tank is tuned by varying the tank circuit inductance (as was the case in the experimental work done in this investigation):

$$Z_r = \frac{|Z_p| + X_c \sin \psi}{\cos \psi}. \tag{17}$$

This formula assumes high-Q and constant coil resistance. The problem of variations in this resistance with the setting of the tank inductance is handled in the next section. A similar expression can be derived for the case where the tank is tuned by varying the tank circuit condenser.

The above group of calculations is to be repeated for different assumed values of θ but for the same assumed E_{pm}.

From this series of calculations we can then plot required tank impedance at antiresonance versus phase angle of the tank impedance for constant rf tank voltage across the detuned tank (Z_r versus ψ for constant E_{pm}).

All of the above is to be repeated for different assumed values of E_{pm}.

For two type 211 tubes operating in parallel, with $E_{bb} = 1{,}000$ volts, $E_{cc} = -150$ volts, $E_{gm} = 200$ volts, and a plate tank circuit condenser fixed at a reactance value of 1,000 ohms, the calculated curves of Z_r versus ψ for values of E_{pm} equal to 1,250, 1,500, 1,750 and 2,000 volts are shown in Fig. 7. The value of Z_r is plotted on a logarithmic scale for convenience in handling its large magnitude range. The full line curves are for negative values of ψ (capacitive tank), while the dashed line curves are for positive values of ψ (inductive tank).

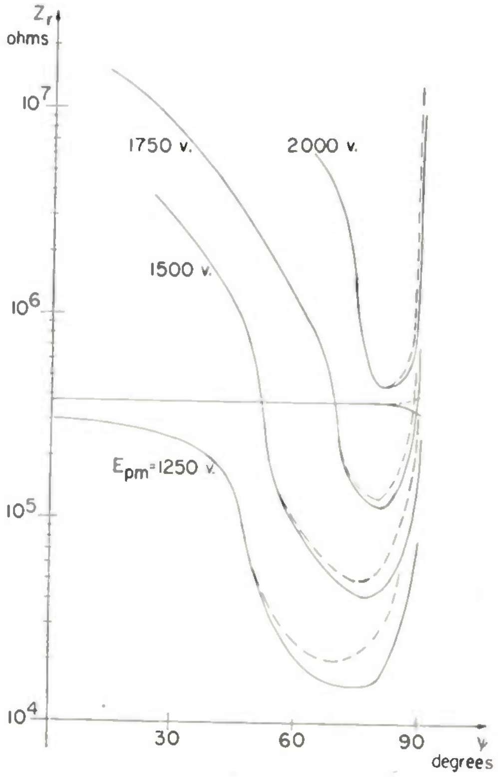


Fig. 7—Magnitude of the plate tank impedance (Z_r) versus the angle of the plate tank impedance (ψ) for constant E_{pm}. This is for two 211 tubes operating in parallel with: $E_{bb} = 1{,}000$ volts, $E_{cc} = -150$ volts, $E_{gm} = 200$ volts, $X_c = 1{,}000$ ohms, $f = 500$ kc. These curves are based on an assumed effective $\mu = 12$ and $r_p = 1{,}100$ ohms. The solid lines are for negative ψ while the dashed lines are for positive ψ. The almost horizontal line is the apparent Z_r for the experimental plate tank circuit.

For constant E_{pm} as ψ increases from zero the required Z_r decreases. This is due to the fact that as ψ increases, θ increases and the ellipse "opens up," which causes an increase in the fundamental component of plate current, which of course reduces the required tank impedance. But for large values of ψ (ψ close to 90°) the ratio of Z_r/Z_p for the tank circuit becomes very large ($Z_r/Z_p = \infty$ for $\psi = 90°$). For large ψ this factor more than compensates for the reduction in Z_p, due to the increase in plate current and causes the value of Z_r to increase with increasing values of ψ.

If on this family of curves a horizontal line is drawn at Z_r equal to the antiresonant impedance of the given plate tank circuit, it is then possible to predict the variations in the rf plate tank voltage with the plate tank circuit tuning.

It is obvious from these curves that increasing the impedance of the tank circuit will increase the rf plate voltage.

Since

$$Z_r = \frac{|Z_p|}{\cos\psi} + X_c \tan\psi$$

the vertical height between the dashed and full line curves for the same E_{pm} is due to the $X_c \tan\psi$ term. As the fixed value of X_c is increased the two curves move apart, and the required Z_r, to produce a given tank voltage, is smaller for negative ψ than for positive ψ. Therefore the maximum possible tank voltage will increase when X_c is increased, even though the impedance of the actual tank circuit at antiresonance is kept fixed by decreasing the Q of the circuit as X_c is increased. This maximum tank voltage will occur when the coil is adjusted to detune the tank capacitively.

Increasing the dc plate voltage or the grid driving voltage or decreasing the dc grid bias will increase the plate current pulse and will therefore increase the maximum possible rf plate voltage for a given tank circuit. Similarly, of course, increase the μ or decreasing the r_p of the tube will increase the rf tank voltage.

EXPERIMENTAL VERIFICATION

The predicted relationship between the rf plate voltage and the plate tank circuit tuning, as obtained above, lends itself very readily to experimental verification.

In order to minimize construction details, a model 8010 transmitter (manufactured by the Radiomarine Corporation of America) was modified somewhat to obtain rf tank voltages higher than the dc plate voltage. This transmitter has an 807 crystal oscillator, an 807 buffer, and two type 211 tubes operating in parallel as the power output stage. The oscillator was set at a frequency of 500 kc.

The only essential modification of the transmitter was the replacement of the plate tank circuit of the final stage by a very-high-Q tank circuit. This tank circuit consisted of a fixed 150-$\mu\mu$f mica condenser and a variable coil. The stator of this coil consisted of 48 turns of #8/16/36 Litz, wound on a form 10.5 inches long and 7 inches outside diameter. The rotor consisted of 13.5 turns of #8/16/36 Litz, wound on a form 2.75 inches long and 5.75 inches outside diameter. At 500 kc the Q of the coil varied from about 405 to 320, depending upon the rotor setting.

Condenser voltage dividers were placed across the plate tank circuit and the grid input circuit of the final power amplifier to permit a cathode-ray oscilloscope to be used to observe the path of operation on the constant current curves (e_g versus e_p). In addition, a small resistor (0.43 ohm) was placed in series with the plate tank circuit to permit a cathode-ray oscilloscope to be used for observation of the plate current pulse.

The point of antiresonance of the plate tank circuit was determined by the scope indication of the path of operation. As explained previously, this can only be a straight line when the tank circuit is antiresonant. As can be seen from the plate current curve in Fig. 9 this antiresonance setting also resulted in a minimum dc plate current. This is similar to what occurs in class-C amplifiers under normal operating conditions ($E_{pm} < E_{bb}$). From this setting of the plate tank coil, and the calibration curves of the coil, it was then possible to determine the total fixed capacity of the tank circuit (including stray capacity, output capacity of the power amplifier, capacity of the plate circuit divider for the scope, and input capacity to the peak reading voltmeter in the plate circuit). The calculated fixed capacitive reactance in the plate tank circuit was 1,000 ohms, and $Z_r = 370{,}000$ ohms.

Had the resistance of the coil been constant then a horizontal line, at $Z_r = 370{,}000$ ohms, drawn on the Z_r versus ψ curves (for constant E_{pm}) would have permitted a determination of the variation in tank voltage with tank tuning. However since the resistance of the tank coil is not constant and since the Z_r versus ψ curves were calculated on the basis of a tank coil with constant resistance, a slight modification of the horizontal line, referred to above, is necessary. For any setting of the tank coil we can find the L and R of the coil. This then permits us to calculate the ψ (phase angle) of the tank impedance and the value of Z_r if the coil resistance were constant at this value. Let us refer to this value of Z_r as the apparent Z_r of the tank circuit.

The resulting curve of apparent Z_r versus ψ for the actual tank circuit is plotted on the Z_r versus ψ (for constant E_{pm}) curves (Fig. 7). It should be noted that the apparent Z_r versus ψ curve for the tank circuit differs only very slightly from the horizontal straight line that would have been obtained if the resistance of the coil were constant.

By interpolation between the Z_r versus ψ (for constant E_{pm}) curves, it is then possible to predict the

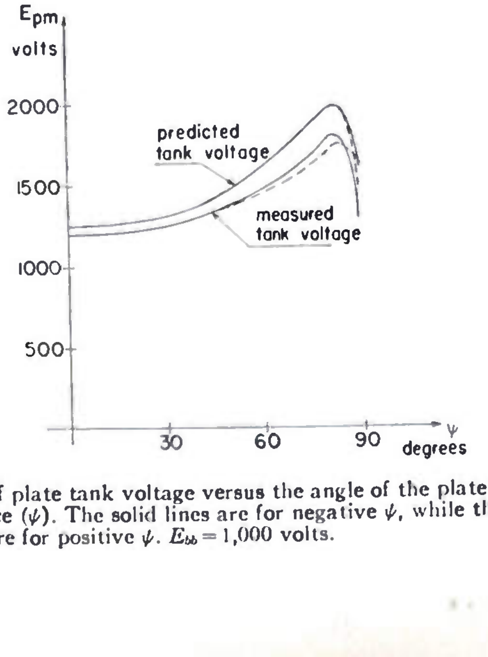


Fig. 8—Rf plate tank voltage versus the angle of the plate tank impedance (ψ). The solid lines are for negative ψ, while the dashed lines are for positive ψ. $E_{bb} = 1{,}000$ volts.

variation of tank voltage with tank circuit tuning. The resulting predicted value of E_{pm} is plotted versus ψ in Fig. 8.

Since the relationship between the setting of the tank coil and ψ (the angle of the tank impedance) has already been calculated, it is possible to plot the predicted value of tank voltage versus the tank coil dial reading. This curve is shown in Fig. 9.

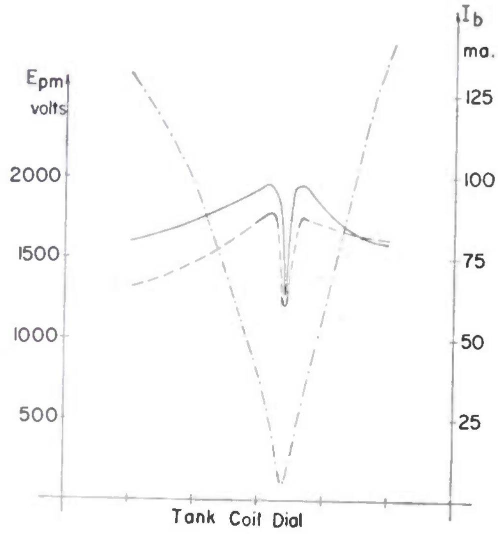


Fig 9—Predicted plate tank voltage (solid line), measured plate tank voltage (dashed line), and measured dc plate current (dash-dot line) versus the plate tank coil dial setting (arbitrary units). E_{bb} = 1 000 volts.

Also in Fig. 9 is plotted the experimentally measured values of tank voltage versus tank coil dial setting. From the known relationship between the tank coil dial setting and ψ, the experimentally measured values of tank voltage are plotted versus ψ in Fig. 8.

The predicted tank voltages differ from the measured values, for most cases, by less than 12 per cent.

In view of the original assumptions (that the constant current curves of the tube are equispaced straight lines, whereas in reality in the region of low plate current the curves are not equispaced, and in the region of low plate voltage the curves are not straight lines, and since those two regions contribute a large percentage of the plate current pulse when the rf plate voltage is larger than the dc voltage) the agreement between the predicted and measured tank voltage is extremely good.

The above assumptions regarding the constant current curves both result in an assumed plate current pulse which in general is larger than the actual plate current pulse, and therefore result in a predicted tank voltage which is higher than the actual tank voltage. This of course could have been compensated for more completely by assuming a higher value of r_p than was done in these calculations. Or possibly a more accurate method of attack would have been to approximate each constant current curve by two straight lines of different slope—one for the region of low plate voltage and another for all remaining values of plate voltage. However the increased accuracy of this method does not justify its use in view of the large increase in work it entails.

It should be noted that at antiresonance the rf tank voltage goes through a minimum point whereas under normal operating conditions, with the rf tank voltage less than the dc plate voltage, the rf tank voltage reaches a maximum at antiresonance. This justifies the shift to the left of the minimum point of the Z_r versus ψ curves as E_{pm} is decreased. With E_{pm} less than E_b the minimum point of these curves will occur very close to $\psi=0$ (tank circuit at antiresonance).

As can be seen from the curves of tank voltage versus tank tuning, the maximum peak rf plate voltage was about 1.8 times the dc plate voltage. When the rf grid drive was increased from 200 volts to 300 volts the peak rf tank voltage was experimentally found to be about 2.5 times the dc plate voltage. Higher values of grid drive or higher values of plate tank impedance will in general cause the ratio of maximum peak rf tank voltage to dc plate voltage to be even greater than this value. Ratios greater than 5 have been obtained experimentally.

This study shows that it is inadvisable to operate the power amplifier at the nominal operating voltages or to perform the power amplifier tuning operation when the antenna is disconnected or presenting a very small load to the power amplifier. Hence there is a lower limit beyond which it is inadvisable to reduce the rf power output from a transmitter by reducing the coupling between the plate tank and the antenna.

Due to the large grid current which flows when the plate voltage is negative the use of grid leak bias will limit the rf plate voltage and will greatly reduce the required voltage rating of the tank condenser.

An alternate method of avoiding the condenser breakdown that this phenomenon produces is to use a protective gap across the plate tank condenser. This method was suggested by J. F. McDonald and has been used by him in a number of Radiomarine transmitters.

No data are at present available as to the influence of this large rf tank voltage on the life of the tube.

ACKNOWLEDGMENT

This study was undertaken as partial fulfillment of the requirements for the degree of Master of Electrical Engineering at the Polytechnic Institute of Brooklyn.

The author wishes to express his appreciation to C. A. Hachemeister for his many helpful suggestions, to H. K. Warner and J. F. McDonald for their assistance with the experimental work, and to Bernard and Julius Dwork for their helpful criticisms.

A New Wide-Range, High-Frequency Oscillator*

O. HEIL† AND J. J. EBERS†, ASSOCIATE, IRE

Summary—**A new tube is described which is fundamentally a Barkhausen oscillator with an electron gun and a cavity resonator. The tube is inherently inefficient, but is capable of tuning over extremely wide frequency ranges. For one single tube, for instance, the wavelength ranges from 4.5 to 12 cm and the output power from 0.1 to 1.0 watt.**

SHORT-WAVE GENERATION in a tube is an energy transformation process. Kinetic energy of electrons is transformed into electric field energy of the electromagnetic oscillating circuit. In extending the range of an oscillator to higher frequencies, there are two limitations set by nature: (1) The conductivity of copper, and (2) the limited emission density of cathodes. The first limitation can be moved farther out by concentrating the electric field in the smallest possible space and allowing the magnetic field the largest possible space. The second limitation can be moved farther out by concentrating the electrons from a large cathode into a narrow beam. The tube described in this paper makes use of field concentration as well as electron concentration. The field concentration is obtained by making use of the fact that electric fields concentrate on sharp corners, whereas, the electronic concentration is obtained by use of a highly efficient electron gun.

The electron mechanism used is the same as in the Barkhausen tube, and is similar to the mechanism of a reflex klystron. In spite of the low efficiency of this mechanism, tubes have been built which tune over a 3-to-1 tuning range by varying only the capacity of the circuit. This variation is possible due to the low resonator gap capacitance. The large tuning range, combined with a simple solid construction, makes the tube useful in spite of its low efficiency.

Fig. 1 shows the electron gun used which increases the current density by a factor of 230 from the cathode to the narrowest cross section of the beam. At the side is shown the distance *d* between a plane cathode and anode which give, at the same voltage, the same current density as the gun, space-charge limitation assumed. The emitting part of the cathode is an ellipsoid with an axial ratio of 1:3. The curvature of this surface is lowest at the center of the cathode where the electrons need less focusing action, since the distance they have to travel from the cathode to the narrowest part of the beam is about 15 per cent longer than the same distance for an electron coming from the edge of the cathode. Two beam elements are shown on the drawing, one coming from the center of the cathode, and the other from the edge. Besides the different lengths and the different focusing action, the defocusing effect of the space charge on the center element has to be smaller than on the outside element, to make the beam parallel in the right distance. In an ideal parallel beam all electrons travel in the same direction. The cathode surface is, therefore, specifically less loaded in the center. The given dimensions have been found experimentally. The density distribution in the beam was checked with a thin carbon sheet as a screen, and the directional distribution of the electrons was checked in a space-charge free space produced by ion trapping. The dimensions shown in Fig. 1 were the final result of a long series of experiments. The dimensions given were in millimeters in the first model. The cathode area was 4.65 cm². However, only the relative dimensions are of importance since the characteristics, such as the increase in current density of 230 and the total space-charge limited current flowing (140 ma at 1,000 volts), are independent of the actual size. The dimensions used in the tubes described in this paper are the original size and half size.[1]

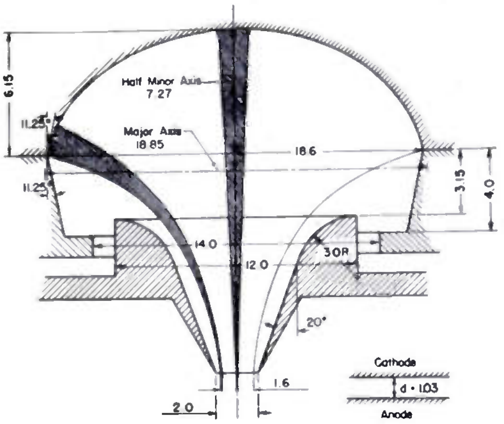
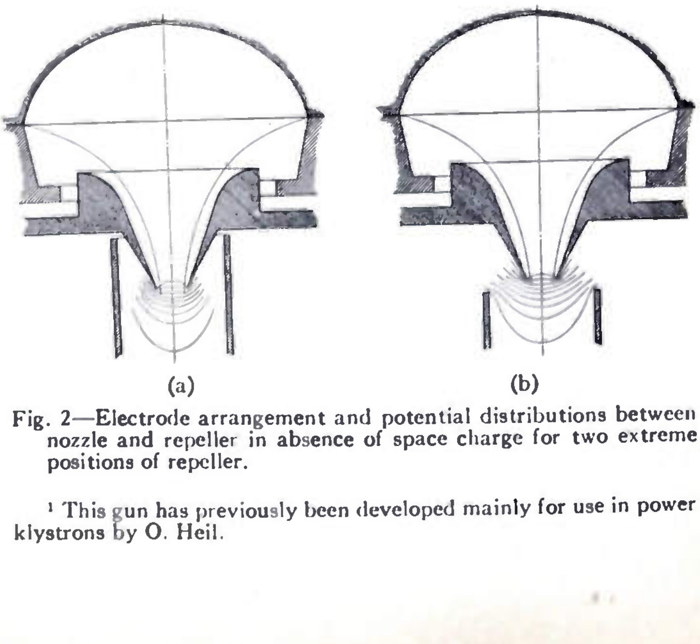


Fig. 1—Electron gun.



(a) (b)

Fig. 2—Electrode arrangement and potential distributions between nozzle and repeller in absence of space charge for two extreme positions of repeller.

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* Decimal classification: R355.912. Original manuscript received by the Institute, October 24, 1949; revised manuscript received, February 16, 1950. The research in this paper was sponsored by the Air Materiel Command, Wright Field, Dayton, Ohio.

† Electron Tube Laboratory, The Ohio State University, Columbus, Ohio.

[1] This gun has previously been developed mainly for use in power klystrons by O. Heil.

Fig. 2 shows how the concentration of the electric field is obtained at the point where the electron beam has its greatest density. The equipotential surfaces surrounding the conic nozzle are pictured. Their form is mainly determined by the nozzle shape, and depends little on the position of the tubular tuner electrode. The equipotential surfaces were taken in an electrolytic trough and do not take into account the space charge of the beam. (For space-charge effect compare Fig. 4.)

Fig. 3 shows in curve 1 the potential plotted along the axis of Fig. 2(a). This potential shape is of double interest, first, as picturing the ac field and second, as picturing the dc or repeller field. For the ac or high-frequency field, the high concentration of the field near the nozzle and the weak field farther away is important. An electron going out and coming back through these fields experiences a strong ac effect going, that is, the modulation effect, and a strong ac effect coming back, which is the working effect. In between, or during the bunching time, the ac effect is weak. (The same point is mentioned by Pierce in a comparison between a Barkhausen tube and a reflex klystron.[2]) The same field, as a dc or repeller field which has to transform velocity variations into bunches, has good transforming ability. In Fig. 3 negative potential is shown positive in order that electrons can be pictured as balls rolling on the potential curve. Curve 2 is a parabola; such a potential curve can never transform velocity variations into bunches because the time an electron takes to go in and come out of the repeller field is independent of its velocity. The electron makes just half an oscillation of an ideal pendulum. If the potential distribution deviates from the parabola, the bunching is stronger the larger the deviation. Curves below the parabola, like curve 3, give bunching similar to a linear klystron while the controlling field is rising; whereas curves above the parabola give bunching similar to most reflex klystrons[3] while the controlling field is falling. The transformation of velocity variations into density variations is good because of the strong deviation of curve 1 from the parabola. These considerations show that the potential distribution obtained in the electrode arrangement of Fig. 2 is favorable for controlling bunching and working of electrons.

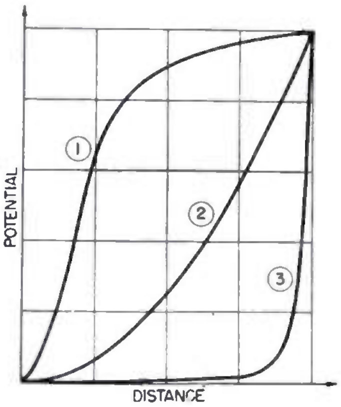

Fig. 3—Axial potential distributions in repeller space.

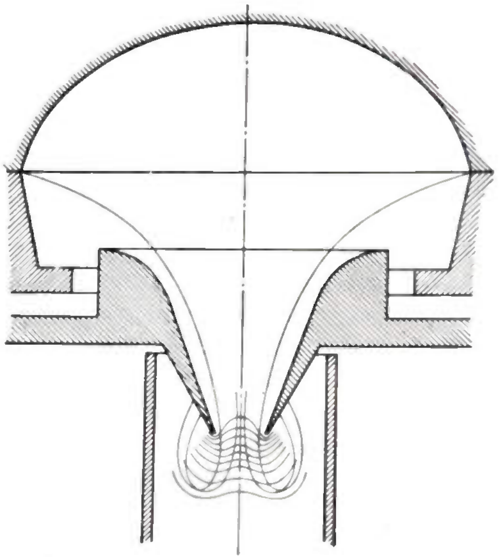

Fig. 4—Potential distribution between nozzle and repeller considering space charge. Shows effect on returning beam.

Apart from the greater simplicity and tunability of this tube, there are three more advantages:

1. All the electronic energy which has to be dissipated is produced on one electrode, the nozzle electrode, which is made of solid copper and conducts the heat away to the outside envelope of the tube.
2. No secondary emission or multipactor discharge loads the oscillating cavity, because the dc field holds all secondaries back on the nozzle and the repeller is never struck by electrons in normal operation.
3. The tube shows no hysteresis effects due to electrons returning in the cathode space, because the nozzle opening is very small and the beam is, by its own space charge, split into a tubular beam on the return. (Fountain effect.) Fig. 4 illustrates this. The distortion of the equipotential surfaces of Fig. 2 by the beam space charge is shown. The path of the electrons is indicated.

Fig. 5—Experimental evidence of formation of ring-shaped returning beam.

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[2] Discussion on "Reflex oscillators," by J. R. Pierce (February, 1945, pp. 112–118); E. U. Condon, A. E. Harrison, W. W. Hansen, J. R. Woodyard, and J. R. Pierce., Proc. I.R.E., pp. 483–485; July, 1945.

[3] An exception is the wide-range tunable reflex klystron described by Ludwig Meyer in an unpublished work.

Experimental evidence of this path is given in Fig. 5. The nozzle was covered with soot. The finely divided carbon glows where it is struck by electrons. The picture was taken through the repeller tubing and shows the nozzle opening in the center surrounded by a dark zone where scarcely any electrons strike the nozzle. Outside of this dark zone is the ring where most of the electrons strike. The irregularity in brightness is due to variations in the soot. When the tube oscillates, the ring becomes a little more diffuse. In tuning the tube, the ring moves up and down on the nozzle. At the high-frequency end of the tuning range the repeller tube has not sufficient refocusing effect on the electrons, and the ring spreads over the base of the nozzle. If this point is reached, the tube ceases to oscillate. The electronic efficiency seems to be higher with the repeller close to the nozzle, that is, on the long wavelength end of the tuning range. With the higher capacitive loading of the circuit at this end, more power is lost in the circuit, and therefore the observed power output does not vary much over the whole tuning range. (It may be mentioned that the soot used for detecting the electrons did not appreciably increase the cavity losses. The splitting of the electron beam into a ring is responsible for the fact that no hysteresis has been observed (see Fig. 13).) There are, however, some electrons out of the center part of the beam which return to the cathode space, but their effect is small because of the relatively long and nonuniform transit time in the cathode space.

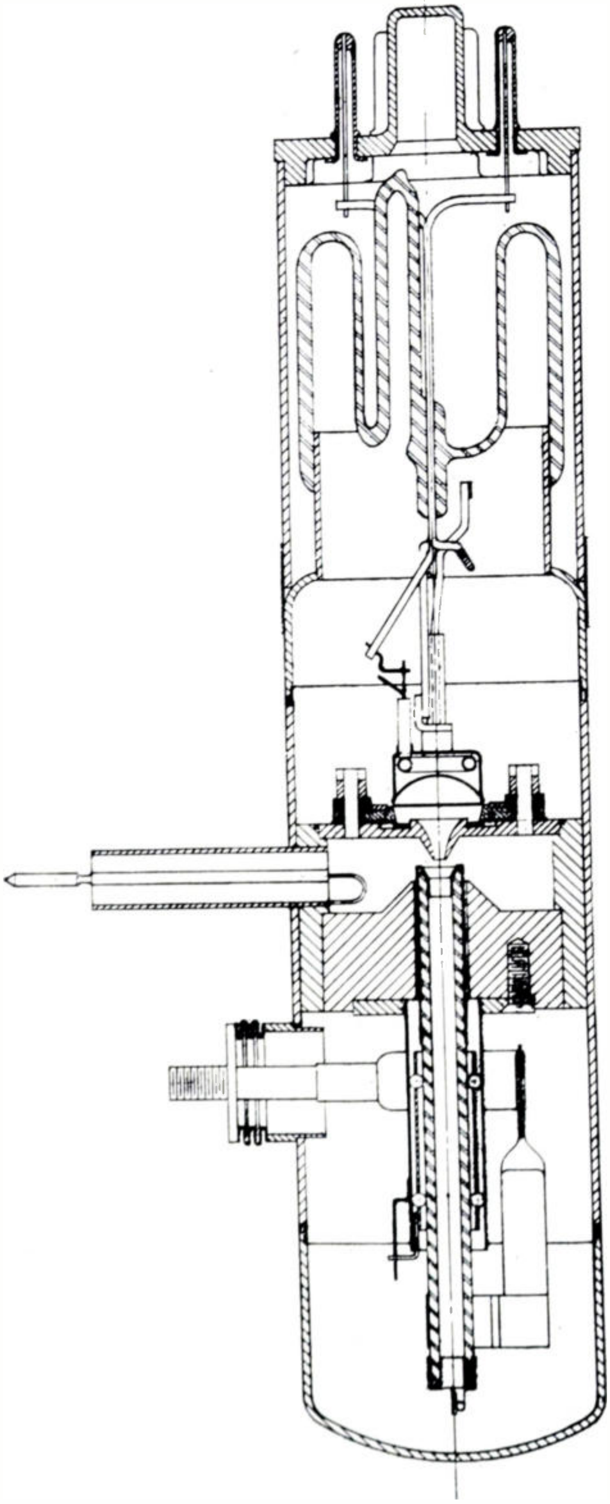

Fig. 6—Tube cross section.

In Fig. 6 is shown a cross section through a sealed-off model of the tube and in Fig. 7 the external view. The cavity is made of copper and is brazed into a standard size radio tube steel envelope. The cathode assembly consisting of hot and cold parts is screwed to the cavity and insulated by a mica disk and ceramic insulators. The mechanism for guiding the repeller tube is screwed to the other side of the cavity. The repeller, having negative potential with respect to the cavity, must move freely without touching the copper block of the cavity. The motion is guided by two sets of three steel balls which roll on the inside of a thin-walled tubing. The elasticity of the metal tubing holds the balls tight onto the repeller and prevents any shaking. The balls are spaced by a ball container. External motion is transferred into the vacuum by a lever mechanism through a short piece of metal bellows. The turning point of the lever is at the bellows. The total motion of the repeller for the full tuning range is about 2.5 mm. The tube shown in Fig. 6 tunes from the 4.2- to 14-cm. wavelength.

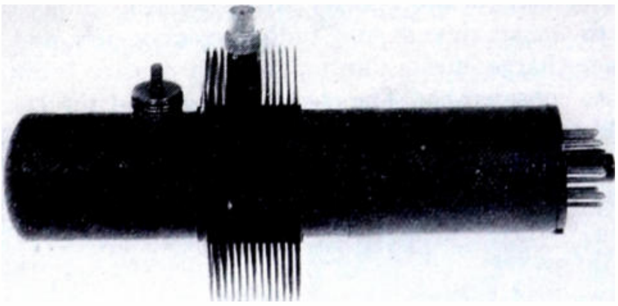

Fig. 7—Tube.

As the cavity is not perfectly closed, high-frequency energy may leak out through the narrow gap between the insulated repeller and the cavity body. This leaking of energy becomes very large, should there exist a resonance in the back volume of the tube. In this case, it can suppress the oscillations. In the tube it is prevented in the following way:

1. The repeller is made of glass or ceramics, and the end sticking into the cavity is silverplated inside and outside.
2. The length of the plating is made shorter than half the shortest wavelength produced so that it cannot resonate.
3. This silverplating is connected electrically over a high resistance (about 10,000 ohms) with silverplating at the other end of the repeller tubing where the repeller connection is made. The resistance is made by coating the inside of the glass or ceramic tubing with carbon (dag).

The tube shown in Figs. 6 and 7 is the model which has been brought into a sealed-off form. Many other versions of the tube varying in electrode shape, in cavity and gun size, and in tuning range and frequency have been studied. All these studies were made on the pump, where quick changes are possible and many variations could be made on the same model. The experimental method of studying this electron mechanism is by far the best, since the electron mechanism is rather complex due to the fact that the electron motion has to be considered in three-dimensional space. The radial components of the velocity are of the same importance as to transit time as longitudinal components, and the space-charge effects due to the high density beam are quite considerable. The electron beam and the electric field configuration, as determined by nozzle and repeller, have been varied in these experiments. On the electron beam, the current as well as the convergence of the electrons at the point where they enter the high-frequency field was varied. The current variation was obtained by moving the cathode farther away from the accelerating electrode. This change of the gun does not affect the quality of the beam appreciably because the beam formation is mainly determined by cathode shape. To move the high-frequency field from the parallel section of the beam into the converging or diverging section, the nozzle was built longer or shorter without changing the outside shape. The best efficiency was obtained by the beam leaving the nozzle parallel and by a spacing of cathode to nozzle which was slightly greater than shown in Fig. 1. The output maximum is very flat, which means the current and the convergence of the beam are not critical. The motion of the cathode in this experiment was done continuously with the tube oscillating by a lever mechanism similar to the one used on the repeller. The varying of the field configuration was done mainly by variation of the nozzle. In one experiment the field was made less homogeneous by adding a cylindrical portion on the end of the nozzle. In other experiments the nozzle opening was made larger or the cone angle was varied. All these experiments showed a decrease in efficiency when deviating from the geometry given in Figs. 2 and 4. On the other hand, it was shown that the dimensioning of the nozzle is not very critical for this low efficient mechanism which is adaptable over a very great frequency range. On one tube, the frequency range produced, with the same exciting unit, was further extended in addition to the normal repeller tuning by changing the size of the cavity. The wavelength varied from 4.75 to 18.8 cm. The actual size of the gun was the same as in Fig. 1, dimensions given in millimeters; whereas, the gun used in the tube shown in Figs. 6 and 7 has half the linear dimensions.

Figs. 8, 9, 10, and 11 present a few characteristics which were obtained with different models of the tube.

Fig. 8 shows the wavelength versus turns of tuning mechanism characteristic of one model of the tube. In this and the following figures, the turns of the tuning mechanism is a linear measure of the distance between nozzle and repeller. This model used a full-size gun with a two-inch diameter cavity. It is seen that one repeller

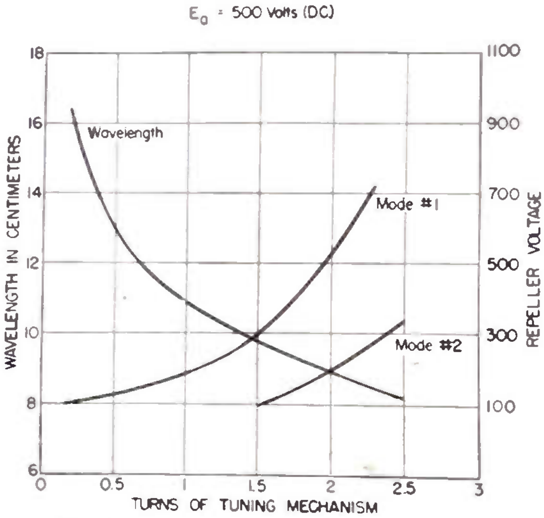

Fig. 8—Tuning characteristics, 8 to 16 cm.

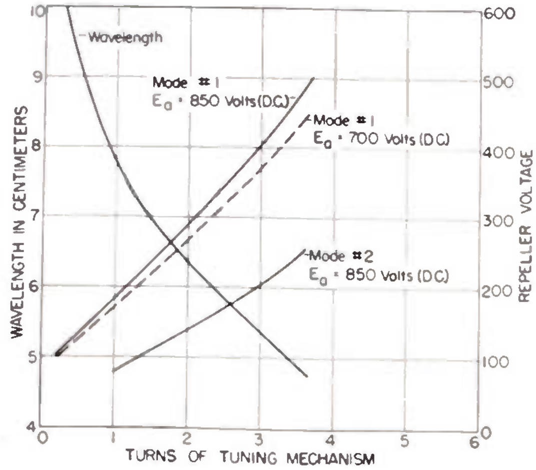

Fig. 9—Tuning characteristics, 5 to 10 cm.

mode covers almost the entire range of the tube. This tube delivered 200 to 250 milliwatts of radio-frequency power across the tuning range at an efficiency of approximately one per cent.

Fig. 9 shows the characteristics of another model similar to the one whose characteristics are shown in Fig. 8, with the exception that the cavity diameter was decreased from two inches to one inch. This tube will oscillate over the range from five to ten centimeters. Two repeller modes are shown: the tube will oscillate over the entire range without shifting modes, and the change in repeller voltage is practically linear with change in repeller position. This tube delivered between 100 and 200 milliwatts across the tuning range.

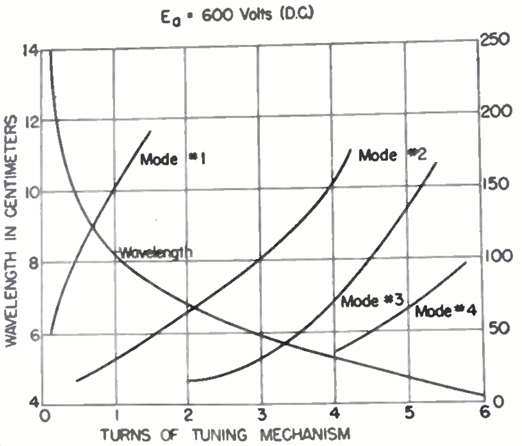

Fig. 10—Tuning characteristics, 4.5 to 12 cm.

The results shown in Fig. 10 were obtained from the tube shown in Figs. 6 and 7. This tube has a half-size gun and a cavity which is about one inch in diameter. It will oscillate over the range from 4.2 to 14 cm. with an output of 100 to 150 milliwatts. In a sealed-off tube, however, the range would have to be limited to about 12 cm, because of danger of field emission between nozzle and repeller. This tube has the disadvantage that no single repeller mode can be used across the entire range. For this tube, mode 1 is the 3/4 mode in conventional reflex klystrons, mode 2 is the 7/4 mode, and so forth.

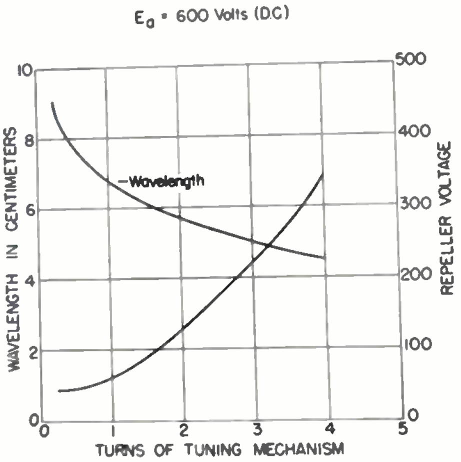

Fig. 11—Tuning characteristics, 4.5 to 9 cm.

Fig. 11 shows the characteristics of a tube similar to the one shown in Figs. 6 and 7, except that the inside and outside diameters of the repeller have been increased. The result is a decrease in the tuning range, but now the whole range can be covered by one repeller mode. The power output of this tube is comparable with the tube whose data are given in Fig. 10.

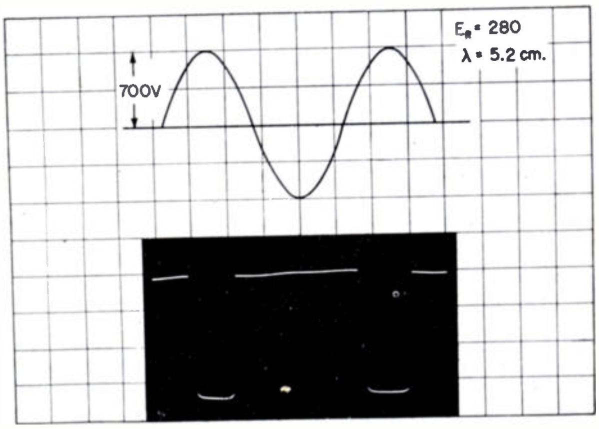

Fig. 12—Variation of power output with anode voltage.

Fig. 12 is an oscillogram of relative power output with varying anode voltage for the tube whose characteristics are shown in Fig. 9. The repeller voltage was adjusted to give maximum output. As is always true, the anode voltage is not critical.

The lack of hysteresis in power output as repeller voltage is varied as shown in Fig. 13. The anode was operated at a constant voltage, and a fixed bias plus sinusoidal modulation was applied to the repeller. The picture on the left shows output with increasing and decreasing repeller voltage. The one on the right shows the two outputs superimposed. The small difference is due to phase shift in the scope amplifier, and practically no hysteresis exists.

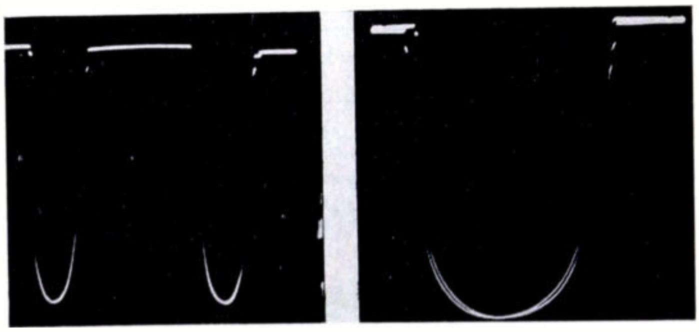

Fig. 13—Variation of power output with repeller voltage showing freedom from hysteresis.

$E_A = 790$ volts direct current

$E_R = \begin{cases} 110 \text{ volts direct current} \\ 120 \text{ volts alternating current} \end{cases}$

$\lambda = 5.3$ cm.

Conclusion

A tube giving a power of 100 to 200 mw and being tunable from 4.2 to 12 cm has many applications because it replaces a series of tubes. It may be used as a laboratory oscillator, for antenna measurements, or as a local oscillator in a receiver. The output and tuning range is achieved with a specific load on the cathode surface of only 40 ma per square centimeter. This assures long tube life and little difficulty in manufacturing. Moreover, it shows that cathode emission is not a limitation when scaling the tube to shorter wavelengths. The other limitation in scaling is the heat dissipation ability of the nozzle. Experiments in this direction show that this limit has not been reached. One of the standard tubes was operated at 1,100 volts and gave 1 watt output with an input power of 84 watts. At this load the tube was run continuously for 24 hours without any change in behavior. In view of the simple construction of the tube, it is therefore believed that scaling to 15,000 Mc is possible.

ACKNOWLEDGMENTS

The authors wish to thank their associates in The Ohio State University Electron Tube Laboratory for their co-operation which has made this research possible. Particular credit is due to Robert W. Wilmarth who has contributed to the mechanical design, to Peter Whibley for the glass work, and to Stanley Taylor, John Cowan, and the men in the electrical engineering machine-shop. Some electrical measurements were made by Richard Neubauer and by Miss S. C. Chiao. We also thank Mrs. Jeannette Reynolds and John Dankworth for the preparation of the figures.

The Influence of the Ground on the Calibration and Use of VHF Field-Intensity Meters*

F. M. GREENE†, MEMBER, IRE

ONE SOURCE OF error usually present in very-high-frequency field-intensity measurements (30 to 300 Mc) results from the effect of the ground beneath the receiving antenna on the value of its radiation impedance. This impedance together with the load impedance Z_L connected to the center terminals determines the ratio of the terminal voltage to the induced voltage in the antenna, namely, the *voltage-transfer ratio*.[1] This in turn affects the value of the *antenna constant*.

Very-high-frequency field-intensity measurements will be generally in error, therefore, if either the receiving antenna height or ground constants are appreciably different from those existing when the field intensity meter was calibrated.

A solution is obtained yielding the approximate radiation or input impedance Z_i of a horizontal half-wave dipole over plane homogeneous earth for finite values of dielectric constant ϵ_r, and conductivity σ. The result is

$$Z_i \simeq Z_{11} + \Gamma Z_{12}, \qquad (1)$$

where Z_{11} is the free-space input impedance,[2] Z_{12} is the mutual impedance[3] between the antenna and its image ($\sigma = \infty$), and Γ is the complex plane-wave reflection coefficient[4] of the ground for vertical incidence. The impedances are referred to the center terminals in all cases.

With the antenna input impedance evaluated, the measurement error may be determined for various values of load impedance Z_L, antenna height in wavelengths h_2/λ, or ground constants ϵ_r and σ.

If the receiving *antenna constant* was determined for a sufficient antenna height h_2 as to have essentially a "free-space" value, the resulting percentage difference between the *true* and *indicated* values of field intensity at any other antenna height will be

$$\delta = \left(\left| \frac{Z_L + Z_{11}}{Z_L + Z_i} \right| - 1 \right) \times 100 \text{ (per cent).} \qquad (2)$$

Values of δ calculated from (2) are well supported by actual measurements at one particular site at a frequency $f = 100$ Mc.

Typical calculated values of this difference or measurement error versus height in wavelengths h_2/λ are shown in Fig. 1 for values of the ground constants shown, $\sigma = \infty$, $\epsilon_r = 9$, 15, and 30. Low loss dielectrics were assumed for the latter three cases, i.e., $(\sigma/\epsilon\omega) \ll 1$. This assumption is permissible for many types of ground over a large portion of the very-high-frequency band.

As shown in Fig. 1, a field-intensity meter ($Z_L = 73\Omega$) calibrated under free-space conditions may indicate values of field intensity over average ground ($\epsilon_r = 15$) which are in error by as much as 10 per cent for values of h_2/λ near 0.3, and 7.5 per cent for values of h_2/λ near 0.6. If this error is to be held to values less than 5 per cent, antenna heights greater than about 0.65 wavelengths should be used for field-intensity measurements under these conditions.

It is somewhat doubtful at present just what maximum values of measurement error of this type should be permitted. The error can obviously be reduced by increasing the load impedance Z_L. For values of the terminating impedance $Z_L = 73$, 150, and 300 ohms, the maximum value of the error calculated for antenna heights greater than 0.15 wavelengths is 10, 7, and 4 per cent, respectively, and approaches zero as Z_L approaches infinity.

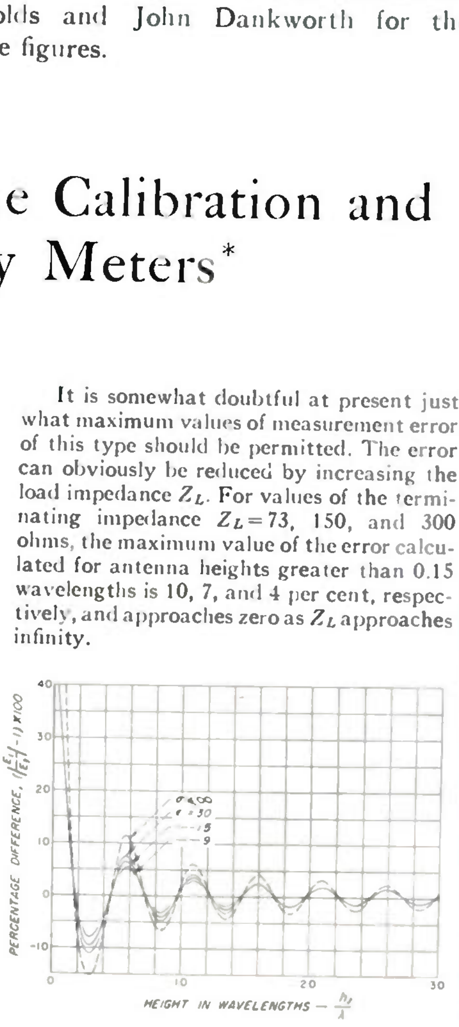

Fig. 1

Apparently the usual changes in the ground constants experienced (due to changing moisture content) have but little effect upon the measurement error as presented here. As seen from Fig. 1, the total variation from "average ground" conditions ($\epsilon_r = 15$) does not exceed 1.5 per cent except for values of $h_2/\lambda < 0.15$.

Since the error is appreciably larger over perfectly conducting ground ($\sigma = \infty$), it would seem inadvisable to use or calibrate very-high-frequency field-intensity meters over a metallic reflecting plane, unless the antenna heights were carefully chosen so as to result in a low value of error.

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* Decimal classification: R271. Original manuscript received by the Institute, June 1, 1949; abstract received, November 18, 1949. This is an abstract of a paper published in the *Journal of Research*, National Bureau of Standards, vol. 44, no. 2, February, 1950. Copies for sale by the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D.C

† Central Radio Propagation Laboratory, National Bureau of Standards, Washington 25, D. C.

[1] "Standards on Radio Wave Propagation. (Measuring Method)," Supplement to PROC. I.R.E., vol. 30, part II, p. 1; July, 1942.

[2] S. A. Schelkunoff, "Electromagnetic Waves," pp. 441–479, D. Van Nostrand Co., Inc., New York, N. Y.; 1943.

[3] P. S. Carter, "Circuit relations in radiating systems and applications to antenna problems," PROC. I.R.E., vol. 20, pp. 1004–1041; June, 1932.

[4] J. A. Stratton, "Electromagnetic Theory," p. 493, McGraw-Hill Book Co., Inc., New York, N. Y.; 1941.

A Philips-Type Ionization Gauge for Measuring of Vacuum From 10^{-7} to 10^{-1}mm. of Mercury*

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Summary—**A Philips (Penning)-type ionization gauge is described which, unlike most former gauges of this type described in he literature, is capable of measuring vacuum down to 10^{-7} mm of Hg. Construction and operation details of the gauge are given in addiion to its advantages and disadvantages as compared to hot-cathode-type ionization gauges.**

THE INCREASE in the use of high vacua has emphasized the need for a convenient means of measuring very low pressures. Among the more practical gauges for measuring high vacuum are the hot-filament and the cold-cathode ionization gauges. The former type has the obvious disadvantage of burnng out if the vacuum is broken while the gauge is in operation. The cold-cathode type is not destroyed by operation at atmospheric pressure, has a greater range and longer life, outgases less, and gives accurate pressure readings.

The Philips (Penning[1,2]) ionization gauge was selected as the most promising type for routine use. A Philips-type gauge which gave accurate pressure readings between 10^{-7} and 10^{-1} mm of mercury was then designed and constructed.[3] Figs. 1 through 4 show two models of this gauge adapted for connection to a glass vacuum system. A permanent Alnico V magnet (illustrated in Fig. 4) producing a field of 1,700 to 2,100 gauss across a 3/4-inch gap, is used to supply the magnetic field. A potential of 1,800 volts dc is supplied between the metal envelope, or cathode, and the wire loop, or anode. The current flowing through the gauge is a function of the pressure and is measured by the circuit shown in Fig. 5.

Operation of the gauge depends upon ionization of the gas molecules within the gauge. When an electron leaves one side of the metal envelope, C_1 in Fig. 1, it is attracted toward the anode loop, and its travel is constrained to a helical path of small diameter by the magnetic field. This electron passes through the anode loop into the retarding electric field caused by the opposite side of the metal envelope C_2, and it is then forced back

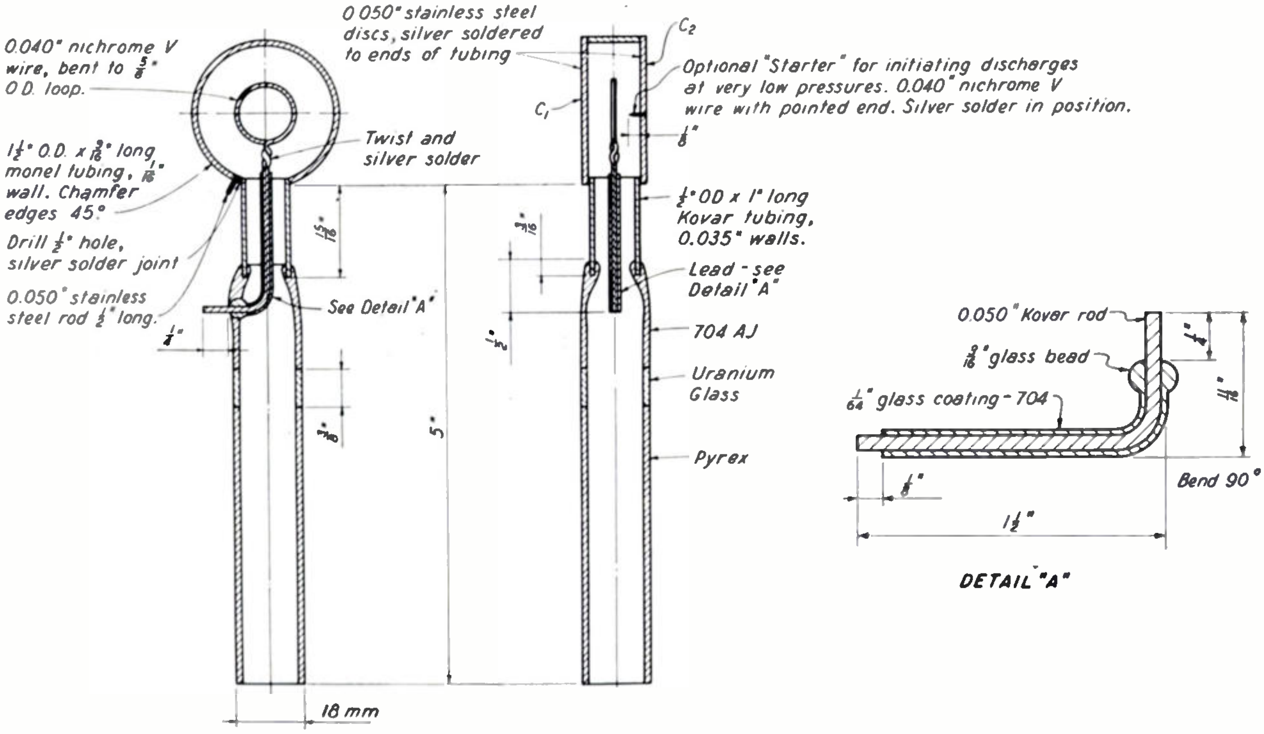

Fig. 1—Philips ion gauge.

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* Decimal classification: 621.375.621. Original manuscript received by the Institute, October 19, 1949; revised manuscript received, February 8, 1950.

This paper is based on work performed for the Atomic Energy Commission by Carbide and Carbon Chemicals Corporation, Oak Ridge, Tenn.

† Carbide and Carbon Chemicals Corporation, Oak Ridge, Tenn.

[1] F. M. Penning, "Method and Device for Measuring Pressures," U. S. Patent No. 2,197,079; April 16, 1940.

[2] F. M. Penning, "Glow discharge at low pressure between coaxial cylinders in an axial magnetic field," *Physica*, vol. 3, pp. 873–894; November, 1936; and "New manometer for low gas pressures, especially between 10^{-3} and 10^{-5} mm," *Physica*, vol. 4, pp. 71–75; February, 1937.

[3] A. O. Nier, "A mass spectrometer for isotope and gas analysis" *Rev. Sci. Instr.*, vol. 18, pp. 399–400; June, 1947.

toward the anode at a somewhat slower velocity and in a helical path of greater diameter. This cycle is repeated many times until the path of the electron extends outward to the anode loop, and the electron is then captured. During its travel, the electron will strike some of the molecules of gas which are present in the gauge and will ionize them. The positive ions will travel to the cathode plates, C_1 and C_2, while the secondary electrons will travel to the anode in the same manner as the primary electron. This current between the anode and the cathode plates is a function of the kind and number of molecules of gas present and may be used as a measure of the pressure.

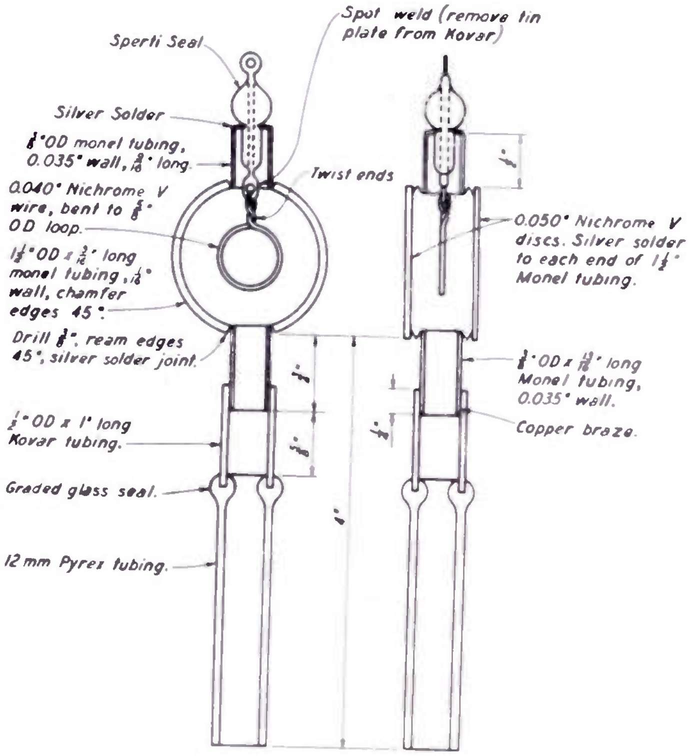

Fig. 2—Philips ion gauge.

Several units of the gauge described have been subjected to various tests and have been found to give accurate pressure readings from 10^{-7} to 10^{-1} mm of mercury. Graphs of the calibration for air are shown in Figs. 6 through 8. The nonlinearity of the curve for pressure above 5×10^{-4} mm is caused by a protective resistor in series with the gauge. Using a lower value of resistance would extend the linear region. The gauge is about six times as sensitive as common hot-filament ionization gauges. Quantities of these gauges have been assembled without special precautions, and the calibrations of the majority of the units completed are within a factor of 1.5. Several gauges have been in routine service continually for more than a year without failure or noticeable change in calibration. As with all ionization gauges, the calibration depends upon the nature of the gas present,[4] a property which makes the gauge useful as a leak detector. Spraying a suspected area with ace-

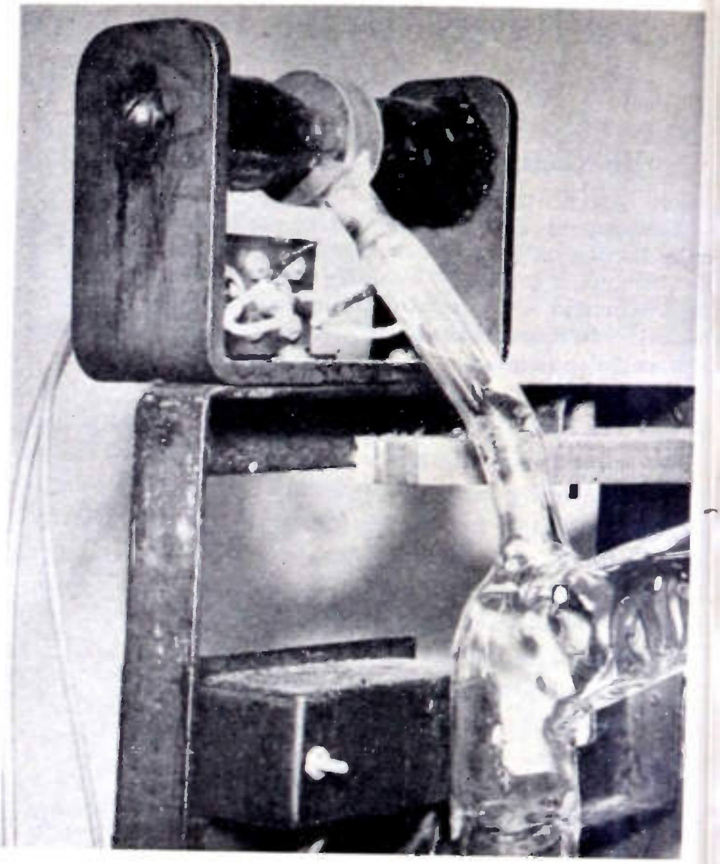

Fig. 3—Philips ion gauge connected to glass system.

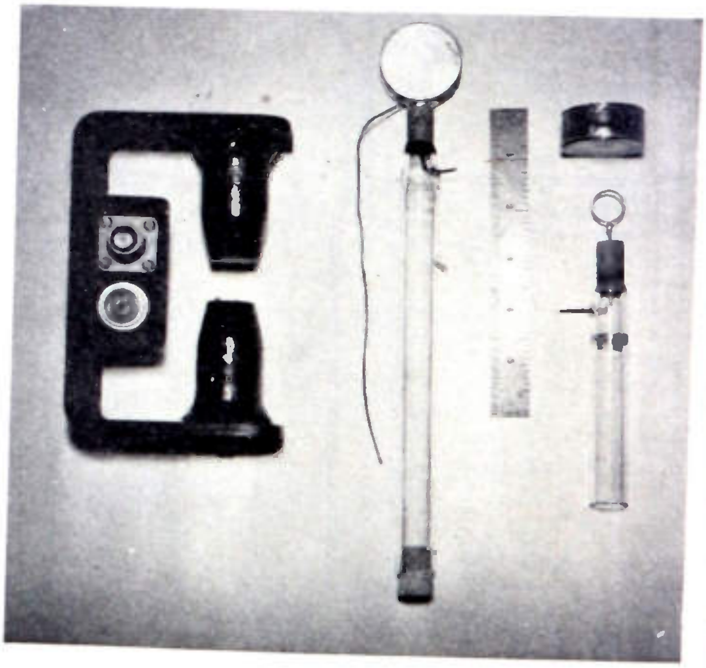

Fig. 4—Philips ion gauge and magnet.

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[4] L. R. Foote, "Effect of Different Gases on Philips Gauges," United States Atomic Energy Commission, AECD 2672. MS June 17, 1944, declassified August 23, 1949; No. 850-A14047.

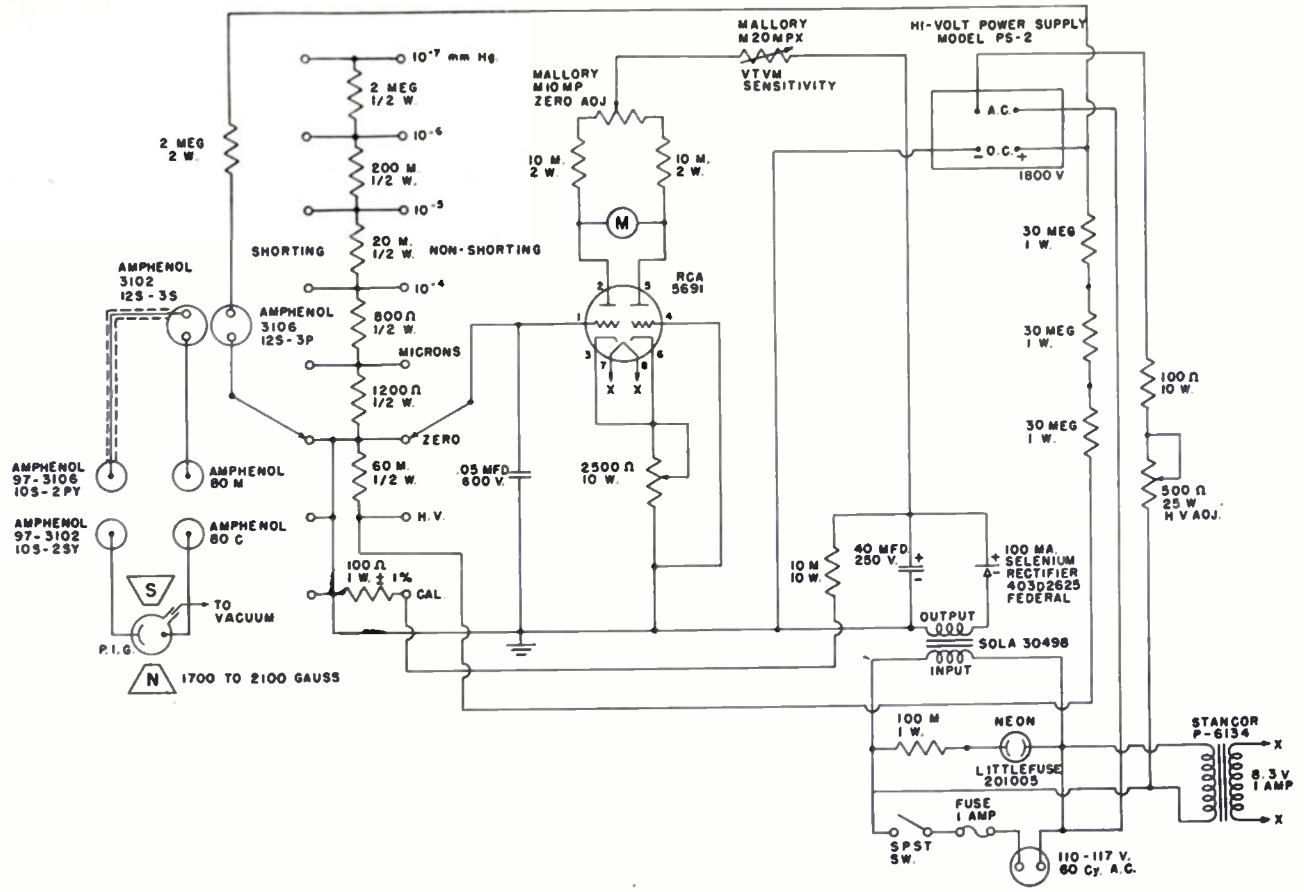
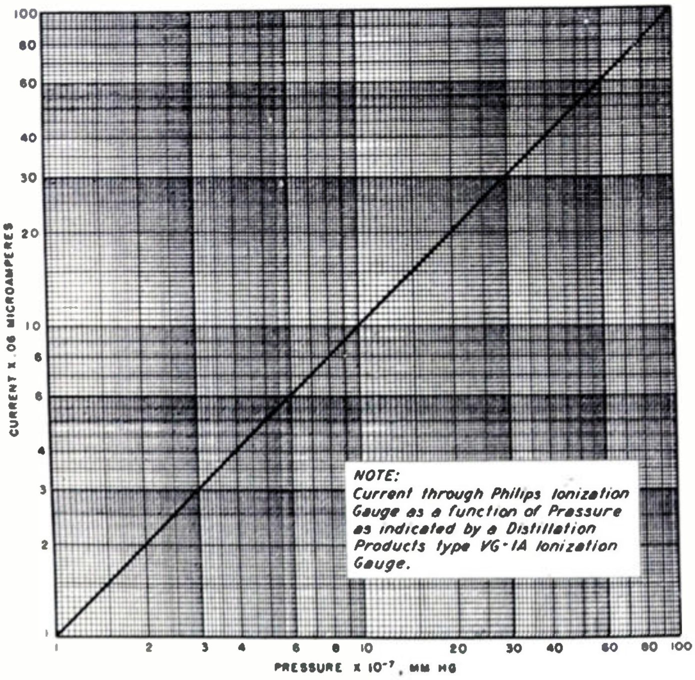


Fig. 5—Miniature Philips ion-gauge circuit. (M = G.E.—D040AY731. F.S. = 500μA—scale 0–10. Cabinet = Bud No. 995 7″×14″×8″.)



NOTE: Current through Philips Ionization Gauge as a function of Pressure as indicated by a Distillation Products type VG-1A Ionization Gauge.

Fig. 6—Calibration of Philips ion gauge.

tone, alcohol, or helium causes a sudden change in the pressure reading when any of the material enters the system through a leak. The response of the gauge to changes in pressure is very rapid. Since the gauge is operated at room temperature, there is little trouble from cracking of hydrocarbon compounds or from outgassing after the gauge has once been degassed. Usually a lower pressure can be obtained than is possible with a hot-filament gauge.

The electronic supply shown in Figs. 5 and 9 contains a high voltage supply to furnish the potential of 1,800 volts dc and a vacuum-tube voltmeter to measure the gauge current. Since all the pressure ranges (except the micron range) are essentially linear, the pressure reading may be read directly from the meter scale times the sensitivity factor. The micron scale is nonlinear and must be added to the meter scale. The two-megohm resistor in series with the gauge is installed both to protect the supply from being overloaded and to protect the operator from high-voltage shocks. Provisions are incorporated into the circuit for checking the electrical zero and the sensitivity of the vacuum-tube voltmeter and the high-voltage potential.

Some additional advantages and disadvantages of the modified Philips gauge are:[5,6]

1. It replaces both the ordinary ionization gauge and the thermocouple or pirani gauge in operating range.

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[5] R. I. Garrod and K. A. Gross, "Combined thermocouple and cold-cathode vacuum gauge," *Jour. of Sci. Instr.*, vol. 25, pp. 378–383; November, 1948.

[6] C. Hayashi, et al, "Several improvements on the 'Philip's gauge,'" *Rev. of Sci. Instr.*, vol. 20, pp. 524–526; July, 1949.

2. It does not cause thermal ionization of the gas, which complicates the pressure reading.
3. It requires no filament emission regulator.
4. Its rugged and simple construction permits easy cleaning.
5. It occasionally fails to strike if turned on when the pressure is below 10^{-6} mm mercury. However, the gauge may be "struck" by warming to raise the pressure momentarily and initiate the discharge. The addition of a "starter" as shown in Fig. 1 will enable the gauge to "strike" at pressures down to 2×10^{-8} mm or lower without the application of heat.

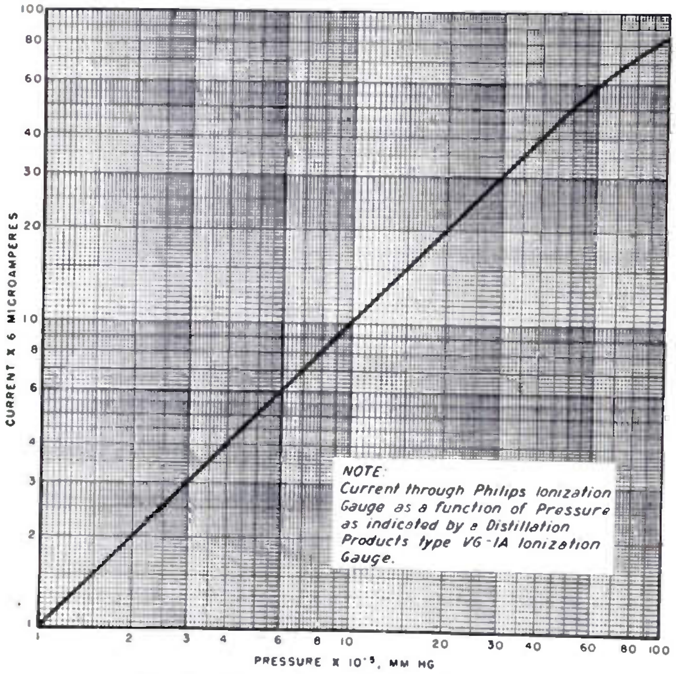

Fig. 7—Calibration of Philips ion gauge.

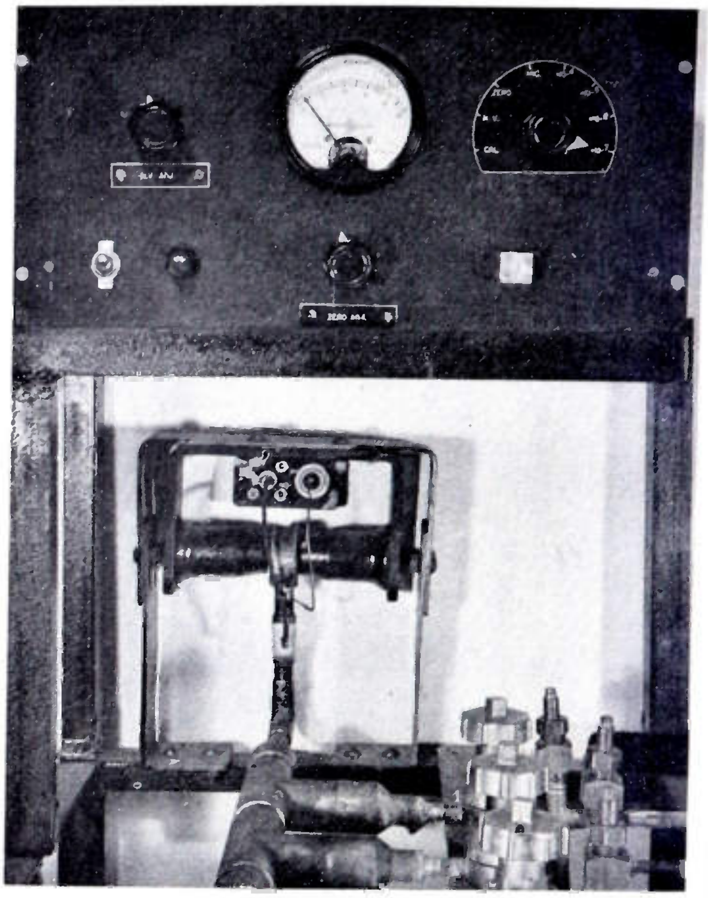

Fig. 9—Philips ion gauge and electronic supply.

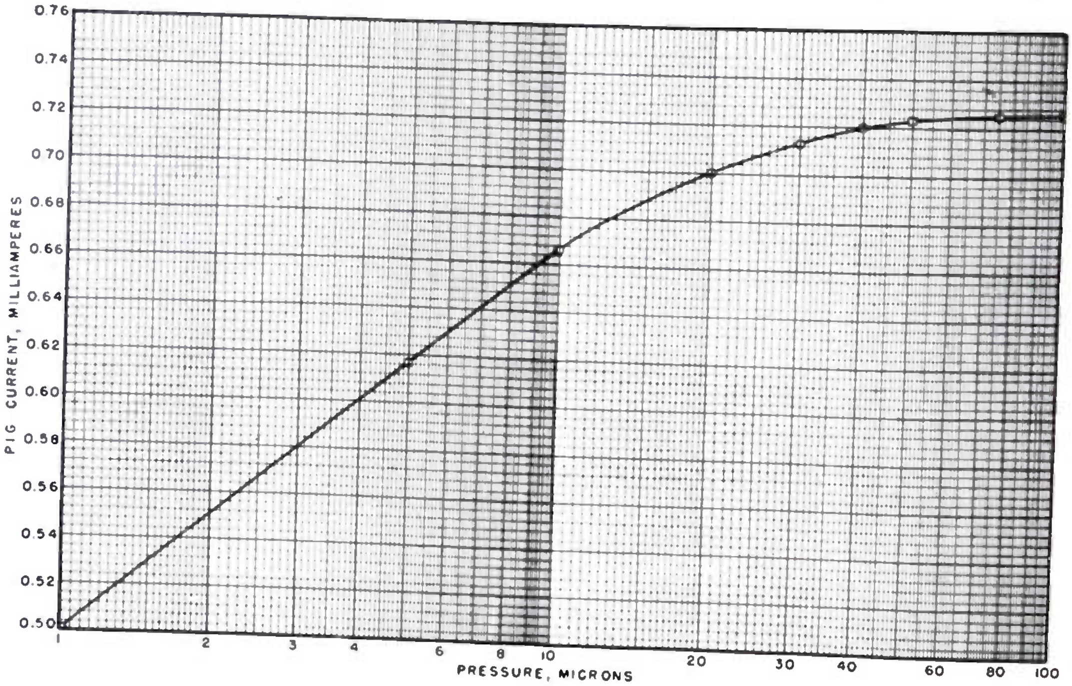

Fig. 8—Philips ion gauge current versus pressure.

Cathode-Coupled Multivibrator Operation*

KEITH GLEGG†, STUDENT, IRE

Summary—**An approximate analysis of the cathode-coupled multivibrator is given. The result is an expression for the period of the output pulse in terms of the circuit elements, rather than voltages which have to be determined graphically. A series expansion of the final expression is used to show that the output period is nearly a linear function of one of the voltages in the circuit, and can be made more nearly linear by a proper choice of the circuit parameters.**

INTRODUCTION

THE CIRCUIT to be analyzed is shown in Fig. 1. An especially fine discussion of this circuit is given in the literature.[1] In view of this discussion, no attempt will be made at a detailed treatment of the circuit operation. Briefly, however, it will be said that the circuit operates with V_2 normally conducting with nearly zero grid bias, and with V_1 cut off. When fed with a positive triggering spike of sufficient magnitude on the grid of V_1, the current switches rapidly from V_2 to V_1, by a regenerative action. V_1 then remains conducting, with V_2 cut off, for a length of time determined by the circuit parameters, at the end of which a spontaneous switching occurs, and V_2 becomes conducting with V_1 cut off as at the start. The circuit remains in this condition until triggered again. The time during which V_1 is conducting is the period of the output pulse. This pulse is positive-going on the plate of V_2 and negative-going on both the plate of V_1 and the cathodes of V_1 and V_2.

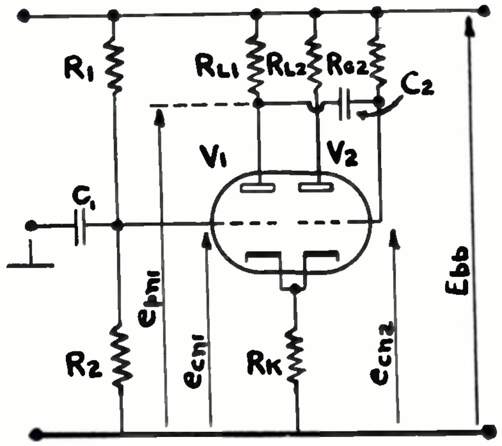

Fig. 1—Cathode-coupled multivibrator.

ANALYSIS OF CIRCUIT

It can be shown, by means of the argument given in the reference previously cited,[1] that the period of the output pulse T is given by the expression

$$T = C_2 R_{g2} \ln \left(\frac{2E_{bb} - e_{pn1}' - e_{cn2}'}{E_{bb} - i_{p1}' R_k + | e_{c02} |} \right) \tag{1}$$

where

C_2, R_{g2}, R_k, and E_{bb} are as indicated in Fig. 1
$|e_{c02}|$ = the modulus of the cutoff voltage of V_2 when V_1 is conducting
e_{pn1}' = the value of e_{pn1} when V_1 is conducting
i_{p1}' = the plate current of V_1 when it is conducting
e_{cn2}' = the value of e_{cn2} when V_2 is conducting.

The object now is to express the quantity inside the bracket in terms of the resistances in the circuit. We start with the quantity e_{c02}.

It is well known that for triodes, the cutoff grid-cathode voltage is given quite accurately by the plate supply voltage divided by the amplification factor of the tube. In the present case therefore, the cutoff voltage is given by

$$| e_{c02} | = \frac{E_{bb} - i_{p1}' R_k}{\mu}. \tag{2}$$

No subscript is used on μ, since it is assumed in all of this analysis that V_1 and V_2 are similar.

We now obtain an expression for e_{cn2}'. Since during the conduction period of V_2 its grid-cathode voltage is nearly zero, actually, very slightly positive, we can treat the plate resistance of the tube as being constant, and the plate characteristic as passing through the origin. This gives immediately

$$e_{cn2}' = E_{bb} \frac{R_k}{R_{p2} + R_k + R_{l2}}. \tag{3}$$

Further, it is clear that e_{pn1}' is given simply by

$$e_{pn1}' = E_{bb} - i_{p1}' R_{l1}. \tag{4}$$

It only remains to get an expression for i_{p1}'. By taking advantage of the strong linearity resulting from the cathode load R_k, it can be shown that i_{p1}' is given approximately by

$$i_{p1}' = E_{bb} \left[\frac{1}{R_p + (\mu + 1) R_k + R_{l1}} + \frac{R_2}{(R_1 + R_2)} \frac{\mu}{(R_p + \mu R_k + R_{l1})} \right]. \tag{5}$$

Substituting (2), (3), (4) and (5) in (1), we get

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* Decimal classification: R146.2. Original manuscript received by the Institute, October 13, 1949; revised manuscript received, February 10, 1950.

† Canadian Marconi Company, Montreal, Que, Canada.

[1] MIT Radar School Staff, "Principles of Radar," Second Ed., pp. 2–53, McGraw Hill Book Co., New York, N. Y.; 1946.

$$T = C_2R_{g2}\ln\left[\left(\frac{\mu}{\mu+1}\right)\frac{\left(1+\dfrac{R_{l1}}{R_p+(\mu+1)R_k+R_{l1}}+\dfrac{R_2}{R_1+R_2}\dfrac{\mu R_{l1}}{R_p+\mu R_k+R_{l1}}-\dfrac{R_k}{R_{p2}+R_{l2}+R_k}\right)}{1-\dfrac{R_k}{R_p+(\mu+1)R_k+R_{l1}}-\dfrac{R_2}{R_1+R_2}\dfrac{\mu R_k}{R_p+\mu R_k+R_{l1}}}\right]. \tag{6}$$

This is the required expression for T and it is in error by about ten per cent (using published values of μ and R_p) when compared with a graphical method used in conjunction with (1). We will now make use of this expression for T to show that for a wide range of conditions T is nearly a linear function of R_2/R_1+R_2 (or e_{cn1}/E_{bb}), and that by a proper choice of parameters, the approach to linearity can be improved. This property of the circuit is well known,[1,2] but a theoretical explanation has so far not appeared in the literature.

CONDITION FOR LINEAR OPERATION

After expanding the logarithm of the quotient in (6), and then making use of the series

$$\ln(1+x) = x - \frac{x^2}{2} + \frac{x^3}{3}\cdots \text{etc.},\ -1 < x < 1, \tag{7}$$

we get, on neglecting terms beyond the second order in R_2/R_1+R_2 and terms less than $1/\mu$:

$$\begin{aligned}T = C_2R_{g2}\Bigg[&\left\{-\frac{1}{\mu}+\frac{R_{l1}+R_k}{R_p+(\mu+1)R_k+R_{l1}}\right.\\&\left.-\frac{R_k}{R_{p2}+R_{l2}+R_k}-\frac{1}{2}\left(\frac{R_k}{R_{p2}+R_{l2}+R_k}\right)^2\right\}\\&+\frac{R_2}{R_1+R_2}\left\{\frac{\mu(R_{l1}+R_k)}{R_p+\mu R_k+R_{l1}}+\frac{\mu R_{l1}}{R_p+\mu R_k+R_{l1}}\right.\\&\left.\cdot\frac{R_k}{R_{p2}+R_{l2}+R_k}+\frac{\mu R_{l1}}{R_p+\mu R_k+R_{l1}}\left(\frac{R_k}{R_{p2}+R_{l2}+R_k}\right)^2\right\}\\&+\left(\frac{R_2}{R_1+R_2}\right)^2\left\{\frac{1}{2}\left(\frac{\mu R_k}{R_p+\mu R_k+R_{l1}}\right)^2\right.\\&-\frac{1}{2}\left(\frac{\mu R_{l1}}{R_p+\mu R_k+R_{l1}}\right)^2-\left(\frac{\mu R_{l1}}{R_p+\mu R_k+R_{l1}}\right)^2\\&\left.\cdot\frac{R_k}{R_{p2}+R_{l2}+R_k}\right\}\Bigg].\end{aligned} \tag{8}$$

Now it is clear from (8) that if we equate the coefficient of $(R_2)^2/(R_1+R_2)$ to zero, the resulting condition will be, to the order of approximation involved in (8), that for a linear relationship between T and R_2/R_1+R_2. When this is done the condition is found to be

$$R_{l1} = \frac{R_k}{\sqrt{1+\dfrac{2R_k}{R_{p2}+R_{l2}+R_k}}}. \tag{9}$$

It is interesting to note that careful plots of T as a function of R_2/R_1+R_2, for various values of R_{l1} using relations (1) and (6) and typical values for 6SN7, show that the value of R_{l1} satisfying (9) does give a more nearly linear relation than either $R_{l1}=R_k$ or $R_{l1}=T_k/2$. In all these cases, however, the departure from linearity is not great, and it is clear that although condition (9) is close to the ideal condition, deterioration is only very slow for variations about this condition. This can readily be understood by examination of relation (8).

An important property of the circuit, namely the rate at which T increases with respect to R_2/R_1+R_2 (or e_{cn1}/E_{bb}) when the circuit operates in the vicinity of (9), is obtained from (8) as

$$\frac{dT}{d\left(\dfrac{R_2}{R_1+R_2}\right)} = \frac{\mu C_2R_{g2}}{R_p+\mu R_k+R_{l1}}\left[R_{l1}+R_k+\frac{R_{l1}R_k}{R_{p2}+R_{l2}+R_k}+R_{l1}\left(\frac{R_k}{R_{p2}+R_{l2}+R_k}\right)^2\right]. \tag{10}$$

Range of R_2/R_1+R_2.

There are two limiting values of R_2/R_1+R_2. The upper limit is obtained when R_2 is made so large that V_2 cannot keep V_1 cut off, and free-running operation ensues. The lower limit is reached when R_2/R_1+R_2 is made so small that T in (8) would have to become negative. This condition has no physical significance, and under these circumstances the circuit cannot be triggered.

It can be shown that the range of R_2/R_1+R_2 is given approximately by

$$\frac{1}{2}\frac{R_k}{R_{p2}+R_{l2}+R_k} < \frac{R_2}{R_1+R_2} < \frac{R_k}{R_{p2}+R_{l2}+R_k}. \tag{11}$$

Here, the minimum value assumes that (9) is satisfied, and both limiting values assume that $\mu \geqq 20$, as is the case with the tubes usually employed in this circuit.

The minimum value given in (11); $(T=0)$, and the slope given in (10), allow the complete characteristics of the circuit to be obtained very rapidly as a function of R_2/R_1+R_2 (or e_{cn1}/E_{bb}).

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[2] B. V. Chance, Hughes, et al, "Waveforms," p. 168, McGraw-Hill Book Co., New York, N. Y.; 1949.

An Impulse Generator-Electronic Switch for Visual Testing of Wide-Band Networks[*]

T. R. FINCH[†], MEMBER, IRE

Summary—**The impulse generator-electronic switch is an instrument developed primarily to facilitate the design and production of radar networks such as delay, pulse-forming, and sweep networks. The instrument may be used to test any network that can be arranged to store a dc charge. Discharge characteristics of microsecond order, produced by a reference network and the network under test, are simultaneously displayed in pictorial form for comparative measurements. The first part of the paper describes representative applications and, with the aid of cathode-ray-oscilloscope traces, illustrates the simplicity of performance; the second part describes the circuit functions in detail.**

TEST FACILITIES designed for pictorial presentation of measured data are now in general use for the rapid evaluation and control of electrical components and networks. An instrument of this type has been provided to expedite the development and manufacture of broad-band delay networks and pulse-forming networks for radar application. With this instrument, any network that can be arranged to store a capacitative charge may be tested; the discharge characteristics of a reference network and the network under test are displayed, in pictorial form, simultaneously, and instantly compared.

The development of this test circuit, designated the impulse generator-electronic switch, was directed toward fulfilling two requirements: namely, to reduce the time required for production testing of wide-band networks and to facilitate the development of new networks by providing a laboratory tool that presents instantaneously and visually the network characteristics of interest. Very early in the production of pulse and delay networks it became evident that steady-state transmission measurements took an amount of time beyond reasonable bounds of available effort. An initial solution was a method of testing whereby the network's discharge characteristic was displayed on a calibrated cathode-ray-oscilloscope screen. This method was a considerable improvement over point-by-point steady-state transmission measurements because it tested directly the network performance feature of interest in about one tenth the time. However, specifying dimensional requirements on the visual pulse patterns required concentrated testing effort to insure the necessary accuracy of measurement. A greater time saving under less tedious testing conditions was realized by comparing directly the transient characteristics of a standard network with the network under test. In this direct comparison test, only a glance at the test patterns is necessary. If the two networks are identical, the resulting patterns appear as a single pattern; if the networks differ there are immediately observable differences between the two patterns.

This impulse generator-electronic switch provides (1) an impulse generator which energizes the networks under test and discharges these networks through a zero impedance switch, (2) a start-stop sweep and beam intensifier synchronized with the impulse generator that may be directly connected to the horizontal plates and grid respectively of a cathode-ray oscilloscope, (3) a wide-band signal delay so that the start-stop sweep circuit may be actuated a fraction of a microsecond before the pulse patterns appear, and (4) a switching circuit which is positively synchronized with the impulse generator and sweep circuit.

The block diagram, shown in Fig. 1, shows the functional relationship of the circuit components. The characteristic pulses developed by the impulse generator are interlaced by the gated amplifier which is phase locked to the repetition frequency of the impulse generator. The synchronized start-stop sweep circuit and beam intensifier provide the time base for the visual indicator. The impulse generator part of the apparatus, shown in Fig. 2, charges and discharges the networks under test (reference and unknown) so that the resultant network discharge voltage pulses may be compared in detail. The heart of this circuit is the switch. In order to test networks ranging in impedance from 1 to 10,000 ohms and to develop voltages of 1 to 25 volts without interaction effects, it is necessary to approach zero imped-

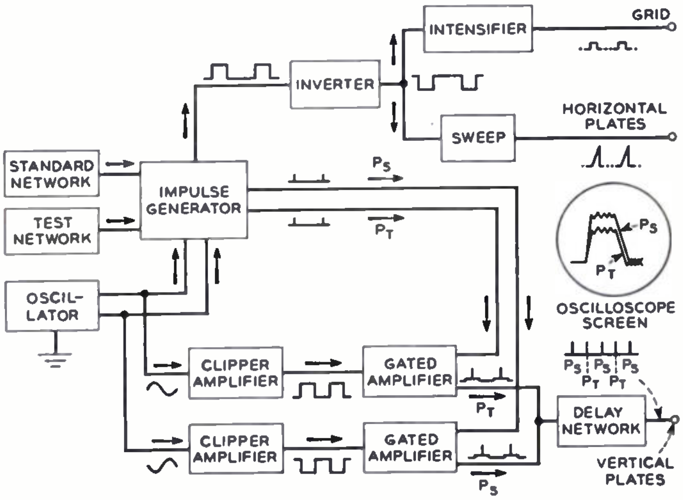

Fig. 1—Block diagram.

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[*] Decimal classification: R371.51×R143. Original manuscript received by the Institute, December 6, 1948; revised manuscript received, December 22, 1949. Presented, 1949 IRE National Convention, March 8, 1949, New York, N. Y.

[†] Bell Telephone Laboratories, Inc., Murray Hill, N. J.

ance, i.e., for this case, less than 0.05 ohm with current discharges up to five amperes. Also, the switch must operate at a sufficiently high repetition frequency so that microsecond pulses may be readily visible on screens of commercial oscilloscopes. The relay that was developed is a reed-type, wick-fed mercury relay designed to operate at 480 closures per second. It is a higher speed modification of the relay described by Brown and Pollard.[1] More recently, additional modifications and improvements have resulted in a relay, currently in use, which operates 1,000 closures per second.

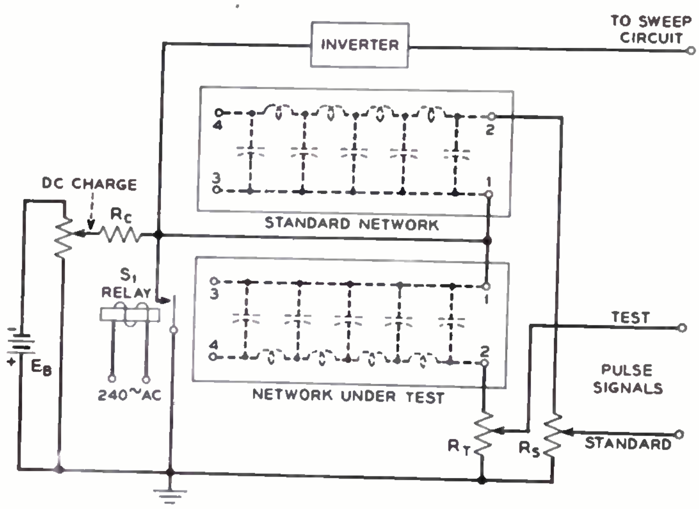

Fig. 2—Schematic diagram of the impulse generator.

The schematic for the complete apparatus is shown in Fig. 3. The top portion, *V*1, *V*2, and *V*3, is the synchronized start-stop sweep circuit. The central part is the impulse generator, and the lower portion, *V*4 through *V*9, is the switching circuit and gated amplifiers. A detailed description of the complete circuit will be given later. The material of initial interest is the description of the testing of wide-band networks, and this is most readily accomplished in pictorial form. The necessity for the signal-delay network is shown in Fig. 4. The network under test is a 1/2-microsecond, 500-ohm delay network with a bandwidth of 5 Mc. The network is connected as a two-terminal network with the output open-circuited and hence operates as a line-type pulse-forming network. The width of the discharge pulse produced is equal to twice the transmission delay, namely one microsecond for this network. Fig. 4(a) illustrates the pulse pattern produced with no signal delay. It is observed that the front edge of the pulse is lost

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[1] J. T. L. Brown and C. E. Pollard, "Mercury contact delays," *Elec. Eng.*, vol. 66, pp. 1106–1109; November, 1947.

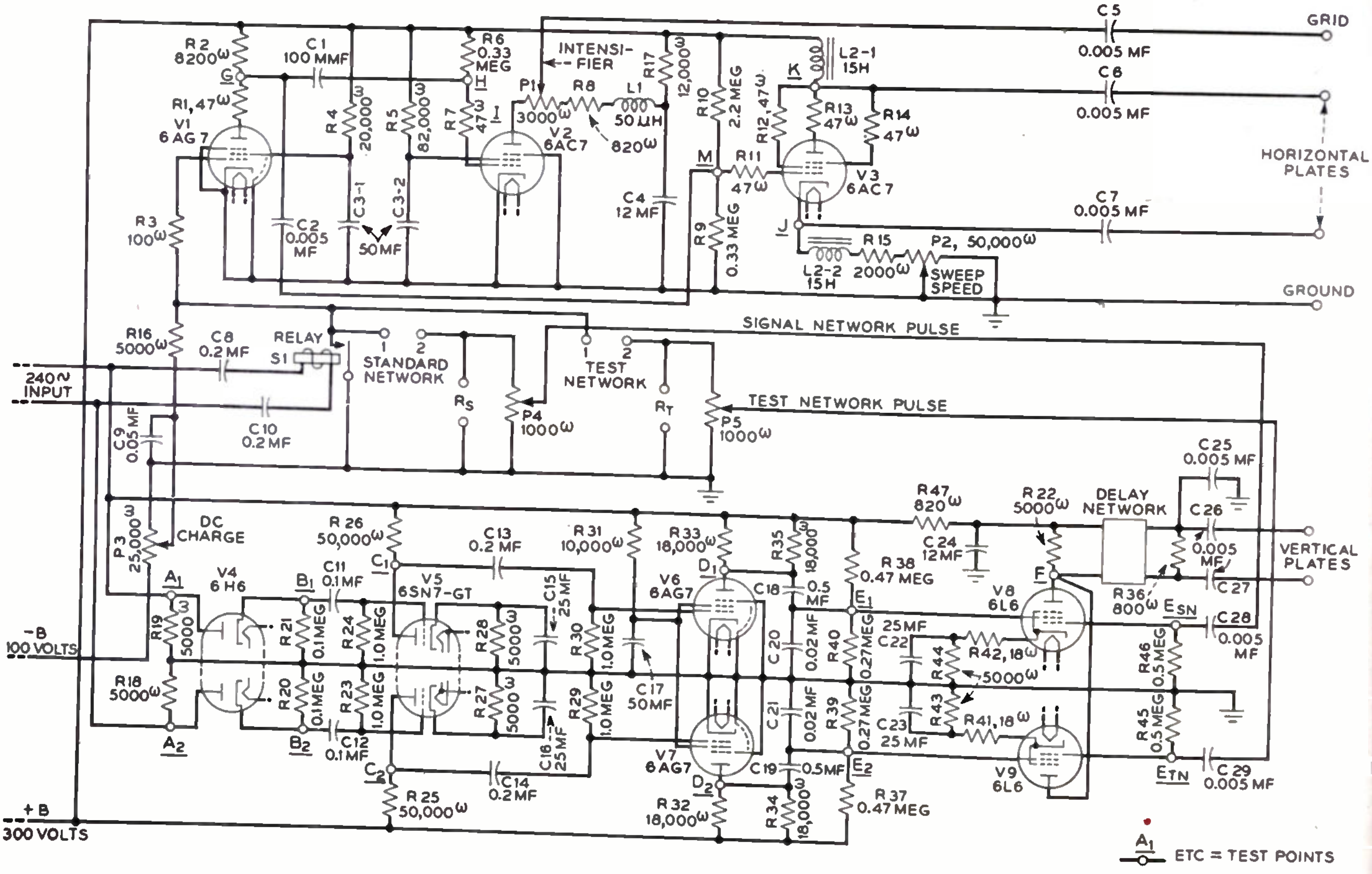


Fig. 3—Schematic diagram of the impulse generator-electronic switch.

and accurate time-of-rise (bandwidth) and pulse-length (delay) measurements are impossible. Fig. 4(b) shows the resultant pulse with signal delay, and it is seen that complete measurements may be made.

In Figs. 4(a) and 4(b) it is observed that there is a base line trace. This is because only one network is connected to the test circuit and, consequently, every other cycle no pulse voltage is produced. Thus it is always possible to tell, when comparison tests are being made by superposition, whether or not both networks under test are connected to the test circuit.

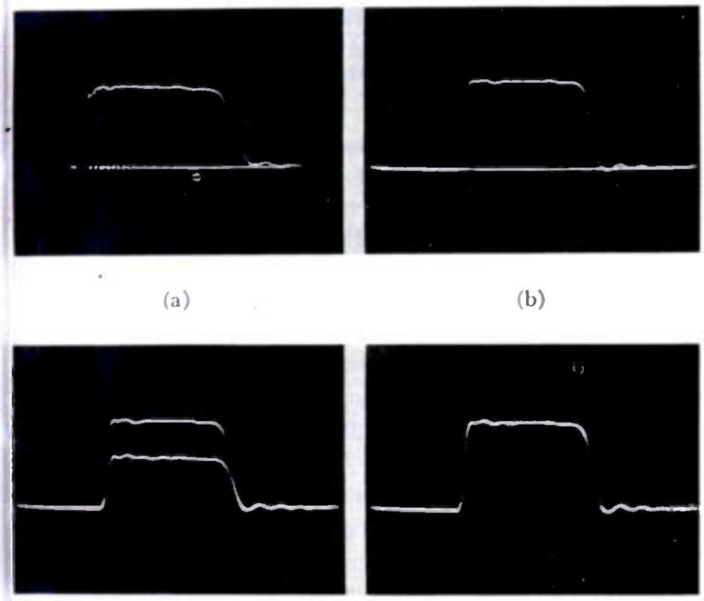

(a) (b)
(c) (d)

Fig. 4—Comparative test patterns: (a) 1-microsecond pulse, no signal delay; (b) 1-microsecond pulse, signal delay; (c) comparison of two 1-microsecond pulses; (d) superposition.

The 1-microsecond pulse patterns shown in Fig. 4(c), produced by two 1/2-microsecond delay networks, are adjusted to different voltage amplitudes by gain controls across the load resistors R_s and R_t. Fig. 4(d) illustrates superposition of the patterns, by gain control adjustment, for direct comparative measurements. It is noted that the two networks are alike in detailed response except for a difference in delay of 0.013 microsecond. This difference in delay is shown by the difference in pulse length. The trailing pulse edges are separated by 0.026 microsecond.

The comparative test patterns shown by Fig. 5 further illustrate the use of the circuit for comparative measurements. The figures are self-explanatory, but a few additional remarks may be informative. Figs. 5(c) and 5(d) show the operation of two delay networks of 125-ohm surge impedance operating into 1,000-ohm loads. Here it is seen that the time delays of the two networks are not quite the same and each reflection shows a greater deviation between the trailing edges. This is a method of making a more exact differential measurement without increasing the sweep speed.

The comparative test patterns shown by Fig. 6 illus-

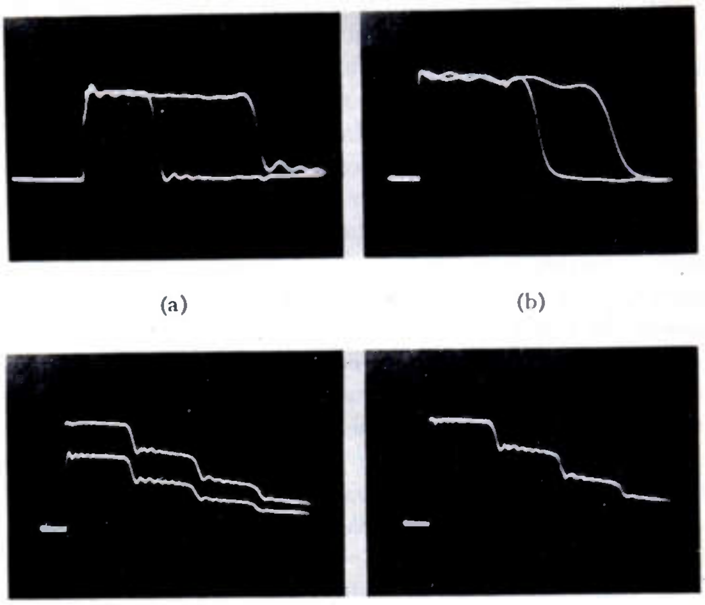

(a) (b)
(c) (d)

Fig. 5—Comparative test patterns: (a) 1- and 2-microsecond pulses; (b) 3- and 5-microsecond pulses; (c) two 2-microsecond pulses, mismatched load; (d) mismatched load, superposition.

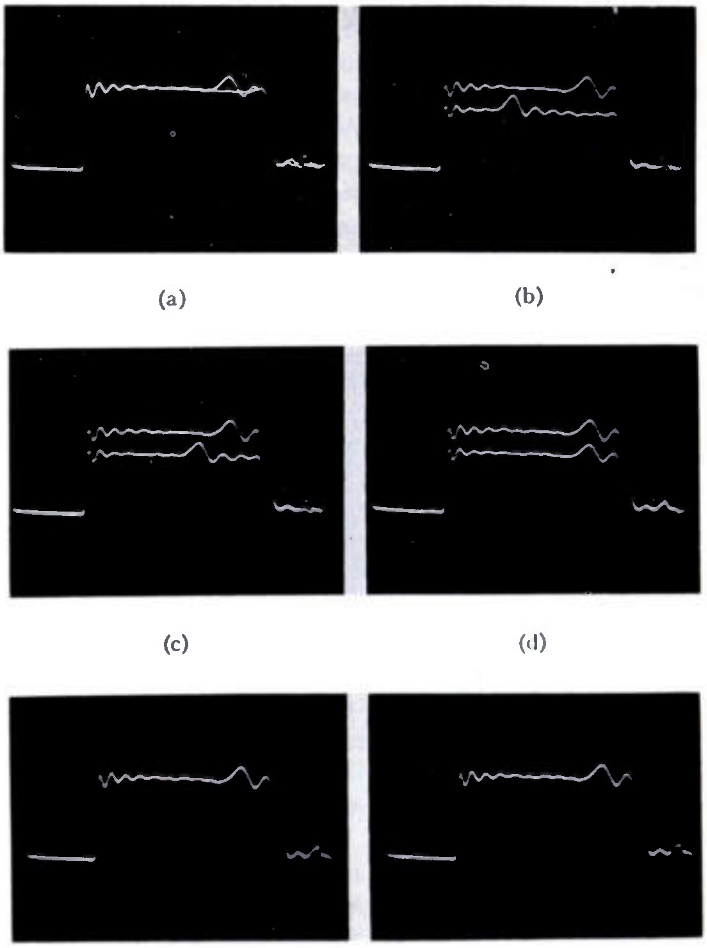

(a) (b)
(c) (d)
(e) (f)

Fig. 6—Troubleshooting a defective network: (a) defect in test network; (b) standard network coil 8–9 shorted; (c) standard network coil 12–13 shorted; (d) standard network coil 15–16 shorted; (e) same defect, superposition; (f) trouble clear, good network.

trate the procedure used for troubleshooting a defective network. Fig. 6(a) shows a "spike" occurring in the pattern of the test network. To facilitate study, the superimposed patterns are separated and an artificial defect inserted in the standard network by shorting a coil. The networks under test in this case are delay networks of twenty tandem sections, that is, 20 series coils and 21 shunt condensers. The series coils are designated starting 1–2 at the test-circuit end to 19–20 at the open-circuit end. The coil short is progressively moved along the delay line until, as shown by Fig. 6(d), the two irregularities occur in the same place. Superposing the patterns, Fig. 6(e), indicates that the defect is common to both networks. Thus the position and type of defect is determined, and it may then be marked and returned to the shop for repair. Fig. 6(f) completes the sequence by showing the repaired network returned and the trouble cleared.

In transient testing broad-band systems, a figure of merit is the rate-of-rise response of the system to a unit step pulse. The pulse response given by Fig. 7 is one way of illustrating the bandwidth of the impulse generator-electronic switch. The superimposed pulse voltages shown are produced by 3-foot lengths of RG65/U cable. The nominal propagation of this cable, is 0.042 microsecond per foot or 0.126-microsecond delay for 3 feet. Thus the open-circuited 3-foot cable tested as a two-terminal network produces a pulse of approximately 1/4-microsecond duration. Also, for a 3-foot length of the cable, the transmission loss is approximately 3 db at 100 Mc. From the transmission loss, the time of rise of the generated 1/4-microsecond pulse is then estimated approximately 0.005 microsecond. The measured rise time taken from the pulse pattern of Fig. 7 is 0.037 microsecond. Thus the rise time of the generated 1/4-microsecond pulse has been increased from 0.005 to 0.037 microsecond in transmission through the electronic switch. Hence a generated pulse with a rise time in the order of 0.035 microsecond is transmitted with negligible increase in wave-front rise time. This corresponds to a bandwidth of approximately 13.5 megacycles.

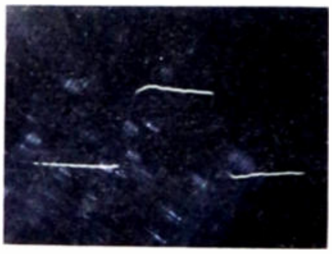


Fig. 7—One-fourth microsecond; 3 feet of RG65/U cable.

A bench-top setup of the impulse generator-electronic switch and associated components is shown by Fig. 8. It is estimated that over a period of three years of manufacturing testing, two of these test circuits saved 30,000 man hours.

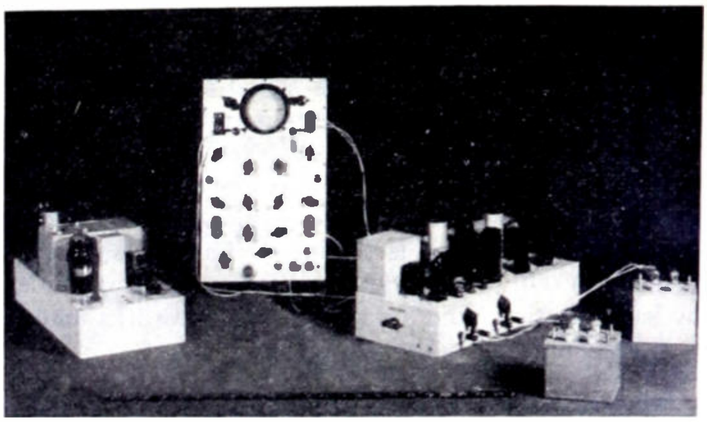


Fig. 8—Test circuit; table-top setup.

DESCRIPTION OF THE CIRCUIT

1. *Impulse Generator*

The operation of the impulse generator was described previously, and is shown by Figs. 1 and 2. The only additional comments that may be informative are that (1) due to the fact that the relay is not a polarized type, the contacts operate at a rate of 480 closures per second when driven from a 240-cycle source, and (2) the closure of the contacts is approximately 90 degrees out of phase with the applied 240-cycle voltage. This is important because the square-wave switching voltage is derived from the 240-cycle sine wave.

2. *Switching Circuit*

A. Clipper Amplifiers—V4, V5, V6, V7: The clipper amplifiers are conventional. The balanced-to-ground 240-cycle sine wave voltage that operates the relay is also amplified and clipped to produce two square-wave voltage waves 180 degrees out of phase with each other. The signal pulses, produced at contact closure, are absolutely phase locked with the square-wave switching voltages because both are derived from the same 240-cycle sine wave. A contact closure occurs approximately centered with each square-wave switching voltage.

B. Gated Amplifiers—V8, V9: The signal voltages developed by the impulse generator are applied to the control grids of the gated amplifier tubes. Signal switching is produced by applying the square-wave voltages to the respective screen grids. The amplitude of the square-wave voltages is approximately ±125 volts with reference to ground, and greater positive screen voltage is attained by a dc bias of 100 volts. Thus the screen voltages alternate between +225 and −25 volts with respect to ground. In order to regulate the screen potential when high screen currents are required, during the short interval that the control grids receive the microsecond positive signal pulses, 0.02 μf condensers are connected between screen grids and ground.

The gated amplifier tubes are normally cathode biased so that, with no signal voltage, only a small amount of late current flows when the screens are +225 volts. Vhen the signal pulses appear simultaneously on the espective control grids, the screen voltage of one tube vill be +225 and the other −25 volts. Thus one tube onducts and amplifies, and the other is blocked. This ondition alternates at the rate of 240 cycles per second nd the two amplified signals are interlaced in the comnon plate circuit.

C. Signal Delay: The start-stop sweep circuit is initiated at the same time that the signal pulses are generated, that is, when contact closure occurs. Because it requires approximately 0.2 microsecond to produce the linear sweep voltage, the important first 0.2 microsecond of the signal pulses would be lost. The $\frac{1}{2}$-microsecond delay network in the common plate circuit of the gated amplifier tubes stores the signal pulse until actuation of the sweep circuit is complete. The delay network transmits the signals to be observed with negligible distortion, but what visible deviations that do occur are common to both signals, i.e., reference-network pulse and test-network pulse.

3. *Sweep Circuit*

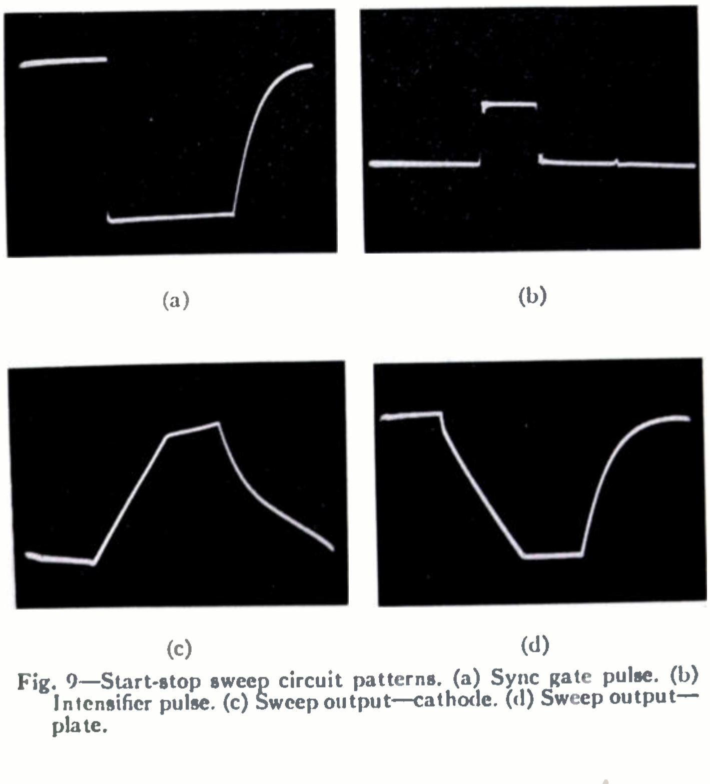
When the relay contacts close, a rising voltage is applied to the grid of the inverter (V_1). This tube is normally biased beyond cutoff by the negative dc charge voltage applied across the relay contacts. The closing of the contacts removes the bias voltage from V_1 and the plate current that flows drops the plate voltage approximately 275 volts. This negative gate synchronizes the sweep circuit and the intensifier so that the sweep and intensity voltages begin at the same time, each cycle, that the test signals are generated.

The intensity voltage is provided by a 6AC7 tube (V_2). Normally, current is flowing through this tube and the applied negative gate voltage drives the control grid beyond cutoff voltage and stops the plate current flow. The positive pulse generated in the plate circuit is applied, by means of the intensity gain control P_1 to the grid of the cathode-ray oscilloscope to control the beam brightness. For most conditions of operation, the intensity control on the oscilloscope is set so that there is normally no beam current and, consequently, no visible spot on the screen. The intensity voltage generated causes beam current to flow only during the sweep interval.

The negative gate voltage is also applied to the control grid of V_3, the sweep generating tube. This tube is normally conducting current, the current flowing through L_{2-1}, L_{2-2}, R_{16} and P_2. Variable resistor P_2 controls the amount of current by controlling the cathode bias voltage applied to the tube. The negative gate, 275 volts, applied to the control grid drives the tube beyond cutoff very suddenly and current ceases to flow through the tube. However, the energy stored in the coils L_{2-1} and L_{2-2} (approximately 15 henries each and magnetically tight coupled) tends to maintain current flow. The magnetic energy stored ($\frac{1}{2} LI^2$) discharges into the distributed capacity of the coil and the external circuits, and magnetic energy is transferred to electric energy in the condensers ($\frac{1}{2} CE^2$). This results in the plate and cathode of V_3 starting toward oscillation peaks of several thousand volts. The cathode makes a negative excursion tending toward an oscillation peak but when it reaches a voltage that sets the control grid-to-cathode voltage again within the conduction region of the tube, the tube conducts current and the oscillation is damped. During the interval that the cathode has traversed negatively, the plate has made an equal positive excursion. Due to the tight magnetic coupling between plate and cathode coils, the plate circuit oscillation is also damped after it has reached a positive excursion of 250 volts. The current flowing through the tube restores the magnetic fields of L_{2-1} and L_{2-2}; and after the negative gate is removed from the control grid, the sweep circuit restores to normal and is ready for another cycle.

As noted above, the cathode and plate voltage excursions are opposite in polarity; hence the sweep output voltage taken between plate and cathode is balanced to ground with an amplitude of 500 volts. The sweep speed is controlled by P_2 which adjusts the current flow and controls the magnetic energy stored in L_{2-1} and L_{2-2}. In that only the first 10 per cent or less of the sine wave oscillation is used, the sweep voltage developed is sufficiently linear. This balanced, linear voltage is applied to the horizontal plates of the oscilloscope and used as a time base.

The voltage wave forms of the sweep circuit are shown by Fig. 9. The duration of the inverter sync voltage is approximately 20 microseconds.


(a) (b)

(c) (d)

Fig. 9—Start-stop sweep circuit patterns. (a) Sync gate pulse. (b) Intensifier pulse. (c) Sweep output—cathode. (d) Sweep output—plate.

Intermediate-Frequency Gain Stabilization with Inverse Feedback*

G. FRANKLIN MONTGOMERY†, MEMBER, IRE

Summary—**Increased gain stability and gain-bandwidth product result from the use of inverse feedback in an intermediate-frequency amplifier. Improvement in gain stability is related to the number of cascaded stages, the stage gain, and the magnitude of the feedback. A circuit is described which uses feedback over a pair of cascaded stages. Generalized selectivity curves for this feedback couple are shown, and the design procedure is outlined. A description of an experimental amplifier concludes the paper.**

INTRODUCTION

INCREASED GAIN STABILITY and gain-bandwidth product result from the use of inverse feedback in an intermediate-frequency amplifier. In addition, the response curve of the amplifier may be designed to have a flatter top and steeper skirts than the curve for an amplifier of cascaded, synchronous, single-tuned stages without feedback. The improvement in flatness and gain-bandwidth product has been described by previous investigators for the case of feedback over a single stage.[1,2] This paper analyzes a method using feedback over pairs of stages and presents experimental confirmation of the design procedure.

Formulas introduced in the text are derived in detail in the appendix, where a complete list of symbol definitions appears.

STABILITY

Consider an amplifier of n identical cascaded stages, each stage having a voltage gain of A_1. The A_1 of each stage may be expected to vary with changes in tube transconductance due to power supply variations and tube aging, and with variations in constants of the interstage coupling networks. In the worst possible case of gain instability, a particular variation in A_1 will occur simultaneously in all stages. We shall examine the effect of such a simultaneous A_1 variation on the over-all gain A_n by forming the ratio of the percentage variation in A_n to the corresponding percentage variation in A_1. A convenient measure of the instability is the limit of this ratio as either of the variations approaches zero. This limit is called the instability factor

$$I = \lim_{\Delta A_1 \to 0} \left[\frac{\dfrac{\Delta A_n}{A_n}}{\dfrac{\Delta A_1}{A_1}} \right] = \left[\frac{A_1}{A_n} \right] \frac{dA_n}{dA_1}$$

and should be as small as possible for good gain stability.

If inverse feedback is applied over groups of m stages, and β is the voltage gain of the feedback network, then the over-all gain of the amplifier is[3]

$$A_n = \left[\frac{A_1^m}{1 + \beta A_1^m} \right]^{n/m}.$$

Differentiation of this expression leads to

$$I = n \left[\frac{A_n^{1/n}}{A_1} \right]^m \tag{1}$$

which is the general expression for I.

In the particular case where A_n is to equal a constant assigned value A_{ns}, we can find the number of stages n which results in the greatest gain stability. That is, if the over-all gain is held constant by increasing n and β simultaneously, the gain stability increases to a maximum (minimum I) and then decreases. For the maximum stability condition

$$n = m \ln A_{ns}$$

$$\beta = \frac{1}{e} - \frac{1}{A_1^m}$$

$$I = \frac{ne}{A_1^m}$$

where $e = 2.718$, and ln indicates the logarithm to the base e.

For a constant A_1, maximum stability is achieved at the cost of stage gain, since, for this condition

$$A_{ns} = e^{n/m}$$

and each m group has a gain of e. In practice, this degree of gain reduction is much larger than is necessary for good stability, the usual gain per stage being several times this value. If, for example, the number of stages n is $1/k$ times the number required for maximum stability, then I will be greater by the factor e^{k-1}/k. This factor is only 1.36 for $k = 2$ and 2.46 for $k = 3$, so that n may be considerably reduced without serious reduction of stability.

Notice from (1) that under any condition of feedback, if A_n and n are both held constant, it is more advantageous to increase feedback by increasing m rather than β, since $(A_n^{1/n}/A_1) < 1$. Because of difficulty in designing the β network, however, m will not usually be greater than three.

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* Decimal classification: R363.13. Original manuscript received by the Institute, March 23, 1949; revised manuscript received, January 30, 1950.

† Central Radio Propagation Laboratory, National Bureau of Standards, Washington, D. C.

[1] S. N. Van Voorhis, "Microwave Receivers," McGraw-Hill Book Co., New York, N. Y., pp. 88–91 and 175–178; 1948.

[2] E. H. B. Bartelink, J. Kahnke, and R. L. Watters, "A flat-response single-tuned if amplifier," PROC. I.R.E., vol. 36, p. 474; April, 1948.

[3] F. E. Terman, "Radio Engineering," McGraw-Hill Book Co., New York, N. Y., p. 248; 1937.

It should not be inferred that the use of feedback inevitably results in reduced gain per stage. Feedback increases the gain-bandwidth product by a factor dependent upon the magnitude of feedback and the β circuit characteristics. If the interstage coupling networks can be properly modified, feedback will produce in a given number of stages a gain higher than that of a single-tuned zero-feedback amplifier while maintaining the same bandwidth and improving the gain stability. The price which must be paid for this improved operation is the use of coupling networks of higher Q than would be required ordinarily. The amount of feedback which can be used in practice for a given bandwidth is thus limited by the obtainable circuit Q's, and this limitation is such that feedback reduces the gain per stage for bandwidths which are a small fraction of the center frequency. Relatively wide-band amplifiers do not suffer this limitation.

For a single-tuned stage with zero feedback, the gain-bandwidth product is

$$H_o = \frac{g_m}{2\pi C_1}.$$

For the feedback couple to be described, the gain-bandwidth product, per stage, is

$$H_B \approx 1.7 H_o \qquad (2)$$

in the more useful range of feedback values.

PARTICULAR METHOD

Fig. 1 is the schematic diagram of a cascaded pair of radio-frequency stages with inverse feedback. Power supply connections are omitted for simplicity. Symbols necessary to the design of this feedback couple are defined in the following list:

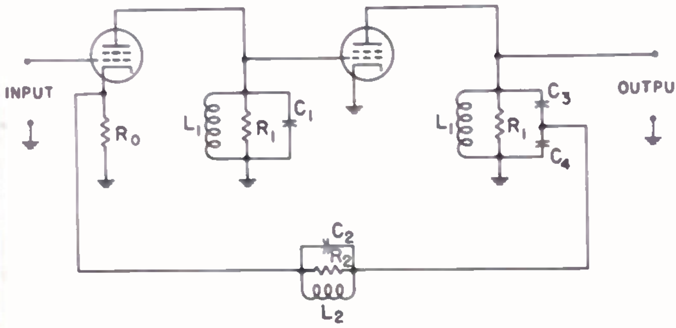

Fig. 1—Schematic diagram of feedback couple.

a = normalized voltage gain of feedback couple
A_0 = center-frequency voltage gain of couple with zero feedback
A_{c0} = center-frequency voltage gain of feedback couple
B = feedback factor
C_1 = plate-load capacitance in farads
C_2 = feedback capacitance in farads
C_3 = divider capacitance in farads
C_4 = divider capacitance in farads
f_0 = center frequency in cycles per second
Δf = bandwidth in cycles per second
g_m = tube transconductance in mhos
N = step-down ratio of tuned output circuit
$\omega_0 = 2\pi f_0$ in radians per second
P = design parameter
Q_1 = Q of tuned circuit consisting of shunt-connected L_1, C_1, R_1
Q_2 = Q of tuned circuit consisting of shunt-connected L_2, C_2, R_2
R_0 = cathode-bias resistance in ohms
R_1 = effective plate-load resistance in ohms
R_2 = effective feedback resistance in ohms.

Normalized gain curves for several values of feedback are given in Fig. 2, where a is plotted against $Q_1(\Delta f/f_0)$.

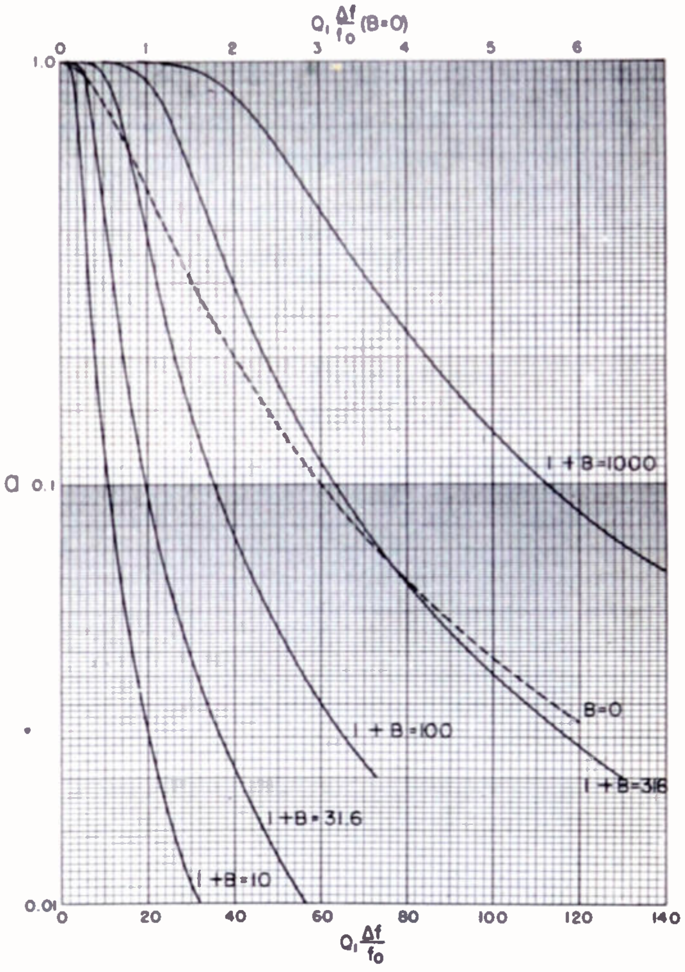

Fig. 2—Normalized gain for smooth response.

The curve labeled $B=0$, plotted to a different horizontal scale, is the normalized response of the couple with zero feedback. Note that the curves for appreciable feedback are more nearly rectangular than the curve for zero feedback.

The design procedure, using formulas developed in the Appendix and information given in the curves, is as follows:

Intermediate-Frequency Gain Stabilization with Inverse Feedback*

G. FRANKLIN MONTGOMERY†, MEMBER, IRE

Summary—**Increased gain stability and gain-bandwidth product result from the use of inverse feedback in an intermediate-frequency amplifier. Improvement in gain stability is related to the number of cascaded stages, the stage gain, and the magnitude of the feedback. A circuit is described which uses feedback over a pair of cascaded stages. Generalized selectivity curves for this feedback couple are shown, and the design procedure is outlined. A description of an experimental amplifier concludes the paper.**

INTRODUCTION

INCREASED GAIN STABILITY and gain-bandwidth product result from the use of inverse feedback in an intermediate-frequency amplifier. In addition, the response curve of the amplifier may be designed to have a flatter top and steeper skirts than the curve for an amplifier of cascaded, synchronous, single-tuned stages without feedback. The improvement in flatness and gain-bandwidth product has been described by previous investigators for the case of feedback over a single stage.[1,2] This paper analyzes a method using feedback over pairs of stages and presents experimental confirmation of the design procedure.

Formulas introduced in the text are derived in detail in the appendix, where a complete list of symbol definitions appears.

STABILITY

Consider an amplifier of n identical cascaded stages, each stage having a voltage gain of A_1. The A_1 of each stage may be expected to vary with changes in tube transconductance due to power supply variations and tube aging, and with variations in constants of the interstage coupling networks. In the worst possible case of gain instability, a particular variation in A_1 will occur simultaneously in all stages. We shall examine the effect of such a simultaneous A_1 variation on the over-all gain A_n by forming the ratio of the percentage variation in A_n to the corresponding percentage variation in A_1. A convenient measure of the instability is the limit of this ratio as either of the variations approaches zero. This limit is called the instability factor

$$I = \lim_{\Delta A_1 \to 0} \left[\frac{\dfrac{\Delta A_n}{A_n}}{\dfrac{\Delta A_1}{A_1}} \right] = \left[\frac{A_1}{A_n} \right] \frac{dA_n}{dA_1}$$

and should be as small as possible for good gain stability.

If inverse feedback is applied over groups of m stages, and β is the voltage gain of the feedback network, then the over-all gain of the amplifier is[3]

$$A_n = \left[\frac{A_1^m}{1 + \beta A_1^m} \right]^{n/m}.$$

Differentiation of this expression leads to

$$I = n \left[\frac{A_n^{1/n}}{A_1} \right]^m \tag{1}$$

which is the general expression for I.

In the particular case where A_n is to equal a constant assigned value A_{ns}, we can find the number of stages n which results in the greatest gain stability. That is, if the over-all gain is held constant by increasing n and β simultaneously, the gain stability increases to a maximum (minimum I) and then decreases. For the maximum stability condition

$$n = m \ln A_{ns}$$

$$\beta = \frac{1}{e} - \frac{1}{A_1^m}$$

$$I = \frac{ne}{A_1^m}$$

where $e = 2.718$, and ln indicates the logarithm to the base e.

For a constant A_1, maximum stability is achieved at the cost of stage gain, since, for this condition

$$A_{ns} = e^{n/m}$$

and each m group has a gain of e. In practice, this degree of gain reduction is much larger than is necessary for good stability, the usual gain per stage being several times this value. If, for example, the number of stages n is $1/k$ times the number required for maximum stability, then I will be greater by the factor e^{k-1}/k. This factor is only 1.36 for $k=2$ and 2.46 for $k=3$, so that n may be considerably reduced without serious reduction of stability.

Notice from (1) that under any condition of feedback, if A_n and n are both held constant, it is more advantageous to increase feedback by increasing m rather than β, since $(A_n^{1/n}/A_1) < 1$. Because of difficulty in designing the β network, however, m will not usually be greater than three.

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* Decimal classification: R363.13. Original manuscript received by the Institute, March 23, 1949; revised manuscript received, January 30, 1950.

† Central Radio Propagation Laboratory, National Bureau of Standards, Washington, D. C.

[1] S. N. Van Voorhis, "Microwave Receivers," McGraw-Hill Book Co., New York, N. Y., pp. 88–91 and 175–178; 1948.

[2] E. H. B. Bartelink, J. Kahnke, and R. L. Watters, "A flat-response single-tuned if amplifier," PROC. I.R.E., vol. 36, p. 474; April, 1948.

[3] F. E. Terman, "Radio Engineering," McGraw-Hill Book Co., New York, N. Y., p. 248; 1937.

It should not be inferred that the use of feedback inevitably results in reduced gain per stage. Feedback increases the gain-bandwidth product by a factor dependent upon the magnitude of feedback and the β circuit characteristics. If the interstage coupling networks can be properly modified, feedback will produce in a given number of stages a gain higher than that of a single-tuned zero-feedback amplifier while maintaining the same bandwidth and improving the gain stability. The price which must be paid for this improved operation is the use of coupling networks of higher Q than would be required ordinarily. The amount of feedback which can be used in practice for a given bandwidth is thus limited by the obtainable circuit Q's, and this limitation is such that feedback reduces the gain per stage for bandwidths which are a small fraction of the center frequency. Relatively wide-band amplifiers do not suffer this limitation.

For a single-tuned stage with zero feedback, the gain-bandwidth product is

$$\Pi_o = \frac{g_m}{2\pi C_1} \cdot$$

For the feedback couple to be described, the gain-bandwidth product, per stage, is

$$\Pi_B \approx 1.7\Pi_o \tag{2}$$

in the more useful range of feedback values.

PARTICULAR METHOD

Fig. 1 is the schematic diagram of a cascaded pair of radio-frequency stages with inverse feedback. Power supply connections are omitted for simplicity. Symbols necessary to the design of this feedback couple are defined in the following list:

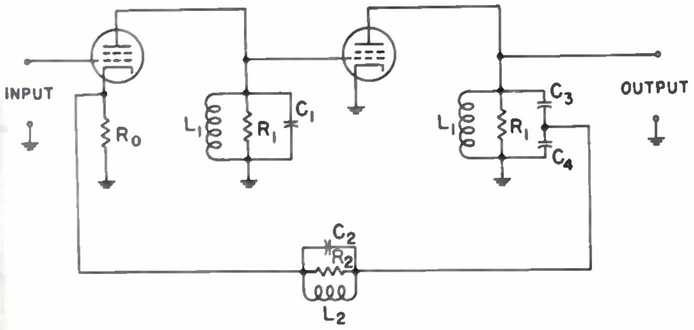

Fig. 1—Schematic diagram of feedback couple.

a = normalized voltage gain of feedback couple
A_0 = center-frequency voltage gain of couple with zero feedback
A_{c0} = center-frequency voltage gain of feedback couple
B = feedback factor
C_1 = plate-load capacitance in farads
C_2 = feedback capacitance in farads
C_3 = divider capacitance in farads
C_4 = divider capacitance in farads
f_0 = center frequency in cycles per second
Δf = bandwidth in cycles per second
g_m = tube transconductance in mhos
N = step-down ratio of tuned output circuit
$\omega_0 = 2\pi f_0$ in radians per second
P = design parameter
Q_1 = Q of tuned circuit consisting of shunt-connected L_1, C_1, R_1
Q_2 = Q of tuned circuit consisting of shunt-connected L_2, C_2, R_2
R_0 = cathode-bias resistance in ohms
R_1 = effective plate-load resistance in ohms
R_2 = effective feedback resistance in ohms.

Normalized gain curves for several values of feedback are given in Fig. 2, where a is plotted against $Q_1(\Delta f/f_0)$.

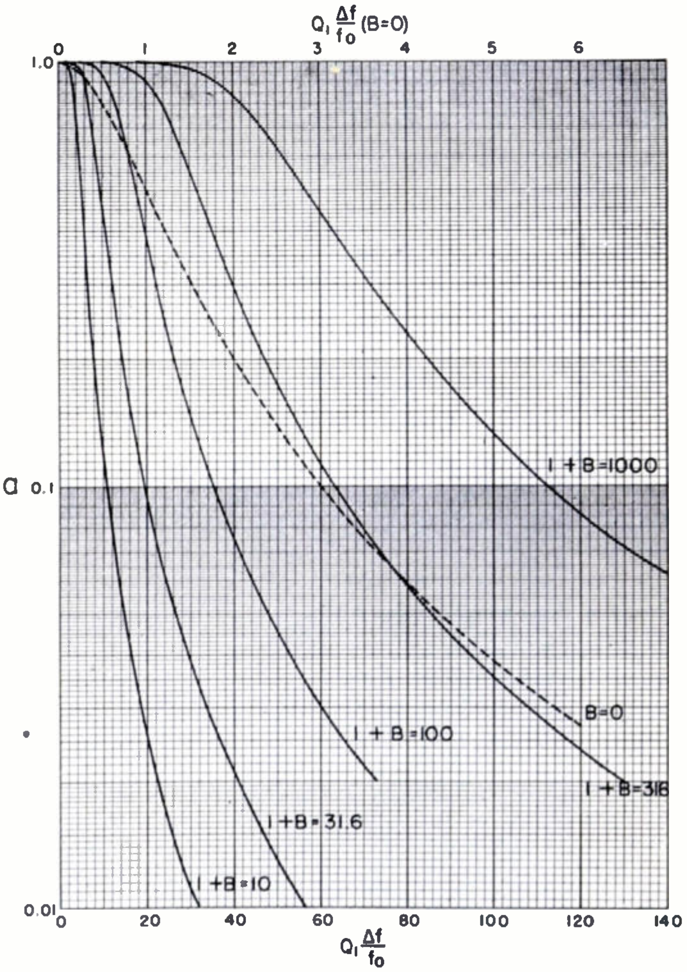

Fig. 2—Normalized gain for smooth response.

The curve labeled $B=0$, plotted to a different horizontal scale, is the normalized response of the couple with zero feedback. Note that the curves for appreciable feedback are more nearly rectangular than the curve for zero feedback.

The design procedure, using formulas developed in the Appendix and information given in the curves, is as follows:

Given Δf, f_0, g_m, and A_{c0}

A. Choose value of $1+B$. In Fig. 2 or 3, find $Q_1(\Delta f/f_0)$ for $a=1/\sqrt{2}$.

B. Calculate Q_1. If Q_1 is impractically large, choose a smaller value of $1+B$.

C. Calculate

$$C_1 = \frac{g_m Q_1}{\omega_0 \sqrt{A_{c0}(1+B)}}\cdot$$

D. Calculate

$$R_1 = \frac{Q_1}{\omega_0 C_1}\cdot$$

E. In Fig. 4, find P and A_0/N.

F. Calculate $A_0=(1+B)A_{c0}$, and N.

G. Choose R_0 from tube data.

H. Calculate

$$R_2 = \left[\frac{A_0}{NB} - 1\right] R_0.$$

I. Verify $N^2R_2 \gg R_1$. If this is not true, choose a smaller $1+B$ and redesign.

J. Calculate

$$Q_2 = \frac{A_0 P}{NB} Q_1.$$

K. Calculate

$$C_2 = \frac{Q_2}{\omega_0 R_2}\cdot$$

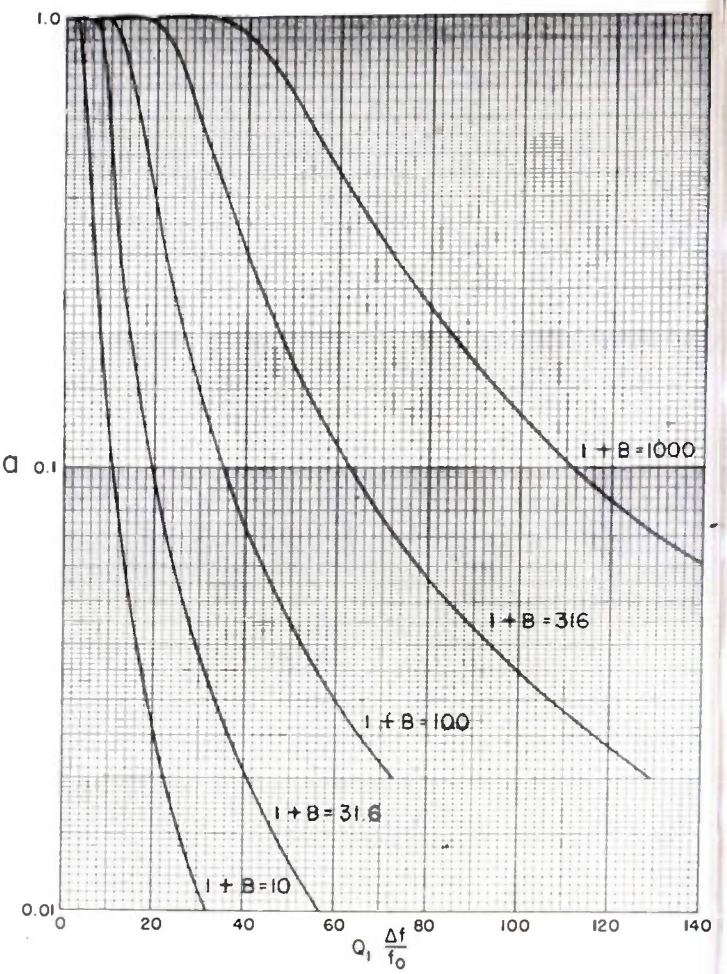

Fig. 3—Normalized gain for peaked response.

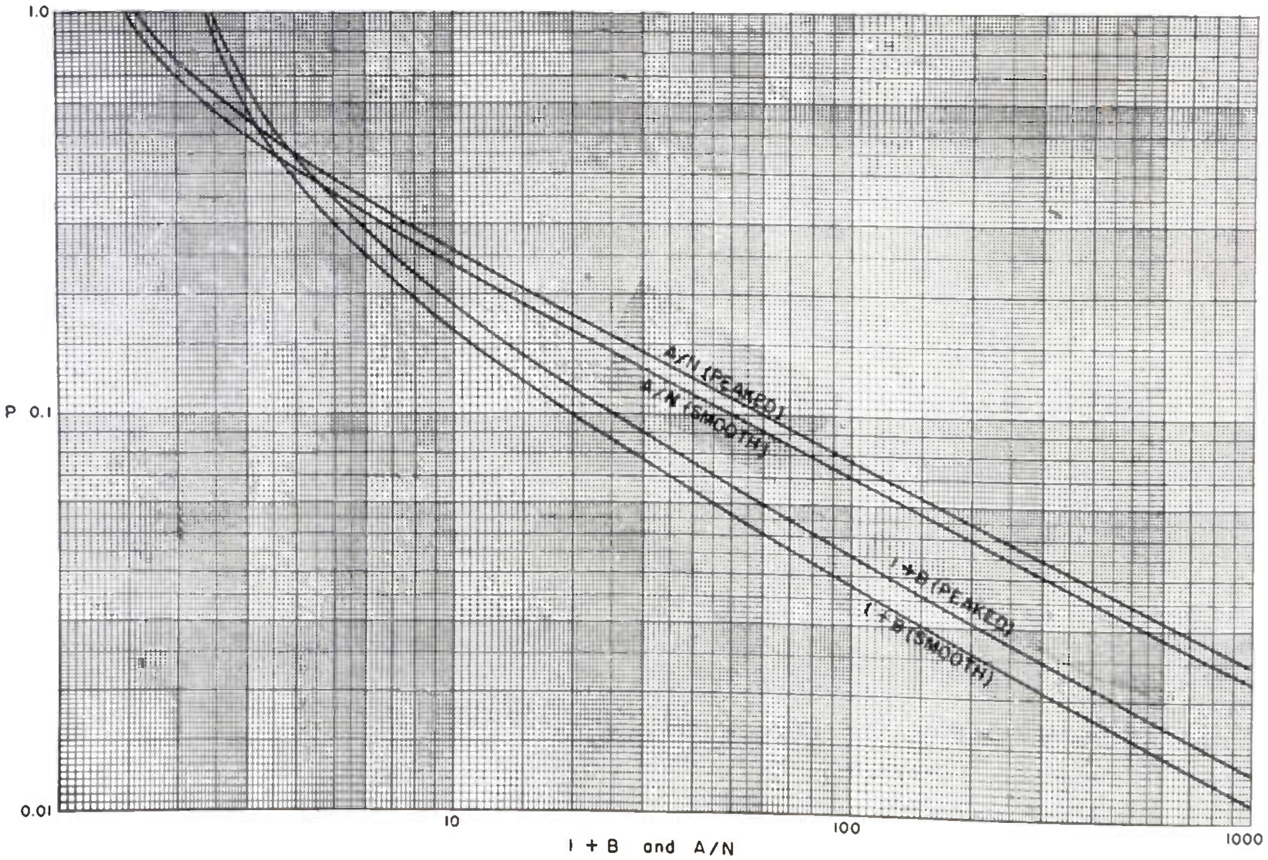

Fig. 4—Design factor chart.

L. Calculate

$$C_3 = \frac{NC_1}{N-1}, \qquad C_4 = NC_1.$$

Generally, the feedback factor B should be chosen as large as possible in order to obtain maximum stability. The improvement in center-frequency gain stability over a zero-feedback amplifier can be calculated directly from (1).

For example, if $n=4$, $m=2$, $A_n=10^4$, $A_1=10^2$, $I=4(0.1)^2=0.04$; a two-stage zero-feedback amplifier with equal gain has $I=2$. Consequently, an amplifier consisting of two feedback couples with the given constants has one-fiftieth the gain variation of a similar two-stage amplifier with no feedback.

In the appendix, it is shown that the design may be proportioned to give peaked responses at the extremes of the pass band. Curves for a slightly peaked response are given in Fig. 3.

EXPERIMENTAL RESULTS

The electrical arrangement represented by Fig. 1 must be duplicated in practice as closely as possible if results are to match the predictions of the design. Several practices, noted as critical during the course of the experimental work, should be followed:

1. Ensure adequate by-passing of the "ground" ends of the tuned plate circuits. This is especially important in the tapped output circuit. The design formulas are based on an output impedance at the tap which is usually a few ohms, and it does not take much reactance in the by-pass capacitor to modify the output impedance considerably.
2. By-pass the screen grid of the first stage directly to the cathode. The formulas for R_0 and C_1 will not be correct if the screen is by-passed to ground.
3. Install power lead decoupling as in an ordinary zero-feedback amplifier. Because of the greater gain-bandwidth product, the gain per stage will be even larger than for a zero-feedback amplifier. A small amount of regeneration may work mischief with a carefully calculated design.

Tuning the amplifier is greatly facilitated by providing a switch to break the feedback line to the cathode of the first stage. The amplifier plate circuits are peaked in the normal fashion. The feedback is then switched in, and the feedback tuned circuit is adjusted for maximum response at the center frequency. If a slight asymmetry of the response develops, it is usually possible to minimize it by detuning the feedback circuit. A large asymmetry indicates a design error, regeneration, or inadequate by-passing. A useful final adjustment is the value of R_0. A bandwidth which is too large can be decreased by decreasing R_0. If the bandwidth is too small, R_0 should be increased.

In Fig. 5 is shown the gain characteristic of a feedback couple using two 6SK7 tubes with the following design values: $\Delta f=30$ kc, $f_0=450$ kc, $g_m=2{,}000$ micromhos, and $A_{c0}=100$. The normalized gain curve for $1+B=10$, taken from Fig. 2, is superimposed for comparison.

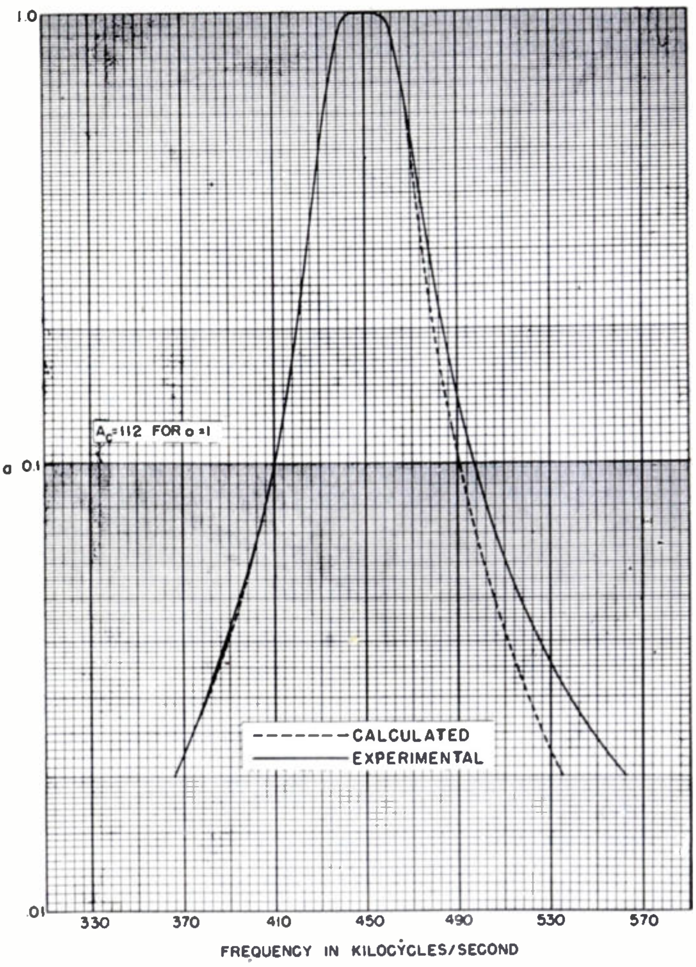

Fig. 5—Response of experimental amplifier.

The over-all gain and bandwidth agree well with the given values. The skirts are not quite as narrow as the normalized gain curve predicts. The lack of agreement is due to approximations that were made in the analysis. In particular, it was assumed that $N^2R_2 \gg R_1$, and that

$$\left|\frac{A}{N}\right| \gg |g_m Z_2|.$$

The latter inequality will not hold as well for frequencies far removed from resonance as for the center frequency. To the extent that the approximations are not achieved in practice, the skirts may be expected to deviate from the calculated values by small amounts.

The normalized gain curves plotted in Figs. 2 and 3 are also approximate in that the quantity u is considered equal to $\Delta f/f_0$, as explained in the appendix. This approximation was chosen because it allows the normalized response curves to be plotted as symmetrical characteristics, facilitating reading of bandwidth values. The approximation fails for bandwidths which are a large fraction of the center frequency, and in such cases it is advisable to plot curves with u equal to its exact value.

In Fig. 6 is shown the variation in gain of the experimental couple with plate supply voltage. A similar curve is shown for the same amplifier with zero feedback.

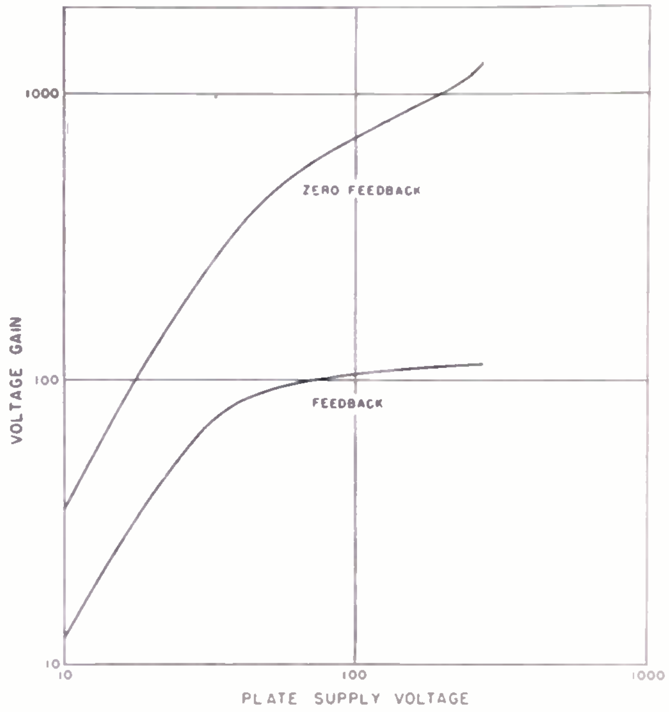

Fig. 6—Gain variation with supply voltage.

APPENDIX

Complete List of Symbols

a = normalized voltage gain of feedback couple
A = complex voltage gain of couple with zero feedback
A_0 = center-frequency voltage gain of couple with zero feedback
A_1 = voltage gain of single stage
A_c = complex voltage gain of feedback couple
A_{c0} = center-frequency voltage gain of feedback couple
A_n = voltage gain of cascade amplifier
A_{ns} = voltage gain of cascade amplifier with maximum stability
B = feedback factor
β = voltage gain of feedback network
C_1 = plate-load capacitance in farads
C_2 = feedback capacitance in farads
C_3 = divider capacitance in farads
C_4 = divider capacitance in farads
e = natural logarithmic base, 2.718
f = frequency in cycles per second
f_0 = center frequency in cycles per second
Δf = bandwidth in cycles per second
g_m = tube transconductance in mhos
I = instability factor
k = reduction factor
L_1 = plate load inductance in henries
L_2 = feedback inductance in henries
m = number of stages in each feedback loop
n = total number of stages
N = step-down ratio of tuned output circuit
ω = $2\pi f$ in radians per second
ω_0 = $2\pi f_0$
$\Delta\omega$ = $2\pi\Delta f$
P, P_1, P_2, P_3 = design parameters
Π_0 = single-tuned gain-bandwidth product
Π_B = feedback couple gain-bandwidth product
Q_1 = Q of tuned circuit consisting of shunt-connected L_1, C_1, R_1
Q_2 = Q of tuned circuit consisting of shunt-connected L_2, C_2, R_2
R_0 = cathode-bias resistance in ohms
R_1 = effective plate load resistance in ohms
R_2 = effective feedback resistance in ohms
σ = root of gain-bandwidth equation

$$u = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}$$

x = $Q_1 u$
x_c = value of x at $a = 1/2$
$\boldsymbol{Z}$ = parallel-tuned circuit impedance in ohms
$\boldsymbol{Z}_2$ = feedback-tuned circuit impedance in ohms.

Feedback Couple Analysis

The complex impedance of a two-pole formed by shunt-connected inductance L, capacitance C, and resistance R is

$$\boldsymbol{Z} = \frac{R}{1 + jQu}$$

where

$$Q = \frac{R}{\omega_0 L},$$

and

$$u = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \approx \frac{\Delta\omega}{\omega_0}.$$

In the circuit of Fig. 1

$$\boldsymbol{A}_c = \frac{\boldsymbol{A}}{1 + \left[\frac{\boldsymbol{A}}{N} + g_m \boldsymbol{Z}_2\right]\left[\frac{R_0}{R_0 + \boldsymbol{Z}_2}\right]}$$

where it is assumed that $N^2 R_2 \gg R_1$ and each plate resistance is much larger than its corresponding load impedance, so that

$$\boldsymbol{A} = \frac{A_0}{(1 + jQ_1 u)^2}.$$

If it is also assumed that

$$\left|\frac{\boldsymbol{A}}{N}\right| \gg |g_m \boldsymbol{Z}_2|,$$

then

$$A_c = \frac{A}{1 + \frac{A}{N}\left[\frac{R_0}{R_0 + Z_2}\right]}$$

which then becomes

$$A_c = \frac{A_0(1+jPx)}{1+B-(1+2P)x^2+j\left[2+\frac{A_0P}{N}+P(1-x^2)\right]x}$$

where

$$B = \frac{A_0}{N\left[1 + \frac{R_2}{R_0}\right]}$$

$$P = \frac{NBQ_2}{A_0Q_1}$$

and $x = Q_1u$.

If we let

$$P_1 = \left[\frac{2 + (1 + A_0/N)P}{1 + B}\right]^2 - 2\left[\frac{1 + 2P}{1 + B}\right]$$

$$P_2 = \frac{1 + 2(1 - A_0/N)P^2}{(1 + B)^2}$$

and

$$P_3 = \frac{P^2}{(1 + B)^2},$$

then the normalized gain

$$a = |A_c|\left[\frac{1 + B}{A_0}\right]$$

becomes

$$a = \left[\frac{1 + P^2x^2}{1 + P_1x^2 + P_2x^4 + P_3x^6}\right]^{1/2}.$$

The relative values of the coefficients in this equation determine the shape of the normalized response.

It can be seen by inspection that, for maximum flatness with no inflection,

$$P_2 = 0, \qquad P_1 = P^2.$$

These two conditions give

$$\frac{A_0}{N} = 1 + \frac{1}{2P^2}$$

and

$$1 + B = \frac{1 + 2P}{P^2}\left[\sqrt{1 + (\tfrac{1}{2} + P)^2} - 1\right]. \tag{3}$$

Plots of these two functions are given in Fig. 4. The equation for a now becomes

$$a = \left[\frac{1}{1 + \frac{P^2x^6}{(1 + B)^2(1 + P^2x^2)}}\right]^{1/2}. \tag{4}$$

Plots of this function are given in Fig. 2.

It should be noted that responses with almost any degree of peaking are available with different choices of the coefficients P_1, P_2, and P_3. For any particular set of these coefficients, it is necessary to derive new expressions for A_0/N, $1+B$, and a in terms of P. Responses may be obtained which give three peaks to the usual selectivity curve, or two inflections in the equation for a. A slightly peaked response with one inflection is obtained by setting

$$P_2 = -1/[5(1 + B)^2], \qquad P_1 = P^2.$$

The corresponding functions are plotted in Figs. 3 and 4.

Gain-Bandwidth Product

It can be shown that

$$\Pi_B = \Pi_0\frac{x_c}{\sqrt{1 + B}}$$

where Π_0 is the gain-bandwidth product for a single-tuned stage, Π_B is the product for the feedback couple, and x_c is the value of x at $a = 1/2$. The ratio $x_c/\sqrt{1+B}$ is approximately 1.7 over the useful range of $1+B$ in Fig. 2, and therefore

$$\Pi_B \approx 1.7\Pi_0. \tag{2}$$

It is interesting to examine the limiting value of Π_B as B and A_0 become infinite. Substitution of $a = 1/2$, $x = x_c$ in (4) yields

$$\sigma^3 = 3\left[\sigma + \frac{1}{P^2(1 + B)}\right]$$

where

$$\sigma = \frac{x_c^2}{1 + B}.$$

From (3), we note that

$$\lim_{B\to\infty} P^2(1 + B) = \sqrt{5/4} - 1.$$

An approximate solution for the real root gives

$$\lim_{B\to\infty} \sigma \approx 3.279$$

and therefore

$$\lim_{B\to\infty} \Pi_B \approx 1.811\Pi_0.$$

The Application of Direct-Current Resonant-Line-Type Pulsers to the Measurement of Vacuum-Tube Static Characteristics*

J. LEFERSON†, ASSOCIATE, IRE

Summary—In order to utilize a power tube efficiently, the static characteristics of this tube must be known. From these characteristics, optimum load lines may be determined and circuit components of the equipment may be defined. The problem of obtaining these characteristics is usually not a simple one, due to the power-input limitations put upon a tube. For this reason, characteristic curves were previously plotted by capacitor-discharge methods or were calculated mathematically, using low power points as references. This paper describes a method of obtaining this information, employing pulse circuits commonly used in radar transmitters.

MEASUREMENT OF STATIC CHARACTERISTICS FOR POWER TUBES

THE PROBLEM of measuring static characteristics of power tubes is aggravated by the fact that these tubes always require positive grid driving voltage which results in appreciable power to be dissipated by the electrodes. These curves, therefore, cannot be obtained by ordinary direct-current means as is common in receiver-type tubes where the grids are mostly operated at a negative potential.

In a practical case, the ML-354, 50-kw FM tube, grid and plate dissipation at the end of a selected load line may be 65 kw, 14 kw on the grid, and 51 kw on the anode. This is not the maximum dissipation that would be encountered by using direct-current means of measurement, as the regions beyond and above the load-line figures quoted above must be explored. Power dissipation loads up to 5 times these quantities may be encountered.

Previously, capacitor-discharge methods were used or curves were obtained by calculation from low level direct-current points.[1,2]

DESCRIPTION OF CIRCUIT

Fig. 1 shows a simplified circuit of the test setup. The circuit to the left of line *A-A* is a conventional-line-type pulser using resonant charging. The principal components (Fig. 2) consist of a direct-current power supply, charging choke, pulse-forming network, load resistor and a 5C22 hydrogen thyratron.[3]

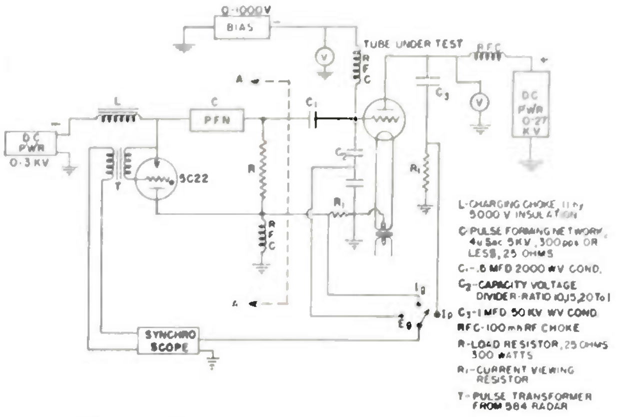

Fig. 1—Simplified schematic diagram of pulser.

Most power tubes have fairly low grid input impedances. The ML-354 with 100 ohms has about the highest. This input impedance should match the network impedance to avoid undesirable effects.

Since the pulser is to be used on many types of tubes, and therefore varying impedances, stability can best be obtained by using a resistor in parallel with the tube load. The tube load is, therefore, a small part of the load impedance for the line pulser resulting in more stable operation. The addition of the parallel resistor results in inefficient operation, but in this case a slightly higher expenditure of power from the direct-current power supply is tolerated to obtain added flexibility.

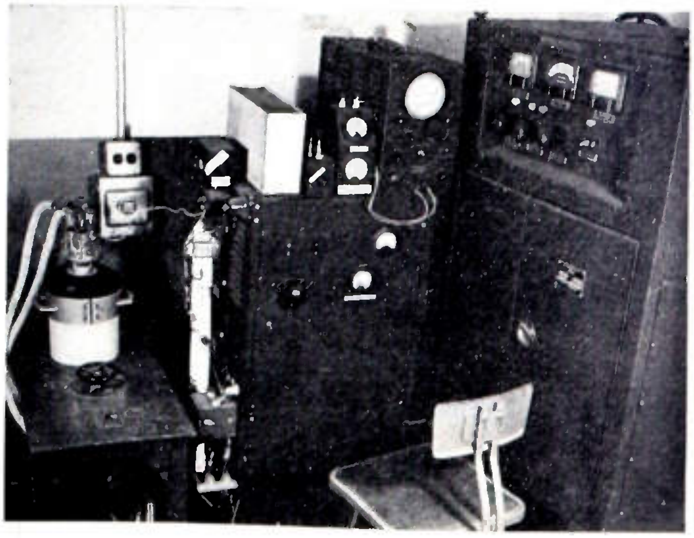

Fig. 2—Setup of equipment. The resistors on the side of the rack comprise the load resistor *R*. The rectangular can on the top left is the pulse-forming network, and the two cased meters read plate voltage applied to the tube under test and deflection voltage on the synchroscope.

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* Decimal classification: R252. Original manuscript received by the Institute, May 12, 1948; revised manuscript received, December 1, 1949.

† Machlett Laboratories, Inc., Springdale, Conn.

[1] E. L. Chaffee, "The characteristic curves of the triode," PROC. I.R.E., vol. 30, pp. 383–395; August, 1942.

[2] J. Bell, J. W. Davies, and B. S. Gossling, "High power valves: construction, testing and operation," *Jour. I.E.E.* (London), vol. 83, pp. 188–193; August, 1938.

[3] "Pulse Generators," edited by G. N. Glascoe and J. V. Lebacqz, Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., pp. 341–354, p. 173; 1948.

The front of a pulse from line-type pulsers is subject to oscillation and other distortions due to the large rate of rise of pulse current, so a fairly long pulse length is desirable. A range of from two to five microseconds has been found satisfactory. The pulse-repetition frequency is kept as low as feasible since average current from the direct-current power supply rises with the repetition frequency and a high repetition rate is not necessary. The tube under test is coupled directly to the output of the pulser to avoid using a pulse transformer. Pulse transformers require fairly close matching to ensure a satisfactory pulse and this is not possible if many tube types are to be measured. There have been occasions when a higher voltage was desirable, and a pulse transformer with a ratio of three to one has been inserted across the load resistor with satisfactory results. For the majority of the measurements, the pulse transformer is not used. Direct coupling necessitates reversing both the power supply and hydrogen thyratron, thereby running the cathode of the 5C22 at high voltage; but a pulse transformer similar to that used in the 584 radar for coupling to the output stage proves quite satisfactory.

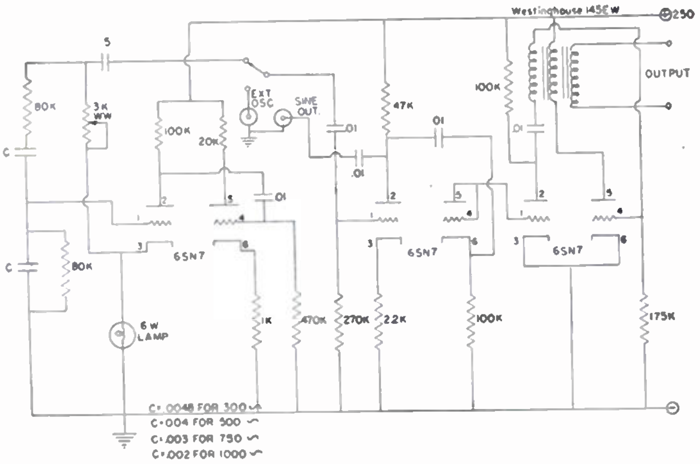

Fig. 3

The 5C22 may be driven by a synchroscope or any other frequency source giving approximately 150 volts. A synchroscope is an oscilloscope constructed so that the horizontal sweep is triggered by either the trigger source actuating the pulser (synchronous operation) or by the signal to be observed (self-synchronous operation). These are manufactured by several manufacturers, among them Sylvania Electric Products Inc., James Millen Manufacturing Company, Inc., and Browning Laboratories, Inc. Fig. 3 is the schematic of the trigger source being used at present to trigger a Browning P4 synchroscope which in turn is driving the 5C22 thyratron.

A description of the operation of the pulser follows. The tube under test is biased to a point where no plate current will flow at the maximum plate voltage to be used. This bias voltage is applied with a direct-current power supply and remains constant on the grid. The pulser impresses positive voltage pulses between the grid and cathode at the rate of 300 per second with a duration of 4×10^{-6} seconds. The amplitude of the pulse is chosen so as to drive the grid voltage to the point to be measured. The plate conducts for the same length of time, 4 microseconds. Plate voltage is applied to the tube by means of a direct-current power supply charging a storage capacitor C_3. This capacitator must be of such size that the voltage remains fairly constant during the time the tube conducts. Proper radio-frequency chokes are incorporated to isolate the pulse voltage from the various supplies.

The grid and plate current are read by means of voltages developed across current-viewing resistors. These resistors must be noninductive in nature since inductance will cause high voltage spikes in accordance with $E = L(di/dt)$. Sprague Koolohm noninductive resistors are quite satisfactory in this respect, and Fig. 4 shows a method of mounting these resistors to afford plug-in convenience.

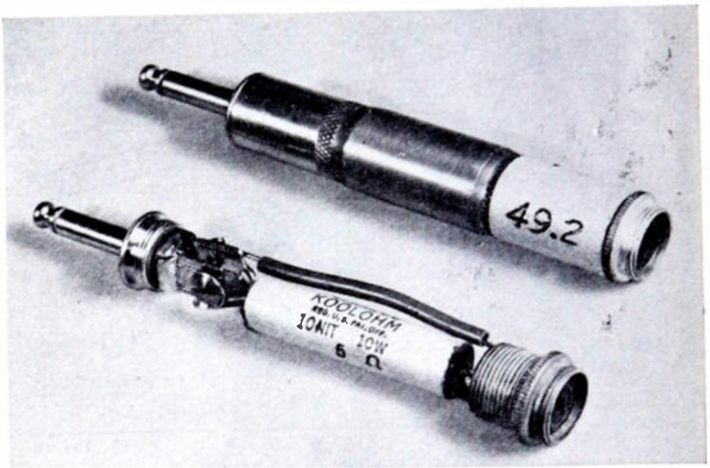

Fig. 4—Construction of current-viewing resistors. Sprague Koolohm noninductive resistors are satisfactory. The completed unit is calibrated on a bridge after assembly.

Bias voltage and plate voltage are read by standard direct-current voltmeters. Grid voltage is read by means of a capacity divider arranged so as to sample the pulse voltage going into the grid. Care must be taken in designing this divider so that reasonable deflection is obtained for any voltage range. This can be done by providing three ratios so that the appropriate one may be selected when needed.[4]

Using a suitable current-viewing resistor, capacity divider, and direct-current meters, an accuracy of ±5 per cent may easily be obtained. By using a little care in reading the synchroscope, an accuracy of ±2 per cent is possible. As a check, low current points are measured on a direct-current test set.

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[4] See pp. 673–677 of footnote reference 3.

Discussion of Results

The above pulser is used for many tests. By tying together the plate and grid of a tube and applying the pulse to these electrodes, peak cathode-emission data may be obtained. This is a very important characteristic since emission varies with cathode temperature and tungsten filaments must run ±15° C for uniform life. Routine measurements on production tubes are a check on standard quality.

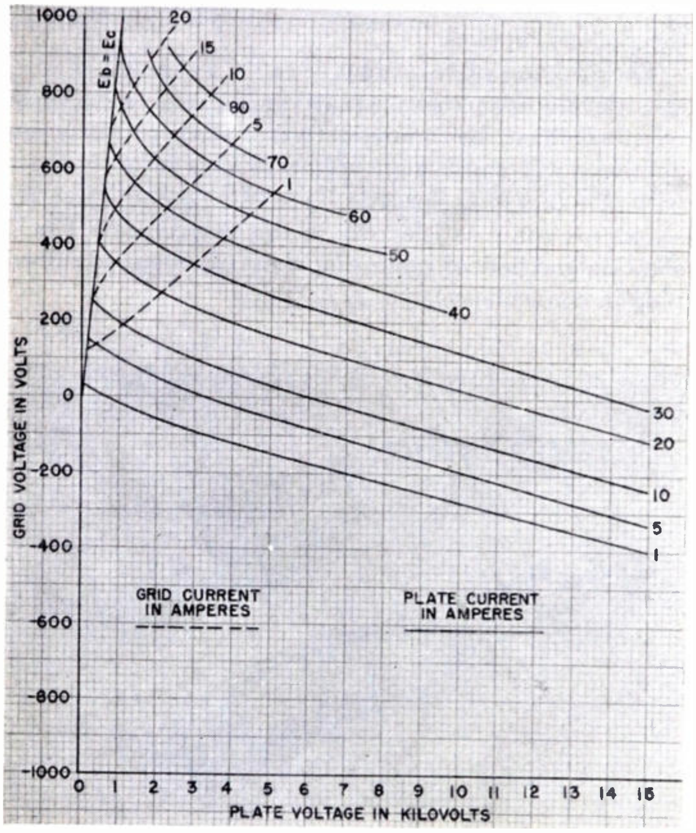

Fig. 5—ML-354 constant current characteristics.

In the experimental tube, cathode emission is very important, as is exploring the emission to ascertain the knee of the plate voltage-cathode current curve. These curves are easily obtained by the pulser method.

Static characteristics are obtained for every condition of operation with no injury to the tube. Fig. 5 is the static characteristic of the 50-kw, 100-Mc FM tube obtained with the above pulser. Since the duty cycle, 300 cycles per second $\times 4 \times 10^{-6}$ seconds equals 0.12 per cent, instead of 65-kw input to the tube, only 78 watts need be dissipated for the above-mentioned load-line point.

Primary grid emission, which varies with grid temperature and material, is of concern in connection with tube characteristics. Since the pulser described does not heat the grid of the tube under test, the effects of increased temperature will, therefore, not be apparent in the curves.

In the standard test for power tubes employed by Machlett Laboratories, alternating current is applied to the grid, the positive cycles used to heat the grid, and the negative cycles to measure the primary emission. A careful check on data obtained this way reveals that primary grid emission in properly designed and processed power tubes is low enough, up to maximum rated input, to be ignored when measuring static characteristics.

Effects of secondary grid emission, however, can be measured. Secondary grid emission is more a function of grid material and shielding at the ends of the grid itself. Tests have been made on the effects of enclosing the "cold" ends of the grid of an ML-846 tube with solid bands in place of the usual wire. Fig. 6 shows the curves of a regular production ML-846 compared with the test tube and the drastic change in grid current can be noted.

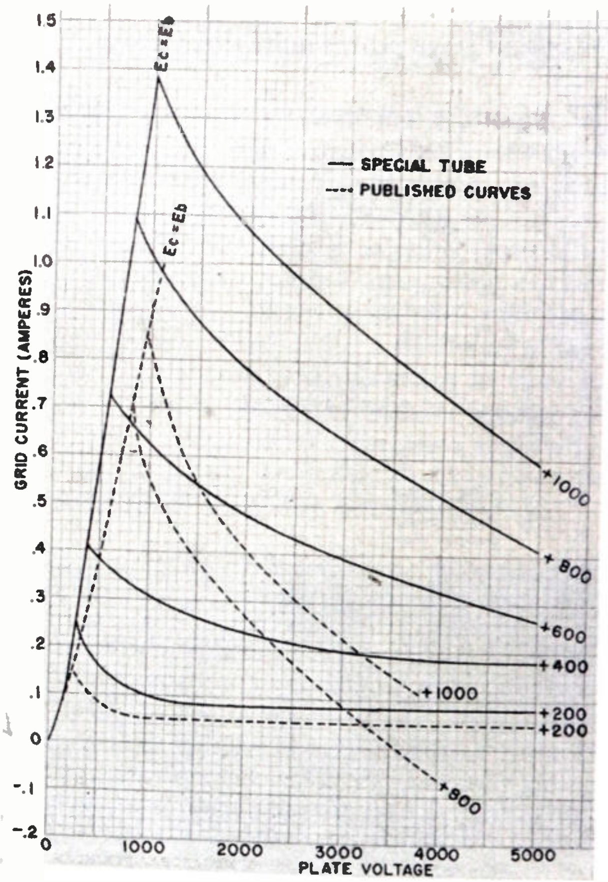

Fig. 6—ML-846 with special grid.

Acknowledgment

The author wishes to thank H. D. Doolittle for his invaluable suggestions, both in building the equipment and in preparation of this paper.

Wide-Range Tunable Waveguide Resonators*

W. W. HARMAN†, ASSOCIATE, IRE

***Summary*—A study is made of the design of broad-band (2:1 tuning range) resonators for use with reflex klystrons of the external cavity type. The resonators considered are variations of a general type consisting of a section of waveguide with movable shorting plungers at each end and the tube located at the center. They are divided into two groups: quarter-wave or fundamental mode resonators whose high-frequency limit is set by the physical size of the tube envelope, and three-quarter-wave resonators which allow operation up to the electronic limit of the tube. Methods of treating the mode interference problem are considered for both classes of resonators.**

I. Introduction

THE INVESTIGATION described in this paper is a small portion of a large program carried on in various laboratories, largely under Government sponsorship, toward the development of reflex klystron oscillators continuously tunable over a frequency range of the order of two to one.[1,2] The majority of this work has been with concentric line resonators. Although waveguide resonators have been built, there has not appeared a comprehensive study and evaluation of them. They assume importance because, for operation at the short wavelength end of the microwave spectrum where a high resonant impedance becomes exceedingly important, they are in general superior to concentric-line resonators.[3]

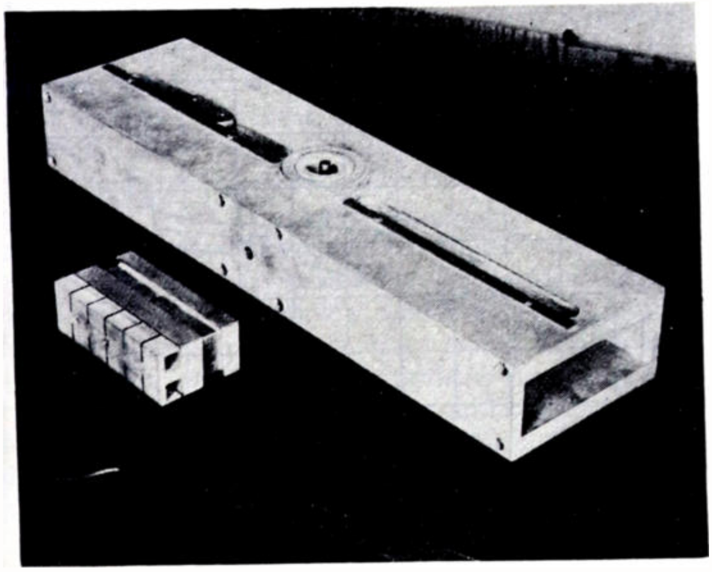

Fig. 1—Typical waveguide resonator with noncontacting shorting plungers. The narrow slots visible on the top of the shorting plunger are filled with aquadag to suppress plunger resonances.

Fig. 1 shows a typical waveguide resonator, consisting of a shorted section of rectangular waveguide with the reflex klystron at the center and two shorting plungers moved symmetrically on either side of the tube. The plungers shown are of the noncontacting choke type. The slots visible on top and bottom are filled with aquadag and have as their purpose the suppression of plunger resonances.

Oscillation Modes

As a preliminary it is necessary to examine the nature of the oscillation modes of a waveguide resonator. For this purpose there is available an excellent tool—the Kron equivalent network.[4] Data for Figs. 2, 3, and 4 were obtained by simulating the resonator on such a network board.

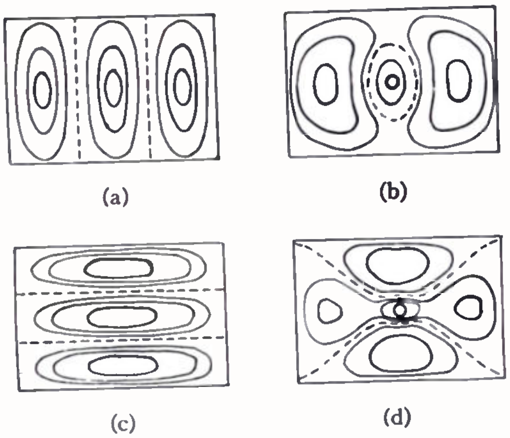

Fig. 2—*H* field configuration for various waveguide resonator modes: (a) [1, 3] mode in shorted waveguide section; (b) [1, 3] mode in rectangular waveguide resonator; (c) [3, 1] mode in waveguide section; (d) [3, 1] mode in rectangular waveguide resonator.

In order for a resonant mode of the shorted waveguide to present a sufficiently high impedance to the tube for oscillations to occur, the tube must be located in a region of high electric field. Since only field configurations having electric fields essentially parallel to the tube axis need be considered, this is equivalent to saying that the center of the resonator must be at a center of a loop of magnetic field. The lowest frequency mode which meets this condition, having a single loop of magnetic field around the tube, is commonly called the "quarter-wave" mode; if the resonator is considered as a section of waveguide operating in the TE_{10} mode, there is something less than a quarter guide wavelength between shorting plunger and tube.

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* Decimal classification: R119.3. Original manuscript received by the Institute, March 1, 1949; revised manuscript received, December 27, 1949. Presented 1948 National Electronics Conference, Chicago, Ill., November 4–6, 1948. This work was performed at Stanford University in connection with joint Office of Naval Research and Signal Corps contract, and under the generous provisions of an RCA fellowship in electronics.

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[1] W. H. Huggins, A.M.C. Special Report No. 185C/4, Army Air Forces Communications Laboratory, Cambridge Field Station, Cambridge, Mass., 1946.

[2] Radio Research Laboratory Staff, "Very High Frequency Techniques," McGraw-Hill Book Co., Inc., New York, N. Y., chap. 31 and 32; 1947.

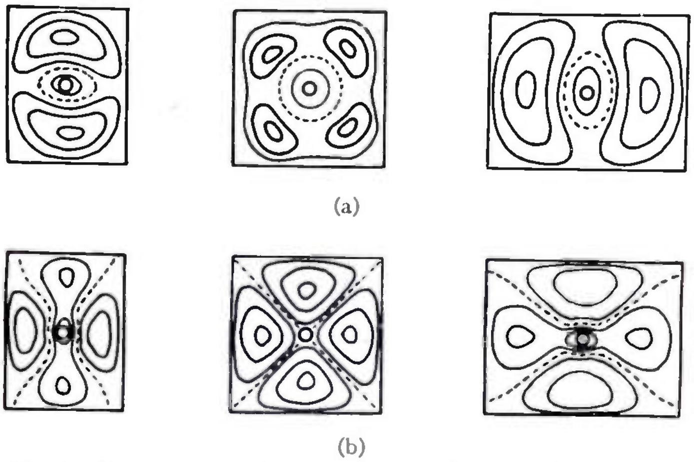
[3] A. V. Haeff, T. E. Hanley, and C. B. Smith, "Wide-range ultra-high-frequency signal generators," Proc. I.R.E., vol. 35, p. 1137; October, 1947.

[4] G. Kron, "Electric circuit models of partial differential equations," *Elec. Eng.*, vol. 67, p. 672; July, 1948.

The next higher frequency resonant mode which can support oscillations has three loops of magnetic field end to end, as shown in Fig. 2(b); the next higher frequency oscillating mode of Fig. 2(d) has three loops of magnetic field side by side. In these field plots the nodal surfaces of electric field are indicated by dotted lines. The effect of the capacitance of the tube grids on the field configurations of these modes, as seen by comparing Figs. 2(a) and 2(c) with Figs. 2(b) and 2(d), is to concentrate the electric field between the grids and to shrink the magnetic field patterns in around the tube.

For convenience, these modes will be symbolized by brackets enclosing two numbers, the first being the number of loops of H field side by side in the waveguide, and the second the number of loops end to end between shorting plungers. The [1, 3] mode of Fig. 2(b) we shall refer to as the "three-quarter-wave" mode, the distance between shorting plunger and tube being somewhat less than three quarters of a guide wavelength for a waveguide operating in the TE_{10} mode.

It is interesting and significant to observe the changes in the field configurations as the shorting plungers are moved. Fig. 3 shows that, for both the modes of Fig. 2, as the resonator is tuned through the square position the field pattern suddenly rotates 90 degrees. In the case of the three-quarter-wave mode the tube is always in a region of high electric field so that this mode presents to the tube a high resonant impedance or shunt resistance. For the mode of Fig. 3(b), it is seen from the position of the nodal surfaces (dotted lines) that the electric field at the tube is very low when the resonator is square. The result of this is that the tube drops out of oscillation when the plungers near this position.



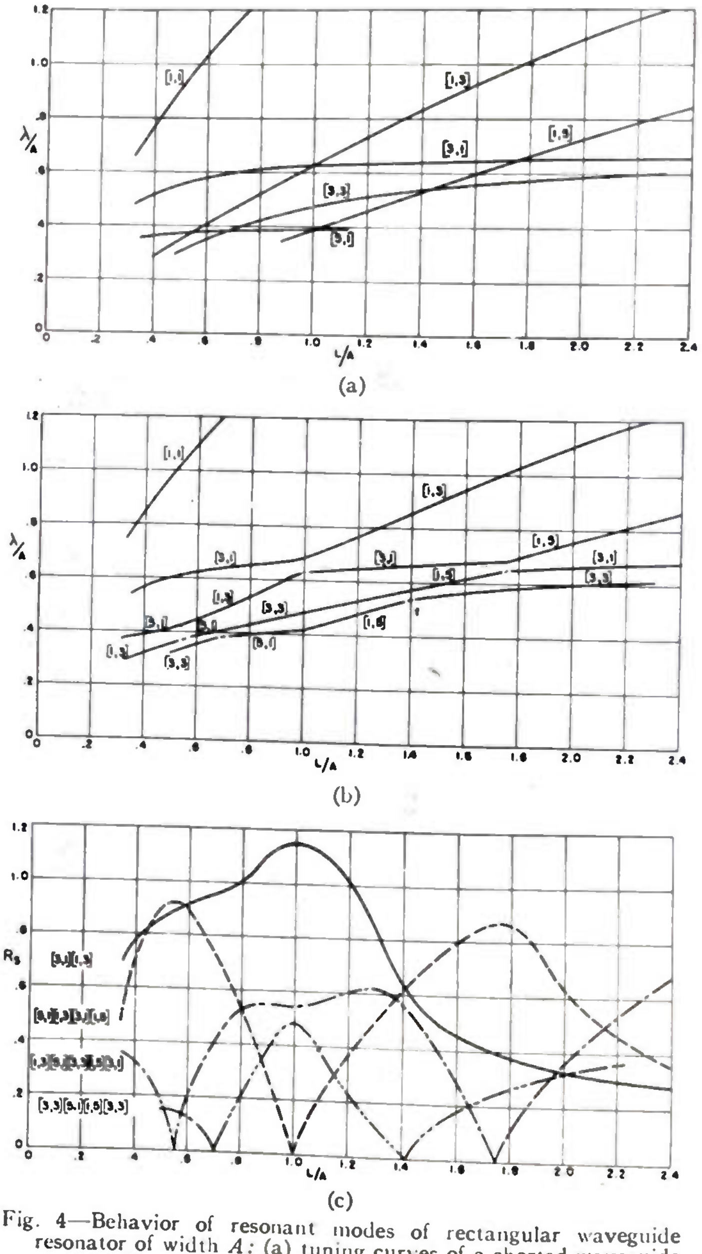
(a)

(b)

Fig. 3—H field configuration of (a) [1, 3] mode and (b) [3, 1] mode in a rectangular waveguide resonator for various plunger positions.

Some light may be shed on this phenomenon by examining the effect of the tube-grid capacitance on the tuning curves of the resonator. Fig. 4(a) shows the tuning curves of wavelength versus resonator length of a shorted section of waveguide for those resonant modes which have a high concentration of electric field at the resonator center. The upper curve is for the quarter-wave mode, the next for the three-quarter-wave mode, and so on. It is to be noted that at certain points marked by the crossings of these curves, the modes are degenerate; that is, two different field configurations have the same resonant frequency. As in the analogous case of coupled tuned circuits, the effect of any coupling between the modes is to split the degeneracy, yielding two different resonant frequencies. Perturbations of the resonator boundaries, such as the re-entrant structure of the tube or asymmetries in the resonator construction, may provide such coupling.

If the two crossing modes are both oscillating modes, the coupling provided by the tube is strong and, at the resonator length corresponding to the degeneracy, two widely separated resonant frequencies appear. All degeneracies vanish and the tuning curves of oscillating modes do not cross. As seen in Fig. 4(b), in each case of degeneracy the low-frequency branches of the tuning



(a)

(b)

(c)

Fig. 4—Behavior of resonant modes of rectangular waveguide resonator of width A: (a) tuning curves of a shorted waveguide section; (b) effect of center capacity on tuning curves of shorted waveguide section; (c) relative shunt resistance of modes of shorted waveguide section with center capacity.

curves join, representing a mode with consistently high shunt resistance; the high-frequency branches join to give a mode which, like the mode of Fig. 3(b), has a low resonant impedance near the degeneracy. The relative shunt resistance of the various modes is shown in Fig. 4(c), the highest curve representing the three-quarter-wave mode. (The quarter-wave mode is not shown, but its resonant impedance is considerably higher, by a factor of three or four, than those shown.) The futility of attempting to represent any one of these modes over its entire tuning range by the suggested combination of two numbers is apparent. Different sections of the same tuning curve are derived from different modes of the unperturbed resonator, and may correspond to completely different field configurations.

Because of the regions of low shunt resistance and nonoscillation for all modes but the quarter-wave and three-quarter-wave, these are the only useful modes for wide-range tuning.[5] They will be considered separately and the resonators considered grouped into quarter-wave and three-quarter-wave types.

Mode Interferences

In the perturbed resonator, the resonant wavelengths of the various modes bear no simple relationship to one another and it is inevitable that the various klystron modes, or regions of oscillation on a plot with co-ordinates of reflector voltage and resonator tuning, will interfere. A potential mode interference exists whenever conditions of reflector voltage and resonator tuning are such that the tube could oscillate in either of two modes. Under such circumstances, the factors of load and tube conductance and previous history that determine which oscillation frequency will actually occur are too complex to consider here.[6] In general, regions of other modes crossing the desired mode must be avoided.

There are other modes not yet considered which do not present a sufficiently low conductance to the tube for oscillation to take place (e.g., modes such as the [1, 2] which have low electric fields in the region between the tube grids). These may still couple to the fields of the desired mode and cause discontinuities or distortions in the tuning curves and "holes" in the mode plots.

Two other types of mode interference, harmonic frequency generation and modes with transit times of an integer plus one-quarter cycles, will be mentioned briefly in a later section.

In the design of a waveguide resonator, there is no wide frequency separation between the desired and undesired modes as in the coaxial resonator. Thus, it becomes much more of a problem to keep the desired mode free of interferences. The general treatment is to calculate tuning curves for all the possible modes in the operating frequency range of the tube, and, from them, to calculate a mode plot on which potential interferences may be identified.[7] It then becomes a matter of juggling design parameters to get the desired mode clear of all interferences. A certain amount may be accomplished by the use of dissipative or resonant mode suppressors when a mode interference can not otherwise be removed, but this is usually accompanied by additional problems so it is more or less of a last resort. This will be elaborated upon in the treatment of specific resonators.

Calculation of Tuning Curves

The resonator is not apt to be of sufficiently simple shape to be very exactly treated by analytical methods. However, by considering the cavity boundaries to be perturbed surfaces of one of the common transmission-line types, the various field configurations to be expected may be identified as modifications of known standing-wave patterns. Although these approximations in some cases appear extremely crude, the resulting tuning curves usually check quite well with experimental curves.

The resonator boundaries may usually be approximated by a shorted section of some type of transmission line. The effect of the re-entrant section containing the tube grids is represented by a lumped capacitance and the circuit solved as a transmission-line problem. The value of capacitance to be used is somewhat greater than the calculated dc capacitance of the tube grids alone (not the measured capacitance between grids as this includes the capacitance between the grid-supporting structures which is more properly a part of the transmission line) and is best determined from an experimental tuning curve. Values of 1 $\mu\mu$f for the 2K28 and 0.8 $\mu\mu$f for the Sylvania SD-835G and 6BL6 tubes have been found to give calculated tuning curves which agree quite well with experimental data. However, there is some indication that the value to be used varies somewhat with frequency.[8] A poor guess for this value of capacitance will result in a shifting of the tuning curve to longer or shorter wavelengths (as the capacitance is over- or under-estimated), but the discrepancy will usually be small compared with the error in estimating the capacitance.

Fig. 5 gives tuning curves for a section of waveguide of length l and arbitrarily shaped cross section, loaded with a capacitance C at the center, in terms of dimensionless parameters λ/λ_c, l/λ_0, and λ_0/α where λ_0 is the

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[5] A similar discussion to this would apply to the concentric-line type resonator. Here the degeneracies of interest in the unperturbed resonator arise from crossings of the tuning curves of modes with radial electric field and modes with electric field parallel to the axis. The perturbation introduced by opening a gap in the center conductor for the tube grids results in similar phenomena to those described for the waveguide resonator. However, in this case it is possible for the three-quarter-wave mode to have a range of low shunt resistance. See F. W. Schott and K. R. Spangenberg, "Analogue studies of losses in reflex oscillator cavities," PROC. I.R.E., vol. 37, pp. 1409–1418; December, 1949.

[6] W. H. Huggins, "Multifrequency bunching in reflex klystrons," PROC. I.R.E., vol. 36, p. 624; May, 1948.

[7] For a discussion of these factors in coaxial resonators, see footnote reference 2.

[8] A more exact equivalent circuit for the tube would be the grid capacitance in series with an inductance representing the grid-supporting pedestals. This inductance is not, in general, negligible and results in the necessity for using an equivalent capacitance which is higher than the dc capacitance and more or less dependent on frequency.

cutoff wavelength. The characteristic length α of the particular waveguide section is defined by

$$\alpha^2 = \pi c C \lambda_c Z_{0\infty} \tag{1}$$

where c is the velocity of light in meters per second and $Z_{0\infty}$ is the characteristic impedance of the guide at infinite frequency. For the rectangular waveguide case $\alpha^2 = \pi^2 bC/\epsilon_0$ where b is the guide height, the dimension parallel to the E field, and ϵ_0 is the dielectric constant of free space. The limiting value of resonant wavelength $(l \rightarrow \infty)$ is the cutoff wavelength for all but the quarter-wave mode, where it is given by

$$\lambda_{l\cdot\infty}^2 = \frac{\lambda_c^2}{2} + \left[\left(\frac{\lambda_c^2}{2}\right)^2 + \alpha^4\right]^{1/2}. \tag{2}$$

These curves indicate that for small capacitive loading an error of 100 per cent in the value of C used may result in less than 10 per cent error in the calculated wavelength. For large capacitive loading, the tuning curve is much more sensitive to changes in capacitance.

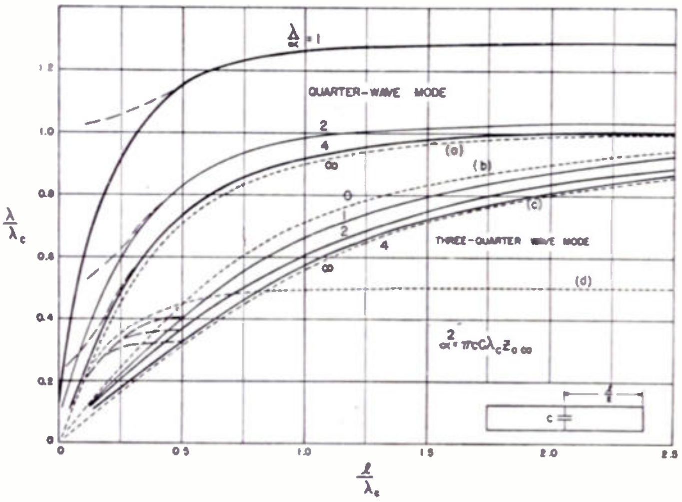

Fig. 5—Universal tuning curves for general waveguide resonator. The dashed curves to the left of $l/\lambda_c = 0.5$ are for rectangular waveguide, in which case the characteristic length α equals $(\pi^2 bC/\epsilon_0)^{\frac{1}{2}}$. Curves (a), (b), (c), and (d) represent the [1, 1], [1, 2], [1, 3], and [2, 1] modes in a shorted waveguide section with no tube present.

The solid curves of Fig. 5 are calculated for a general waveguide with the assumption that the character of the transmission-line mode does not change over the tuning range. While this assumption is probably fairly valid for a ridge waveguide resonator, it does not hold for the rectangular waveguide.

For plunger separations less than the guide width (i.e., for l/λ_c less than 0.5 in Fig. 5) in a rectangular waveguide resonator, the field configurations shift in such a way that the roles of guide length and width are interchanged, causing the tuning curves to take the shape indicated by the dashed lines in Fig. 5. Thus the "three-quarter-wavelength" mode is the [1, 3] mode for l greater than the guide width, but for l less than the width the configuration changes as indicated in Fig. 3(a). Similarly the field configuration of the [3, 1] mode changes according to Fig. 3(b) as the length of the waveguide passes through the square position. As previously discussed, these two modes, which would have the same field configuration in a square resonator with no tube, because of the presence of the tube have different field configurations and resonant frequencies.[9]

The tuning curves may be used to predict possible mode interferences as will be discussed presently. In order to predict holes resulting from nonoscillating modes, it is desirable to compute the tuning curves of these modes. These may be approximated by simply neglecting the effect of the tube-grid capacitance, since it is located in a region of low electric field. If the resonator were perfectly symmetrical, these modes should not couple to the oscillating modes. In the practical case, they usually have some coupling and result in disturbances to the oscillating mode in narrow regions where their tuning curves cross that of the desired mode. These disturbances range from a slight decrease in amplitude to a "hole" in the mode plot and accompanying discontinuity in the tuning curve.

II. Quarter-Wave Mode Resonators

Of the two useful oscillating modes, the quarter-wave mode has a much higher shunt resistance and is relatively free of interference from other modes. It has, however, a limited frequency range. When the shorting plungers are moved all the way up to the tube envelope, representing the highest possible frequency in the quarter-wave mode, (corresponding to a wavelength of about 7 cm in the 6BL6 for example), the tube may still be capable electronically of going to a higher frequency. In order to produce higher frequencies, it is necessary to have an effective short inside the tube walls produced by a physical shorting plunger a half wavelength away—in other words, the three-quarter-wave mode.

The range is also limited on the long wavelength end by the guide cutoff frequency, as seen in the experimental tuning curves of Fig. 6 for an SD-835G tube in resonators with cutoff wavelengths of 10 and 15 cm. The asymptotic values which the resonant wavelength approaches as the plunger separation is increased are about 13 cm and 17 cm, somewhat greater than the cutoff wavelengths. This is characteristic of the quarter-wave mode. At frequencies below the cutoff frequency, a long section of waveguide appears at the end as an inductive reactance which resonates with the capacitance of the tube grids at a frequency somewhat lower than the cutoff frequency.

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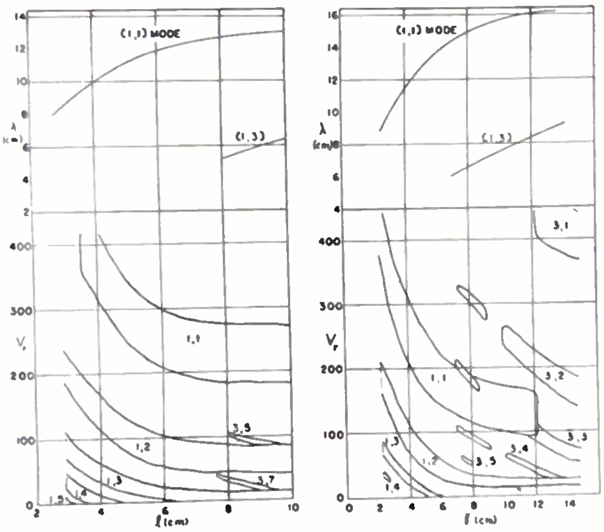
[9] Although the field configurations for the square resonators shown in Fig. 3 appear to bear little resemblance to those for [1, 3] and [3, 1] modes, it will be seen that patterns similar to these result from superpositions of fields of [1, 3] and [3, 1] modes with different phases. In the case of Fig. 3(a), a voltage maximum appears at the center of the pattern and the concentration of electric field between the tube grids results in the resonant frequency being considerably lower than it would be with no tube present. If the [1, 3] and [3, 1] modes are superimposed with one having the opposite phase, the pattern of Fig. 3(b) results, with zero electric field at the center. In this case, the presence of the tube has little effect on the field configuration and resonant frequency, and the conductance is too high for oscillations to occur in this mode in a nearly square resonator.

The mode plots of Fig. 6 show regions of oscillation on a plot with co-ordinates of repeller voltage and resonator tuning. The first of the identifying numbers on the modes indicates the resonator mode, "1" signifying quarter-wave; "3," three-quarter-wave. The second indicates transit time of the electrons in the reflecting space, "1" meaning one and three-quarter cycles, and so on. It is not possible to increase the guide dimensions indefinitely without encountering three-quarter-wave mode interference as shown at the lower right.



(a) (b)

Fig. 6—Experimental mode plot and tuning curves of SD-835G tube in rectangular waveguide resonator ⅜-inch high; (a) guide width, 2 inches; (b) guide width, 3 inches.

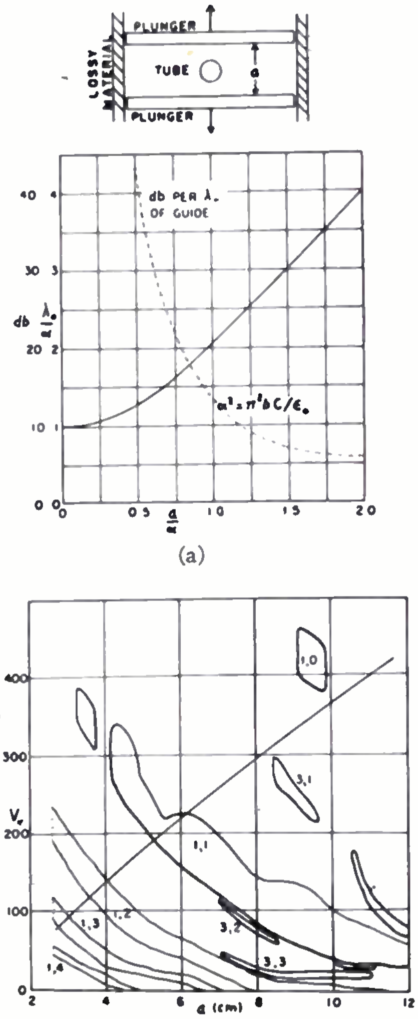
In the design of a rectangular waveguide resonator for a given tuning range, the primary design parameters are guide width and height. The height should be equal to the spacing of the grid contacting rings if the highest frequency of the tube in the quarter-wave mode is to be realized. Increasing the height decreases the resonator loss conductance and allows for better noncontacting plunger design, but it makes interfering resonances more likely to appear. Increasing the width reduces the guide wavelength and hence the plunger travel for a given tuning range, and makes the wavelength-plunger position tuning curve more linear. Here again, however, the limiting consideration is mode interference.

The next higher frequency modes which are likely to be excited by the tube are the [1, 3] and the [3, 1]. Thus the width of the waveguide must be such that the longest wavelength in the desired tuning range is less than the asymptotic oscillation wavelength for infinite plunger separation, and yet, to avoid possibility of mode interference with other modes, the guide should not be wider than about 1.5 times the shortest wavelength at which the tube will oscillate. This may limit the longest wavelength of the tuning range, and for ranges which lie completely below two or three thousand megacycles, the coaxial-line resonator is generally considered to be preferable.

Although these general considerations may be used to choose tentative dimensions, the final test of the design is a calculated mode plot. The generalized tuning curves of Fig. 5 may be used in conjunction with manufacturer's data on the tube modes to give the expected mode plot.

Variable-Width Waveguide Resonator

It may be difficult to obtain a 2:1 tuning range in the quarter-wave mode without running the risk of mode interference from the three-quarter-wave modes as in Fig. 6. An ideal resonator would have only one resonant mode—or at least its higher order modes would have resonant frequencies well beyond the oscillation range of the tube. A very interesting cavity which approaches this ideal consists of a long section of rectangular waveguide whose width is variable, with lossy terminations at both ends (Fig. 7).



(a)

(b)

Fig. 7—Variable-width waveguide resonator. (a) Theoretical tuning curve; (b) experimental mode plot and tuning curve for resonator ⅜-inch high using SD-835G tube.

In connection with the tuning curves of the quarter-wave waveguide resonator, it was shown that as the plunger separation increases to large values the resonant wavelength approaches an asymptotic value. Since the cutoff wavelength of the waveguide is less than the resonant wavelength, the fields die off exponentially with distance away from the tube and, at a sufficient distance, the plungers may be replaced by lossy terminations without appreciably affecting the Q of the cavity. On the other hand, the higher order modes tend to be suppressed by the lossy material. In other words, the dissipative material is placed in a region where the fields of the dominant mode are very weak but the fields of the higher order modes tend to be strong. If there are no reflections from the ends (infinitely long waveguide or perfect terminations), the dominant mode is the only oscillation mode which can exist. In the mode plot shown in Fig. 7, the fragmentary extraneous modes are the result of imperfect terminations, the reflections from which allowed a weak [3, 1] mode oscillation.

A theoretical tuning curve for this resonator is plotted from (2) in terms of the dimensionless parameters λ/α and a/α. This illustrates a property of having the tuning much less crowded toward the high-frequency end of the range than most other types of resonators.

The dotted curve in Fig. 7 gives the db attenuation of the fields per wavelength distance down the waveguide away from the tube, showing the necessity in this resonator of designing for small values of a/α to avoid requiring an excessive length of guide to isolate the terminations from the tube.

Because of the absence of re-entrant areas and the minimum of surface carrying high currents, the Q and shunt resistance of this resonator are quite high.

III. THREE-QUARTER-WAVE RESONATORS

It has been seen that, with present tube dimensions, in order to attain the electronic high-frequency limit of the tubes it is necessary to operate on a resonator mode which has a nodal surface of the E field between the cavity boundaries and the tube grids. The design of the three-quarter-wave mode resonator presents a much more difficult problem than either the quarter-wave mode waveguide resonator or the three-quarter-wave coaxial resonator, chiefly because of the severe quarter-wave mode interference brought about by the nonlinear tuning properties of the waveguide resonator.

As previously discussed, for a plunger separation large compared with the waveguide width the resonant wavelength of the quarter-wave mode approaches a value which is independent of the plunger position. This means that certain regions of repeller voltage will be dominated by the strong quarter-wave mode. On the mode plot these appear as wide horizontal bands which in most cases completely obliterate the desired three-quarter-wave mode in these regions. This tendency appears in Fig. 6.

By increasing the guide width, i.e., cutoff wavelength, the wavelength which the quarter-wave mode approaches can be increased to such a value that oscillations will occur at very low repeller voltages or not at all (because of electrons striking the repeller). The limitation to increasing guide width is again higher order mode interference.

There are several possible ways in which the mode interference problem for the three-quarter-wave resonator might be attacked. The discussion of these methods will use the resonator tuning curves of Fig. 5 to illustrate the means by which the interference-free regions of the desired 3, 2 and 3, 3 modes[10] may be increased. These curves furnish ample assurance that a 2:1 tuning range with no mode interference will not come easily. In order to obtain the 3, 2 mode with no 1, 3 quarter-wave mode interference, it is necessary to have the wavelengths of the two modes in at least the ratio of the cycles of transit time of the two modes, in this case 11:7, over the entire range. But, besides this requirement, it is desirable to avoid the regions of tuning where nonoscillating coupled resonances may cause holes. Most likely of these to cause trouble are the [1, 2] and the [2, 1] (curves b and d in Fig. 5). The crossings of the tuning curves for these modes with that of the [1, 3] are seen to occur when the cavity is nearly square. If they are to be avoided, l must be maintained either greater or less than the guide width over the tuning range.

The latter choice results in a resonator which is essentially a fixed length of waveguide of variable width. The useful tuning range of such a resonator is limited by quarter-wave mode interference at the short wavelength end and by the [1, 2] mode crossing at the other, usually to considerably less than a 2:1 ratio.

Quarter-Wave Mode Suppression

Within the limits set by the desired tuning range, λ_c may be made small so that the portion of the tuning curve used is that to the right of $l/\lambda_c = 0.5$ in Fig. 5. The three-quarter-wave mode is ordinarily used only near the short wavelength oscillation limit of the tube so that this minimizes the possibility of higher order mode interference. It usually results, however, in serious quarter-wave mode interference. A certain amount may be accomplished toward removing such interference by the use of dissipative or nondissipative mode suppressors.

In the coaxial resonator some success was had with nondissipative or tuned mode suppression.[11] In the waveguide resonator the interference is usually much more severe and this method is an unsatisfactory solution.

Dissipative mode suppression is more successful. Reference to Fig. 2(b) will show that there are regions at the sides of a waveguide resonator, opposite the tube, where the fields of the [1, 3] mode are quite weak and the fields of the [1, 1] mode are strong. Dissipative material placed in these regions has the effect of lowering the conductance presented to the tube by the [1, 1]

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[10] These are the most useful modes. Although the 1¾-cycle transit time mode has higher amplitude in general, it requires excessively high repeller voltages at the high-frequency end and its conductance is lower so that it may be unstable with loading.

[11] Radio Research Laboratory Staff, "Very High Frequency Techniques," McGraw-Hill Book Co., Inc., New York, N. Y., art. 32–17, 18; 1947.

mode below a value where the tube will oscillate. At the same time the conductance of the [1, 3] mode is scarcely affected.

Unfortunately these regions move as the resonator is tuned, so that, for a fixed position of the mode suppressors, the tuning range over which the desired modes are not appreciably affected is limited to something like 20 per cent. This suggests arranging to have the mode suppressors moved mechanically to follow the nodal surface as the main shorting plungers are tuned. It is not felt that this method, although quite feasible, offers a very acceptable solution to the mode interference problem.

Resonator Dimensioning

Quarter-wave mode interference may be eliminated by making the guide dimensions large (increasing λ_c and α in Fig. 5). This has the added attraction of giving increased Q and decreased resonator loss conductance. The waveguide width and height must be increased to the point where the desired portions of the 3, 1; 3, 2; or 3, 3 modes are clear of the 1, 1 mode. (We do not concern ourselves with the 1, 0 mode because even if it occurs its conductance is so low that it will disappear with a small amount of coupled load or will be recessive to the desired modes.) Again the height should probably not be over half the width from higher order mode considerations.

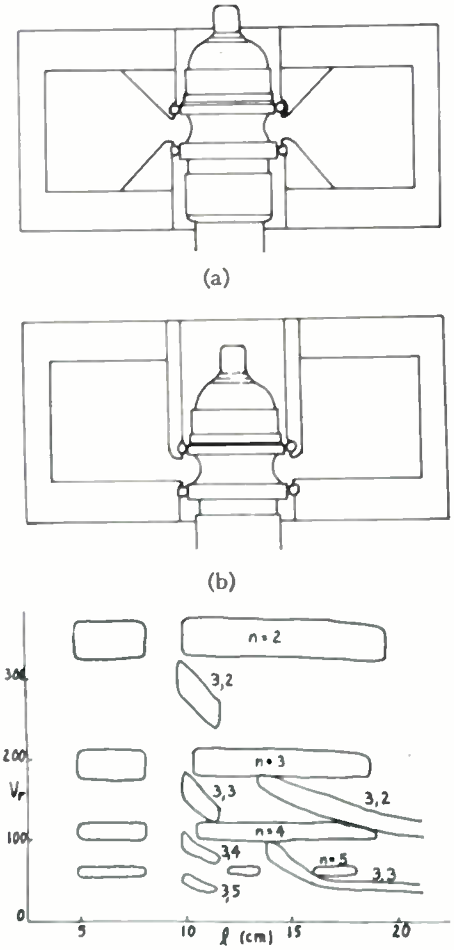

Fig. 8—Tube holders for resonator of Fig. 1; (a) conical tube holder; (b) re-entrant cylinder holder; (c) [3, 1] mode interference with cylindrical holder.

Interference might be anticipated from such modes as the [3, 1] and [3, 3], the next higher frequency oscillating modes. Experimentally this interference proves to be very sensitive to the shape of the tube mount. Fig. 8 illustrates serious [3, 1] mode interference consisting of strong horizontal bands across the mode plot corresponding to a wavelength of about 6.6 cm, almost completely blocking out the desired modes. With a tube holder consisting of two re-entrant cones of approximately 45 degrees cone angle, mode interference from these modes was found to be not objectionable.

The same can not be said for the nonoscillating modes, those which have low electric field density near the position of the tube and hence present too high a conductance for oscillation to occur. In making λ_c large to avoid quarter-wave mode interference, the tuning range has been shifted to the left in Fig. 5 to embrace the region of crossing of the tuning curves for the [1, 2] and [2, 1] modes. Depending on the degree of coupling, these parasitic resonances may or may not give serious trouble. In the case of the resonator of Fig. 1, for which tuning curves and mode plot are shown in Fig. 9, the [2, 1] coupled resonance resulted in only a slight dip in the power output but the [1, 2] caused a hole at a wavelength of 6.2 cm.

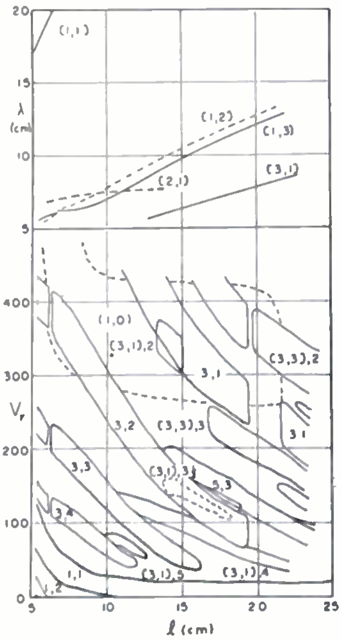

Fig. 9—Experimental tuning curves and mode plot of $3\frac{1}{2} \times 1\frac{1}{2}$-inch rectangular waveguide resonator. Dotted tuning curves represent nonoscillating modes. Tube is SD-835G.

With this resonator the 1, 0 mode appeared in the region indicated by the dotted contour when contacting plungers were used. With noncontacting plungers designed for the short wavelength end of the tuning range, plunger losses were sufficient to prevent its occurring. The [3, 1], $3\frac{1}{2}$ mode is a multifrequency bunching effect discussed by Huggins[6] which disappears with a small amount of loading of the oscillator. This resonator gave satisfactory operation over a 2:1 tuning range except for the hole already mentioned. It had a minimum of mode interference and the desired mode appeared stronger than other modes in the vicinity. The unloaded shunt resistance at a wavelength of 9.6 cm was measured to be 18,000 ohms compared with 5,000 to 10,000 ohms at the same frequency for typical coaxial resonators.

Ridge Waveguide Resonator

Whereas the normal ratio of cutoff wavelengths of the TE_{10} and TE_{30} waveguide transmission modes is 3:1, by the use of ridge waveguide it can be increased to 5:1 or 6:1. This allows designing a resonator with a large λ_c while keeping the range of [3, 1] resonance above the frequency limit of oscillation of the tube. In addition, as previously discussed, for $l/\lambda_c<0.5$, the ridge waveguide tuning curves follow the solid lines in Fig. 5 eliminating the intersection with the [1, 2] mode. Thus, potential mode interference may be restricted to the effects of the coupled [2, 1] mode, which may not be objectionable if the resonator is sufficiently symmetrical to keep the coupling between modes to a low value. The potential tuning range is seen to be well over 2:1.

"Two-Way Stretch" Resonator

So long as λ_c is held constant over the tuning range it has proven very difficult to obtain a 2:1 range completely free of potential mode interference. This consideration suggests varying λ_c and l together so that their ratio stays in an interference-free region in Fig. 5. This is accomplished in a resonator so constructed that the sides move out as the distance between the end shorting plungers is increased.

Harmonic Frequency Generation

As may be seen from Fig. 6, the 1, 1 is by far the most vigorous mode in a waveguide resonator with dimensions small enough to exclude all but the TE_{10} modes. If a cavity which is simultaneously resonant to two frequencies in a 2:1 ratio is operated in this mode, power at the second harmonic frequency is produced by the strong second harmonic content of the sharply bunched electron stream. The amount of second harmonic power obtainable in this manner is quite comparable with that available with the same cavity mode at the same frequency when the mode is excited in the normal manner.

Harmonic frequency generation has certain potential advantages, at least for fixed- or narrow-band tunable-frequency operation. First, the oscillations are exceptionally stable. The presence and frequency of the second harmonic output are not affected by any amount of loading at the second harmonic frequency; the amplitude does depend on the loading because bunching is a nonlinear phenomenon and the amplitude of the output at any frequency is dependent on the bunching voltages of all frequencies. Second, the frequency range of a given tube is extended. For example, output at a wavelength of 3.75 cm was obtained from the third harmonic of a 2K28 operating at 11.26 cm. Normally this tube stops oscillating at 5 cm. Third, the mode interference problem is minimized. Ordinarily the 1, 1 mode is so strong that other modes do not interfere, and, since this mode is not directly loaded, it is quite stable in spite of the low transit time and associated low electronic conductance.

Further discussion and experimental results on these last three types of resonators are given in a previous paper.[12]

IV. CONCLUSION

It does not appear, in general, that the waveguide type of resonator has much to offer in the way of advantages over coaxial resonators for frequency ranges below a few thousand megacycles. At the short wavelength end of the microwave spectrum, however, they are probably superior for most uses. Waveguide resonators and shorting plungers are simple in construction. They are more flexible in design than the coaxial types, offering more latitude in the variation of characteristics by changing design parameters. Designing three-quarter-wave resonators without mode interference is not simple, but the coaxial resonator also has its higher order mode problems. Waveguide resonators have two shorting plungers which must be mechanically connected to move simultaneously.[13]

Considerably more difficult to calculate and measure than tuning curves and mode plots, but nonetheless important in evaluating the merit of a resonator because of this, is resonator loss conductance. A method for determining the loss conductance from the resonator tuning curves and a Q measurement, together with universal curves useful in design, is given in another paper.[14]

At high frequencies the waveguide cavities appear to be somewhat better from the standpoint of loss conductance, both by measurement and as indicated by the frequent presence of the 1, 0 mode and the $n+\frac{1}{4}$ modes (Fig. 9) discussed by Huggins in his multifrequency bunching studies. These are seldom or never observed with the same tubes used in coaxial resonators.

In view of the mode interference problems and relatively high loss conductances associated with three-quarter-wave resonators, consideration might be given to the approach of doing away with the need for them. In other words, a tube could be so designed as to have a mechanical high-frequency limit essentially the same as its electrical limit. This might be accomplished in a tube having a long, narrow rather than circular cross section in which the electrons flow in a thin sheet and the grids are narrow slits. The glass walls could then be

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[12] W. W. Harman, "Tunable waveguide cavity resonators for broadband operation of reflex klystrons," *Proc. NEC*, vol. 4, pp. 233–252; 1948.

[13] Over a 20-per cent bandwidth or so, one plunger only may be moved, the other remaining in a fixed position.

[14] W. W. Harman, "Resonant impedance of transmission lines and waveguides," *Jour. Appl. Phys.*, vol. 20, pp. 1252–1255; December, 1949.

put in such a position as to allow very narrow separation of shorting plungers and corresponding high-frequency operation. Such a tube could be used to advantage in either of the quarter-wave resonators discussed in this paper.

ACKNOWLEDGEMENT

This work was done under the guidance of Karl Spangenberg, to whom appreciation is expressed. Valuable assistance was received from the author's co-workers, J. H. Tillotson and R. C. Honey.

The Effect of a Bend and Other Discontinuities on a Two-Wire Transmission Line*

K. TOMIYASU†, MEMBER, IRE

Summary—**The theoretical problem of a bend on a two-wire transmission line is analyzed by the vector-potential method. The equivalent circuit elements are obtained by comparing the variable-line parameters near the bend with the conventional line parameters found on an infinite line. The effect of the nonrotational symmetry of the axial vector potential on the conductor surfaces is treated by an approximate method to obtain one of the equivalent series elements. Experimental values of the circuit elements for the bend were found to be in good agreement with the theoretical. Good agreement was also found between experimental and theoretical values for open-ended and bridged-end lines.**

INTRODUCTION

THE ANALYSIS of the effect of a single bend on a two-wire transmission line is based upon the condition that the line spacing is very small compared to the wavelength. This implies that radiation from the line is negligible, and that the effect of a bend is nondissipative. Since this effect is dependent upon the distribution of current through the bend, the equivalent network of the bend must contain both series and shunt reactive elements.

FORMULATION OF THE PROBLEM

The analysis of discontinuities on two-wire lines is based upon the differences between the actual distributed line parameters and the corresponding parameters on an infinite line.[1] The equivalent circuit elements of the discontinuity are obtained by integrating the differences over a short distance within the "zone of discontinuity."

In Fig. 1(a), one conductor of radius a is shown bent at an angle of θ degrees. The second conductor of the two-wire line is parallel to the first at a distance b. The co-ordinates w and u have their origins at and are directed away from the bend. The generator is assumed to be at $z=0$. The lengths of the lines L and T are, respectively, s and s_T.

From the electric field vector and the potential equation of continuity the second-order differential equation for the scalar-potential difference on line L has the form[2]

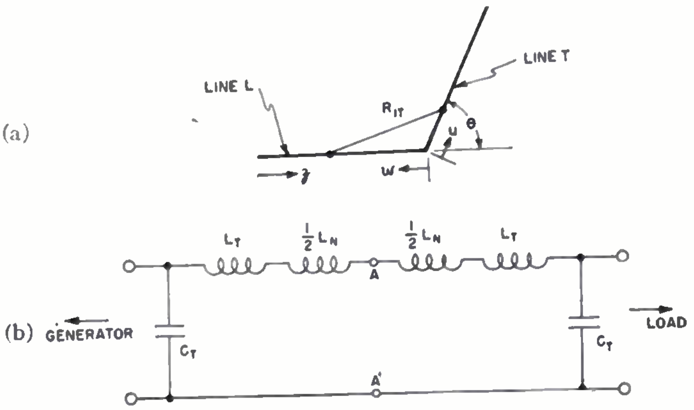

Fig. 1—(a) A two-wire line is bent at an angle of θ. (b) Equivalent circuit for the bend.

$$\frac{\partial^2 V(w)}{\partial w^2} + \beta^2 V_L(w) = \frac{\partial}{\partial w}\left[z^i I_{zL}(w) + j\omega W_{zT}(w)\right]. \tag{1}$$

The angle $\psi(u')$ in [12] and [15b] is explicitly written as θ when computing the vector and scalar potentials. The distances used in this problem are

$$\begin{aligned} R_a &= \sqrt{(w-w')^2 + a^2} \\ R_b &= \sqrt{(w-w')^2 + b^2} \\ R_{1T} &= \sqrt{(w+u'\cos\theta)^2 + (u'\sin\theta)^2 + a^2} \\ R_{2T} &= \sqrt{(w+u'\cos\theta)^2 + (u'\sin\theta)^2 + b^2} \end{aligned} \tag{2}$$

Approximate solutions for [15a]–[15c] are

$$\begin{aligned} k_0(w) &\doteq 2\ln\frac{b}{a} - \ln\frac{w+\sqrt{w^2+b^2}}{w+\sqrt{w^2+a^2}} \\ k_{0T}(w) &= \cos\theta\, k_{0T}'(w) \doteq \cos\theta\ln\frac{w\cos\theta+\sqrt{w^2+b^2}}{w\cos\theta+\sqrt{w^2+a^2}}. \end{aligned} \tag{3}$$

Equations [15d]–[15f] are not used, since this higher-order correction is small.

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* Decimal classification: R117.13. Original manuscript received by the Institute, May 23, 1949; revised manuscript, December 12, 1949. Presented, American Physical Society, June 16, 1949, Cambridge, Mass. The research reported in this document was supported in part jointly by the Navy Department (Office of Naval Research), the Signal Corps of the U. S. Army, and the U. S. Air Force, under Harvard University, O.N.R. Contract N5-ori-76, T.O. 1.

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[1] R. W. P. King and K. Tomiyasu, "Terminal impedance and generalized two-wire line theory," PROC. I.R.E., vol. 37, pp. 1134–1139; October, 1949.

[2] This is eq. [31] of footnote reference 1. Bracketed equation numbers refer to equations in footnote reference 1; those in parentheses, to this paper.

Assuming a lossless line and calculating $W_{zT}(w)$ and $V_L(w)$ using (3), (1) simplifies to

$$\frac{\partial^2 V(w)}{\partial w^2} + \omega^2 c(w) l^e(w) V(w) = 0, \tag{4}$$

where $\phi_1(w)$ and $a_1(w)$ defined in [21] are contained in the terms

$$\begin{aligned} c(w) &= \frac{q_L(w)}{V(w)} = \phi_1(w) c_0(w) = \frac{2\pi\epsilon}{k_0(w) + k_{0T}'(w)} \\ l^e(w) &= \frac{W_z(w)}{I_{zL}(w)} = a_1(w) l_0^e(w) = \frac{k_0(w) + k_{0T}(w)}{2\pi\nu}. \end{aligned} \tag{5}$$

When $w^2 \gg b^2$, the end effect of the line and the coupling effect of the load are negligible and the values of the valuable line parameters in (5) approach the constant values of the line parameters c_0 and l_0^e for an infinite line

$$\begin{aligned} c(w)\big|_{w^2>>b^2} &= \frac{\pi\epsilon}{\ln \frac{b}{a}} = c_0 \\ l^e(w)\big|_{w^2-b^2} &= \frac{\ln \frac{b}{a}}{\pi\nu} = l_0^e. \end{aligned} \tag{6}$$

The circuit elements C_T and L_T in the equivalent network shown in Fig. 1(b) are given by the integrals

$$\left.\begin{aligned} C_T &= \int_0^{d\to\infty} [c(w) - c_0] dw \\ L_T &= \int_0^{d\to\infty} [l^e(w) - l_0^e] dw \end{aligned}\right\}. \tag{7}$$

If the following notations are used

$$\left.\begin{aligned} A &\equiv \ln \frac{x + \sqrt{x^2+1}}{x + \sqrt{x^2 + \frac{a^2}{b^2}}} \\ B &\equiv \ln \frac{x\cos\theta + \sqrt{x^2+1}}{x\cos\theta + \sqrt{x^2 + \frac{a^2}{b^2}}} \\ x &\equiv \frac{w}{b}, \end{aligned}\right\} \tag{8}$$

the integrals (7) may be rewritten as

$$\frac{C_T}{c_0} = b \int_0^{d/b} \frac{A - B}{2\ln \frac{b}{a} - A + B} dx \tag{9}$$

$$\frac{L_T}{l_0^e} = \frac{b}{2\ln \frac{b}{a}} \int_0^{d/b} (B\cos\theta - A) dx. \tag{10}$$

Two C_T's and two L_T's are required in the equivalent circuit since the center of the bend is at AA'. A very small difference in the input impedance is found if both C_T's are lumped together at AA'.

NUMERICAL INTEGRATION FOR C_T AND L_T

The integrals (9) and (10) which have been evaluated by graphical methods for the special case of $b = 2.0$ cm, $a = 0.1588$ cm, and $\theta = 0°$, 30°, 60°, and 90° are tabulated in Table I. The upper limits of the integrals were taken to be $d/b = 10$, since the contributions to the integrals are negligible for $d/b > 10$. The tabulated ratios have the dimensions of centimeters. For this two-wire transmission line, $c_0 = 0.1096$ $\mu\mu$f/cm and $l_0^e = 0.01014$ μh/cm which yields a characteristic impedance of 304 ohms. Note that the integrals (9) and (10) are proportional to line spacing b if b/a is constant.

TABLE I

θ	0°	30°	60°	90°
$\frac{C_T}{c_0}$	0	−0.0163	−0.0714	−0.1889
$\frac{L_T}{l_0^e}$	0	−0.0334	−0.141	−0.353

NONROTATIONAL SYMMETRY

The analysis thus far assumes a rotational symmetry for the vector potential on the conductor surfaces and an abrupt discontinuity in the direction of current at the bend. Strictly speaking, these conditions do not exist. For a nonrotational symmetry of current density in the bend, the effective radius of the conductor is decreased.

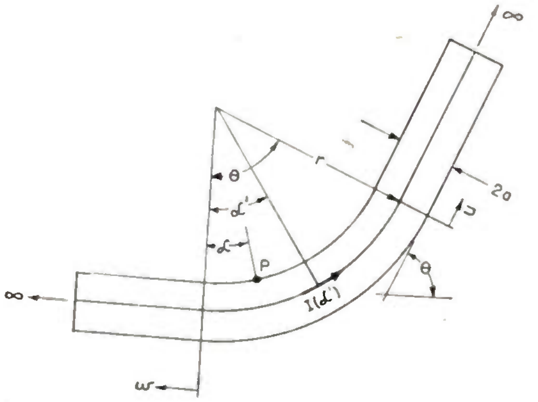

Fig. 2—One line conductor with an arc for the bend.

By considering the bend as an arc of a circle as shown in Fig. 2, the vector potential, and hence the per-unit inductance, can be calculated at all points along the conductor. The second conductor of the two-wire line is parallel at a distance b from the one illustrated. The vector potential is calculated on the inside surface of the bend, where the current density is greatest. If the per-unit inductance is calculated at the center of the arc, where it is largest and assumed to be nearly constant over the length of the arc, the error involved will not be very large since the per-unit inductance at $\alpha=0$ and $\alpha=\theta$ is actually larger than it would have been had the bend made an abrupt angle of θ. With the primed co-ordinates designating the variables of integration and the unprimed co-ordinates the points where the vector potential is computed, and assuming a uniformly distributed filamentary current on the axis of the conductor whose radius has the effective value, the vector potential difference $W(\alpha)$ at a point P (angle α) neglecting retardation is calculated from the following equations:

$$W(\alpha) = W_L(\alpha) + W_a(\alpha) + W_T(\alpha) \tag{11}$$

where

$$W_L(\alpha) = \frac{I \cos \alpha}{4\pi\nu} \int_0^\infty \left(\frac{1}{R_1} - \frac{1}{R_2}\right) dw' \tag{12}$$

$$W_a(\alpha) = \frac{Ir}{4\pi\nu} \int_0^\theta \cos(\alpha - \alpha')\left(\frac{1}{R_3} - \frac{1}{R_4}\right) d\alpha' \tag{13}$$

$$W_T(\alpha) = \frac{I \cos(\theta - \alpha)}{4\pi\nu} \int_0^\infty \left(\frac{1}{R_5} - \frac{1}{R_6}\right) du' \tag{14}$$

and

$$\left.\begin{aligned} R_1 &= \sqrt{[(r-a)\sin\alpha + w']^2 + [r-(r-a)\cos\alpha]^2} \\ R_2 &= \sqrt{R_1^2+b^2} \\ R_3 &= \sqrt{r^2+(r-a)^2-2r(r-a)\cos(\alpha-\alpha')} \\ R_4 &= \sqrt{R_3^2+b^2} \\ R_5 &= \sqrt{[u'+(r-a)\sin(\theta-\alpha)]^2 + [r-(r-a)\cos(\theta-\alpha)]^2} \\ R_6 &= \sqrt{R_5^2+b^2} \end{aligned}\right\}. \tag{15}$$

Equation (13) can be expressed in terms of elliptic functions of the first and second kind, $K(\phi, k)$ and $E(\phi, k)$ respectively

$$\begin{aligned} W_a(\alpha) = \frac{2Ir}{\pi\nu}\Bigg\{&\left[\left(\frac{2}{k_a^2} - 1\right) K(\phi, k_a) - \frac{2}{k_a^2} E(\phi, k_a)\right]_{\pi/2-\theta/4}^{\pi/2} \\ &+ \frac{1}{\sqrt{b^2+(2r-a)^2}}\left[\left(\frac{2}{k_b^2} - 1\right) K(\phi, k_b) - \frac{2}{k_b^2} E(\phi, k_b)\right]_{\pi/2-\theta/4}^{\pi/2}\Bigg\}, \end{aligned} \tag{16}$$

where

$$\begin{aligned} k_a^2 &= \frac{4r(r-a)}{(2r-a)^2} = \sin^2 \beta_a \\ k_b^2 &= \frac{4r(r-a)}{b^2+(2r-a)^2} = \sin^2 \beta_b. \end{aligned} \tag{17}$$

For $\theta=15°$, $\alpha=\theta/2$, and assuming $r=b$ and $b=12.6a$,

$$W_L(\alpha) = W_T(\alpha) = 0.712 \frac{I}{\pi\nu}. \tag{18}$$

In evaluating $W_a(\alpha)$, let us assume the effective conductor radius $a_e=0.643a$. Then the parameter $\beta_a=88.5°$ and $\beta_b=62.75°$. Using tables of elliptic functions,[3]

$$W_a(\alpha) = 1.609 \frac{I}{\pi\nu}. \tag{19}$$

Thus

$$W(\alpha) = 3.033 \frac{I}{\pi\nu} \tag{20}$$

and

$$l\,\big|_{\alpha=\theta/2} = 1.198 l_0^e. \tag{21}$$

This shows an increase of 19.8 per cent in the per-unit inductance. Since the arc length is 0.524 cm,

$$\frac{L_N}{l_0^e} = +0.104 \text{ cm}. \tag{22}$$

The subscript $_N$ is for nonrotational symmetry. (See Fig. 1(b).) When $\theta=30°$ and assuming $a_e=0.643a$,

$$\frac{L_N}{l_0^e} = +0.22 \text{ cm}. \tag{23}$$

For $\theta>30°$, three difficult and important factors become apparent. The radius of curvature r and the effective conductor radius a_e both decrease, which increases the inductance; the shortened current path, on the other hand, decreases the inductance. Because of the complexity of the problem, the value of L_N is assumed constant for $\theta \geqq 30°$.

COMPARISON WITH EXPERIMENTAL VALUES

It is evident from the equivalent circuit shown in Fig. 1(b) that the elements $2L_T+L_N$ and $2C_T$ can be measured experimentally by letting the load assume either very low or very high impedances. This range in load impedance can be obtained by adjusting the length of shorted line beyond the bend.

Since the line was unloaded, only the phase functions

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[3] A. M. Legendre, "Traité des Fonctions Elliptiques," vol. 2, Paris, Imprimerie de Huzard, Coucier; 1826.

were measured by the resonance-curve method. For small values of $2C_T$ and $2L_T+L_N$, as expected here, the ratios $2C_T/c_0$ and $(2L_T+L_N)$ ratio l_0^e were simply the change in over-all length of line to give resonance. It was found that small differences in the location of the distribution extremes at the center of the bend did not affect the measured values. Comparison between the theoretical and experimental values are shown in Figs. 3 and 4. The agreement in $2C_T$ and $2L_T+L_N$, which is fairly good over the whole range, is within the radius of the line conductor.

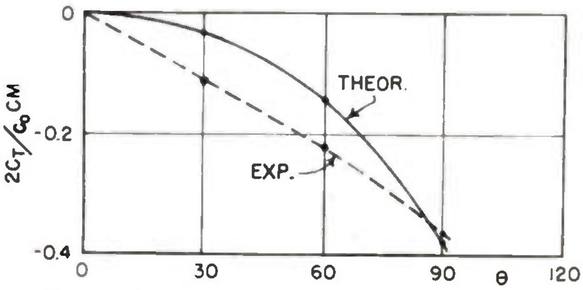

Fig. 3—Comparison of $2C_T/c_0$ as a function of θ.

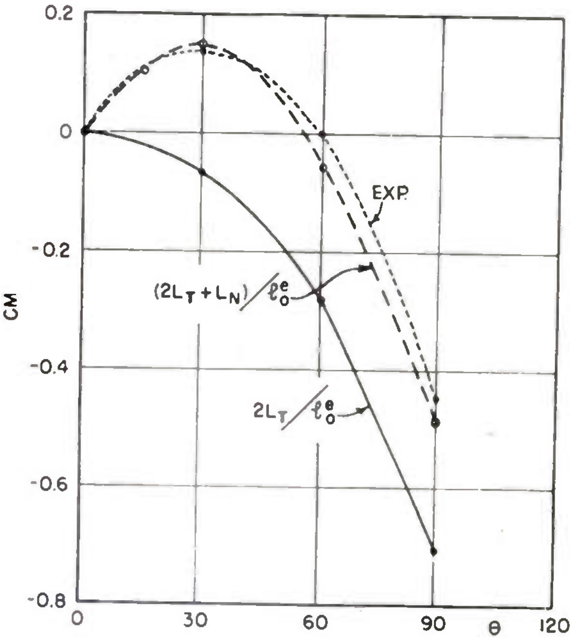

Fig. 4—Comparison of $2L_T/l_0^e$ and $(2L_T+L_N)/l_0^e$ as functions of θ.

The measurements were made at a frequency of 300 Mc by using a very sensitive crystal-af amplifier detector. The measurements were readable to within 0.02 cm and reproducible to within 0.05 cm.

Open-ended and Bridged-end Lines

For an open-ended line, C_T was computed by setting the quantity B in (9) equal to zero. Evaluating the integral numerically gave a value of $C_T/c_0=0.416$ cm. Open-ended lines differing by the end surfaces on the conductor were measured experimentally. These results are tabulated in Table II: (a) flat closed ends on the conductors, (b) hemispherical ends, and (c) open tubing. It appears that the innersurfaces of the open tubing are being charged, whereas fringing is at a minimum for the hemispherical ends. Since the currents do not vanish except at the center of the ends of the conductors, the length of the conductors should perhaps be measured along the surface of the conductor and to this point. By adding the radius of the conductor, the corrected theoretical value of C_T/c_0 becomes 0.575 cm. The reference plane was at the end of the conductors.

TABLE II

Flat closed ends	0.60 cm
Hemispherical ends	0.58 cm
Open tubing	0.65 cm
Theoretical (corrected)	0.575 cm

The end effect of a bridge on a two-wire line is considered by adding two inductances in series, L_T and L_S, where L_T is the value given in Table I for $\theta=90°$ and L_S the self-inductance of the bridge given by the equation[4]

$$L_S = \frac{b}{2\pi\nu}\left[\sinh^{-1}\left(\frac{b}{a'}\right)+\frac{a'}{b}-\sqrt{1+\frac{a'^2}{b^2}}\right]. \quad (24)$$

Three values of a', radius of the bridge wire, were chosen, 0.0794, 0.1588, and 0.2382 cm, and L_S was calculated for each.

Comparison of theoretical and measured values of $(L_S+L_T)/l_0^e$ are plotted in Fig. 5 for three radii of the bridge wire. The agreement is good for the smaller radii. The reference plane was at the center of the bridge.

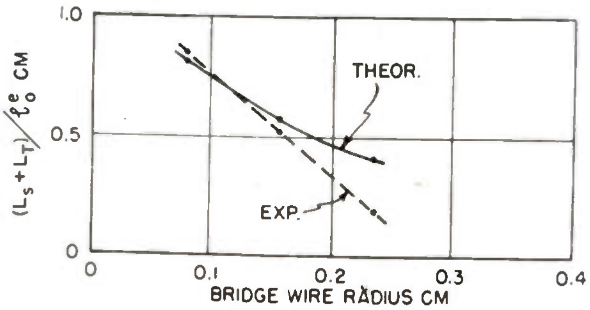

Fig. 5—Comparison of $(L_S+L_T)/l_0^e$ as a function of bridge wire radius.

For the bridge with the largest radius, the radius of curvature r and effective conductor radius a_e are larger and the current path is shorter in proportion than for the two smaller bridge radii. This means that the actual L_S is smaller for the largest bridge radius and explains the departure shown in Fig. 5. The effect of L_N is contained in the modified L_S.

Acknowledgment

The author wishes to thank R. W. P. King for valuable suggestions, and Edith Stokey for performing the numerical integrations, necessary for the publication of this paper.

[4] R. W. P. King, "Electromagnetic Engineering," vol. I, chap. VI, McGraw-Hill Book Co., New York, N. Y.; 1945.

Discussion on

"An Approach to the Approximate Solution of the Ionosphere Absorption Problem*"

JAMES E. HACKE, JR.

Norman Balabanian:[1] In the discussion immediately preceding equation (16) in his paper, Mr. Hacke states: "It (the apparent height of reflection x') can be obtained by integrating, over the upward path, the group velocity of the wave divided by the velocity of light. In the x co-ordinates used here, since the group velocity is the velocity of light divided by the index of refraction,

$$x' = x_L + \int_{x_L}^{x_0} \frac{dx}{\mu}. \qquad (16)\text{"}$$

The definition given for group velocity is actually that for phase velocity. Phase velocity in a medium is defined as the velocity of light c, divided by the refractive index of the medium $v_p = c/\mu$. From equation (12) of the paper, μ is less than unity and hence v_p is greater than c. In the ionosphere, group and phase velocities are related by the following expression:

$$v_g = c^2/v_p.$$

Since v_p is greater than c, v_g must be less than c. Also, substituting $c/v_p = \mu$, $v_g = \mu c$.

Mr. Hacke proposes to obtain x' by integrating v_g/c over the upward path. But v_g/c is less than unity everywhere along the path. Hence, the integration would yield an apparent height less than the actual height. This is obviously erroneous.

Actually, the apparent height can be obtained by integrating c/v_g over the upward path. This can easily be shown to be true. Since a pulse of energy travels at approximately the group velocity, the time it takes a pulse to traverse the upward path is given by the integral

$$t = \int_0^{x_0} \frac{dx}{v_g}.$$

Multiplying both sides of this equation by c yields

$$ct = \int_0^{x_0} \frac{c}{v_g}\, dx.$$

But the left-hand side of this equation is the apparent height. Hence, the integral on the right must also give the apparent height. Substituting $v_g = \mu c$,

$$x' = ct = \int_0^{x_0} \frac{dx}{\mu}.$$

This is the same as equation (16) in the paper.

Attention should also be called to equation (4). Substitution of equation (3) into equation (2) will yield

$$N = \frac{N_0}{\sqrt{\sec x}} \exp \tfrac{1}{2}(1 - x - \epsilon^{-x}).$$

This error might be attributed to the printer. However, equation (4) appears in the same form in a later paper.

A. H. Waynick:[2] Since Mr. Hacke has not been engaged in ionospheric work for over a year, I am taking the liberty of attempting to answer Mr. Balabanian's comments on his paper for him.

In the case of the simplified dispersion equation used by Mr. Hacke, whose real part is his equation (12), it is readily shown that

$$\frac{c}{U} = \mu + \omega \frac{\partial \mu}{\partial \omega} = \mu' = \frac{1}{\mu}. \qquad (1)$$

Here

c = velocity of light in a vacuum
U = group velocity
μ = phase refractive index
ω = angular wave frequency
μ' = group refractive index.

Now

$$P' = c\Delta t = c \int \frac{ds}{U}.$$

From the above relation this may here be written

$$P' = \int \frac{ds}{\mu}$$

or, in x height units,

$$= \int \frac{dx}{\mu}$$

as used in Hacke's (16).

Here $P' = 2h'$ = group path or, following Hacke's terminology, apparent path.

I believe the above illustrates:

(A) The wording of the two sentences quoted from Mr. Hacke's paper is incorrect as indicated in the discussion.

(B) However, the Hacke equation (16), which was based on relations (1) above, is correct. This is proven by similar reasoning in the discussion.

(C) While the words of (A) are incorrect, the mathematics of (B) is correct as covered above. The discussion comment "Hence, the integration would yield an apparent height less than the actual height" is covered

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* Proc. I.R.E., vol. 36, pp. 724–728; June, 1948.

[1] Syracuse University, Syracuse, N. Y.

[2] Pennsylvania State College, State College, Pa.

in Figs. 4 and 3 of the paper. These illustrate that, for any given R, the height of apparent reflection, *measured down from the level of maximum ionization in the layer*, is less than the true height. Consequently, the group height is shown to be greater than the true height by the figures included in the paper.

The comments concerning Hacke's equation (4), and the identity (5), are obviously correct.

I do not know whether Mr. Balabanian will not wish to have his discussion published in view of the above comments. The error in equation (4) is so obvious, it alone would not appear to warrant publication of its correction.

However, for Mr. Balabanian's information, the "P_1 alone" curve of Fig. 5 has recently been found to be in error due to computing mistakes. The balance of this figure, however, is correct and the conclusions drawn concerning the invalidity of using the single parabola approximation in absorption calculations are still valid.

Norman Balabanian:[1] I wish to thank Professor Waynick for his comments on the discussion.

Professor Waynick apparently agrees with the writer's original discussion except for his point (C). The writer's comment, "Hence the integration would yield an apparent height less than the actual height" applies to Mr. Hacke's intended integration, "*over the upward path*, of the group velocity of the wave divided by the velocity of light." Here there is no question of measuring height downward from the level of maximum ionization in the layer. In Professor Waynick's notation,

$$h' = \frac{P'}{2} = c\,\frac{t}{2}\,.$$

Obviously, h' is measured upward from the surface of the earth. However, Professor Waynick's statement concerning the height of apparent reflection, measured down from the height of maximum ionization, is certainly true.

James E. Hacke, Jr.:[3] Thanks are due Mr. Balabanian for pointing out two careless errors in my paper, "An Approach to the Approximate Solution of the Ionosphere Absorption Problem."

1. Preceding my introduction of equation (16), I wrote,

"In the x co-ordinates used here, since the group velocity is the velocity of light divided by the index of refraction . . . . " This is, of course, an error; the group velocity is the velocity of light *times* the index of refraction

$$v_g = \mu c = dx/dt; \tag{15a}$$

$$dt = dx/(\mu c). \tag{15b}$$

When this is divided by the velocity of light and integrated over the path, the result is the apparent or group height, as noted by Mr. Balabanian and implied in the preceding statement from the paper, quoted by Mr. Balabanian: the apparent height " . . . can be obtained by integrating, over the upward path, the group velocity divided by the velocity of light." What Mr. Balabanian is apparently trying to say in the fourth paragraph of his note is that if I had actually integrated using c/μ, and not μc, I should have got a group height differing in the wrong direction from the geometric height.

2. Mr. Balabanian is quite right that equation (4) should read,

$$N = (N_0 \div \sqrt{\sec \chi}) \exp \tfrac{1}{2}(1 - x - \epsilon^{-x}), \tag{4}$$

with appropriate changes in equation (5). This is the way the equations appear in my notes; the error was evidently made in preparing the typescript. A similar error was made in equation (1.5) of a subsequent paper, "An Approximate Solution of the Problem of Path and Absorption of a Radio Wave in a Deviating Ionosphere Layer"; this equation is part of a summary of the results of the present paper, and was made from the typescript.

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[3] 600 Haven St., Evanston, Ill.

Correspondence

Note on Low-Noise Figure Input Circuits*

In the excellent paper "Design Factors in Low-Noise Figure Input Circuits," by M. T. Lebenbaum,[1] is found the following equation (Table I):

$$M = (\lambda_1\lambda_2 - \lambda_3)^{1/2}.$$

There are many cases where $(\lambda_1\lambda_2 - \lambda_3)$ is a small difference of two large quantities, and therefore a slide-rule calculation of M by this equation is subject to considerable error. This difficulty can be avoided as follows, referring to Table I,

$$\lambda_2 = \frac{1}{C_2\omega_0^2}\left(1 + \frac{\alpha_2^2}{8} + \frac{3\alpha_1^2}{8}\right)$$

$$= \frac{1}{C_2\omega_0^2}(1 + \Delta_2)$$

$$\lambda_1 = \frac{1}{C_1\omega_0^2}\left(1 + \frac{\alpha_1^2}{8} + \frac{3\alpha_2^2}{8}\right)$$

$$= \frac{1}{C_1\omega_0^2}(1 + \Delta_1)$$

where the meaning of Δ_1 and Δ_2 is obvious.

$$\lambda_1\lambda_2 = \frac{1}{C_1C_2\omega_0^4}(1 + \Delta_1)(1 + \Delta_2)$$

$$= (\lambda_3)(1 + \Delta_1 + \Delta_2 + \Delta_1\Delta_2).$$

$$\lambda_1\lambda_2 - \lambda_3 = (\lambda_3(\Delta_1 + \Delta_2 + \Delta_1\Delta_2).$$

Taking the root, we have

$$M = \sqrt{\lambda_3}\sqrt{\Delta_1 + \Delta_2 + \Delta_1\Delta_2}.$$

This equation can be used to determine M but usually simplification is possible, namely,

Case 1:

$$\alpha_1 < 0.5; \qquad \alpha_2 < 0.5.$$

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* Received by the Institute, February 1, 1950.

[1] M. T. Lebenbaum, "Design factors in low-noise figure input circuits," Proc. I.R.E., vol. 38, pp. 75–80; January, 1950.

Correspondence

Then $\Delta_1\Delta_2$ is negligible compared to $+\Delta_2$.

$$M = \sqrt{\lambda_3(\Delta_1 + \Delta_2)}$$

$$= \sqrt{\frac{\lambda_3}{2}\,(\alpha_1^2 + \alpha_2^2)}$$

ccurate to one per cent if α_1 and α_2 are each ss than 0.3.

Case 2: Primary loading only

$$M = \alpha_1\sqrt{\frac{\lambda_3}{2}} \quad \text{(accurate to one per cent if } \alpha_1 < 0.5).$$

Case 3: Secondary loading only

$$M = \alpha_2\sqrt{\frac{\lambda_3}{2}} \quad \text{(accurate to one per cent if } \alpha_2 < 0.5).$$

Incidentally, I believe that in the second quation of Table I, $\sqrt{2f}$ is a misprint for $\sqrt{2b}$. [This has been pointed out by the uthor in a separate communication.]

A. C. Hudson
Microwave Section
National Research Council
Ottawa, Canada

The Diurnal Variation of the Vertical Incidence Ionospheric Absorption at 150 Kc*

It is the purpose of this letter to present some recent experimental information on absorption at low frequencies.

The Radio Propagation Laboratory of the Pennsylvania State College has been conducting a program of measurements of the ionospheric absorption of radio waves at vertical incidence at 150 kc. Both the experimental and theoretical phases of this program have been in progress since March, 1949. The measurements were taken by manual and semiautomatic receiving methods on pulses from a one-half megawatt pulse transmitter at a low repetition rate.

During the summer months, because of insufficient system gain and high absorption, the absorption records were incomplete for several hours around local noon. However, as the season progressed, the records became more nearly complete. A number of complete daily records have been obtained in recent months. One of these is shown in Fig. 1. This figure is a plot of $|\log \rho|$ against time for August 20, 1949. The curve was obtained by the method of Appleton[1] from the observed values of the virtual height of the layer h, the relative amplitudes of the ground pulse G, and the first and second echoes E and E'. Then $\rho = (2hE/kG)$ with $k = hE^2$ $0.96GE$ where the 0.96 factor enters due to ground reflection. This curve is a running mean of the instantaneously observed values of the echo amplitude. Of particular interest is the marked correspondence of ground sunset and sunrise with the absorption curve transitions.

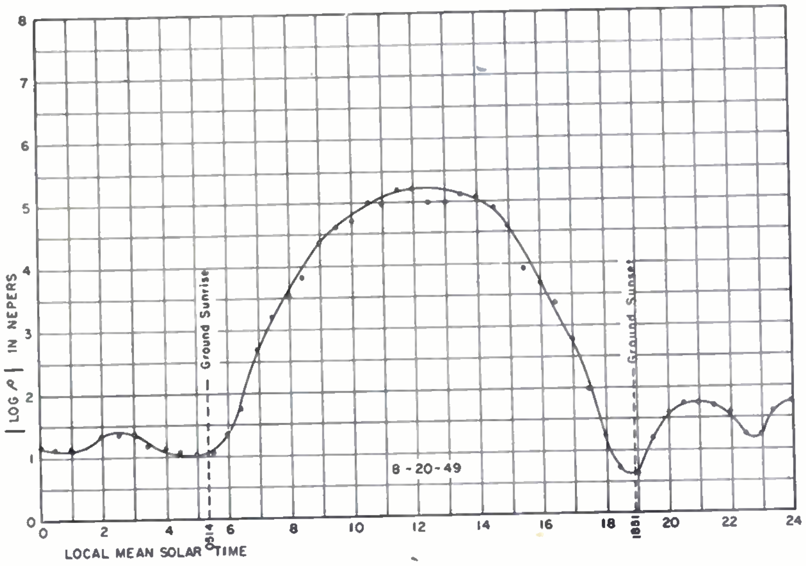

Fig. 1—A plot of $|\log \rho|$ against time at vertical incidence at 150 kc for August 20, 1949.

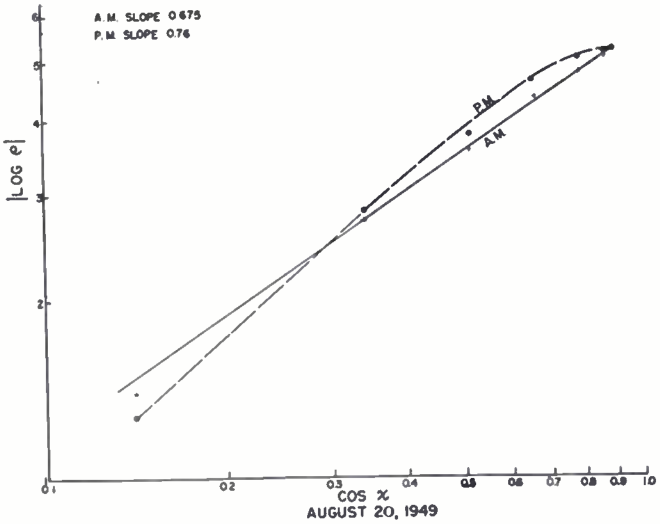

Fig. 2—Plot to determine exponent of $\cos^n x$ for particular case of August 20, 1949.

A further interesting feature of this curve is the slope in terms of the sun's zenith angle. Fig. 2 shows a plot of $\log |\log \rho|$ vs. $\log \cos \chi$. This gives a morning slope of 0.675 and on the linear portion of the afternoon curve a slope of 0.76. These numerical values obviously apply as the exponent in $|\log \rho| \alpha \cos^n \chi$ for the relevant periods.

Other similar records are now being analyzed to obtain a statistical treatment of the data. Preliminary comparison of the experimental and theoretical results shows a rather close agreement as regards total absorption. A paper covering the complete program of investigations will be published at a later date. This work is being conducted with the support of the Geophysical Research Directorate of the Air Force Cambridge Research Laboratories, AMC, USAF.

A. H. Benner
Radio Propagation Lab.
Pennsylvania State College
State College, Pa.

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* Received by the Institute, October 24, 1949.

[1] E. V. Appleton, "Regularities and irregularities in the ionosphere," *Proc. Royal Soc.*, vol. 162, October, 1937.

Contributors to Proceedings of the I.R.E.

Ross Bateman

Ross Bateman (A'42) was born in Toledo, Ore., on July 22, 1912. He received the B.S. degree in electrical engineering from the Oregon State College in 1934. From 1934 to 1937 he was employed by the American Pacific Whaling Company and the Alaska Steamship Company in connection with the installation and operation of land and ship radio stations. Following this, from 1937 to 1942, he was associated with the Federal Communications Commission. Mr. Bateman joined the staff of the operational research group in 1942, in the office of the Chief Signal Officer, serving there until 1945. He is now employed by the National Bureau of Standards as chief of the ionospheric research section at the Central Radio Propagation Laboratory, in Washington, D. C.

⁂

K. E. Burmaster

Kenneth E. Burmaster (M'48) was born in Ransomville, N. Y., on January 29, 1912. He attended Niagara University for a short time and then became engaged in the electronic and electrical engineering field for several years. Then, he entered the employ of Oldbury Electro-ChemicalCo., Niagara Falls, N. Y., in 1937. Here he was associated with electrical maintenance of industrial equipment, including electrical machinery for generation, conversion, and distribution of electric power, and also the operation and maintenance of megawatt arc furnaces and associated control equipment.

Mr. Burmaster attended Niagara University in 1943 and 1944 to study electronics and electrical engineering. He accepted a position during August, 1944, with Carbide and Carbon Chemicals Corporation at Oak Ridge, Tenn., supervising maintenance of mass spectrometers for the Instrument Division. In 1945, he joined the Laboratory Division of Carbide and Carbon Chemicals Corporation at Oak Ridge K-25 Plant and supervised the organization of electronics, vacuum tube, machine shop, and engineering drafting groups for development and maintenance of laboratory analytical instruments. At present, he is assistant laboratory department supervisor, supervising groups engaged in design, construction, and maintenance of laboratory instruments and equipment associated with the operation and development of the gaseous diffusion process for separating uranium isotopes, including mass spectrometers for isotope measurements, spectrophotometers, and X-ray analyzers

Mr. Burmaster is a member of the American Institute of Electrical Engineers and the Instrument Society of America, of which he is the vice president of Oak Ridge Section.

⁂

L. A. Byam, Jr.

L. A. Byam, Jr., was born on August 19, 1906, in Chelmsford, Mass. After serving for one year as Morse operator with the Western Union Telegraph Company and three years as a radio operator with the Radiomarine Corporation, Mr. Byam entered the University of Delaware to study electrical engineering. He was graduated in 1932 with the B.E.E. degree. Following graduation, he enrolled in a special course of study at the Massachusetts Institute of Technology, completing the course in 1933. Rejoining Western Union in July, 1933, he has been associated with that company up to the present time.

On January 1, 1937, Mr. Byam was appointed division manager in charge of operation, and served in that capacity until May, 1941, when he was called to active service with the Navy. During the war he was engaged in radar installation and maintenance under the direction of the Bureau of Ships. He resumed service with Western Union in January, 1946, and is now engaged in radio research with the development and research department. He is a member of Phi Kappa Phi and Tau Beta Pi.

⁂

R. E. Corby

Robert E. Corby was born in 1916 in Freeport, N. Y. He was graduated from Rensselaer Polytechnic Institute in 1940 with the B.E.E. degree, and received the M.E.E. degree the following year from the same institution. During the war years he was an instructor and later an assistant professor at Rensselaer. He also did consulting work for the Sprague Electric Company, in North Adams, Mass. In 1945 he was awarded the Ph.D. degree in physics from Rensselaer.

Dr. Corby joined the staff of the University of Arizona in 1946, and at present he is an associate professor of physics. He is a member of Sigma Xi, Pi Mu Epsilon, and the American Physical Society.

Leo E. Dwork

Leo E. Dwork (M'46) was born in New York, N. Y., on August 31, 1920. He received the B.E.E. degree in 1941 from the College of the City o New York, and the M.E.E. degree in 1948 from the Polytechnic Institute of Brooklyn. In the summer of 1941 he was employed by the Kenyon Transformer Company. During 1941 and 1942 he was successively employed by the Babcock and Wilcox Company, and the Brooklyn Navy Yard. He taught electrical engineering at the College of the City of New York from 1942 to 1944, and served in the U. S. Navy from 1944 to 1946. He is at present senior instructor in charge of the RCA Institutes electrical technology department, having joined the staff in 1942.

Mr. Dwork is an associate member of the American Institute of Electrical Engineers, a member of the Acoustical Society of America and Tau Beta Pi, and an associate member of Sigma Xi.

⁂

J. J. Ebers

J. J. Ebers (S'46–A'48) was born on November 25, 1921, in Grand Rapids, Mich. He received the B.S. degree from Antioch College in 1946, his education having been interrupted by three years' service in the U. S. Army. He obtained the M.S. degree in electrical engineering from Ohio State University in 1947. Since 1947, he has been an instructor in the electrical engineering department of this University, as well as a research associate for the Ohio State University Research Foundation. His research activities have been in the field of high-frequency oscillators.

Mr. Ebers is a member of Eta Kappa Nu, Sigma Xi, and the American Physical Society.

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For a photograph and biography of W. W. Harman, see page 1345 of the November, 1949, issue of the Proceedings of the I.R.E.

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For a photograph and biography of K. Tomiyasu, see page 1156 of the October, 1949, issue of the Proceedings of the I.R.E.

Ernest C. Evans (S'44–A'46) was born n July 19, 1920, in New York, N. Y. After ttending the New Jersey College of Commerce and studying radio engineering at Rutgers University, he was engaged for a short time in instrument work at the Signal Corps General Development Laboratories at Red Bank, N. J. Following this he entered Massachusetts Institute of Technology for further study in electrical engineering, leaving in 1944 to work on the atomic bomb project in the Army. After his discharge in 1946, he was employed by the Laboratory Division of Carbide and Carbon Chemicals Corporation at Oak Ridge, Tenn., as a research engineer. At the present time, he is a development engineer for this Corporation and is engaged in design and development work on mass spectrometers, laboratory instruments, electronic controls, and electromagnets. He is at present engaged in graduate studies at the University of Tennessee.


E. C. Evans

❖

T. R. Finch (M'46) was born on December 26, 1914, in Colorado Springs, Colo. He attended the University of Colorado where he received the B.S. degree in 1938 and the M.S. degree in 1939. Since then, Mr. Finch has been employed by the Bell Telephone Laboratories, where he has been primarily interested in the design and development of transmission networks. He is at present engaged in the development of networks for wide-band coaxial repeater amplifiers.


T. R. Finch

❖

Edwin F. Florman (A'41) was born on February 16, 1904, in Venice, Ill. He worked as an electrician in a steel mill, during which time he completed high school and undergraduate college work. In 1932 he was graduated from Washington University with the B.S. degree in electrical engineering, and in 1934 he received the M.S. degree in electrical engineering and physics from the same institution. From 1934 to 1941 he was employed as research physicist at the Western Cartridge Company of East Alton, Ill., in charge of the development of electronic ballistics-testing equipment. He was a radio engineer at the National Bureau of Standards during 1941–1944, in charge of completing a radio device for measuring upper-air wind velocities up to altitudes of 100,000 feet.

From 1944 to 1946 Mr. Florman was associated with the Philco Corporation in the research department. In 1946 he returned to the National Bureau of Standards, where he is at present conducting low-frequency radio-wave-propagation experiments in the Central Radio Propagation Laboratory. Mr. Florman is a member of Tau Beta Pi and Sigma Xi.


Edwin F. Florman

❖

Keith C. M. Glegg (S'47) was born in Kingston, Jamaica, B.W.I., on January 7, 1926. He received the B. Eng. degree in engineering physics in 1947 and the M.Sc. degree in physics in 1949, both from McGill University. While working for M.Sc. degree, Mr. Glegg was a demonstrator in electrical engineering at McGill University.

Since 1949 Mr. Glegg has been engaged in radar development at the Canadian Marconi Company. He is a member of the Xi.


Keith C. M. Glegg

❖

Oskar Heil was born in Germany on March 20, 1908. He received the Ph.D. at the University of Göttingen in 1932 and did research at the University of Göttingen, in the Cavendish Laboratory, Cambridge, England, and with Standard Telephones and Cables, C. H. Lorenz, A.G., Julius Pintsch K.G. and Telefunken. At present, he is working for the government and engaged in research and development work in the Tube Laboratory at Ohio State University.


O. Heil

❖

George A. Hufford (S'45–A'48) was born on June 1, 1927, in San Francisco, Calif. He received the B.S. degree in engineering from the California Institute of Technology in February, 1946, and the M.S. degree in electrical engineering in 1948 from the University of Washington. Since his graduation he has been with the Central Radio Propagation Laboratory of the National Bureau of Standards, studying various problems in the theory of wave propagation.


G. A. Hufford

Fred J. Kamphoefner (S'43–A'45–M'46) was born in San Francisco, Calif., on March 23, 1921. He received the B.S. degree from the University of California in 1943, then joined the Radio Research Laboratory at Harvard University as a research associate, in the receiver group. In 1946 the Laboratory was disbanded and he resumed studies at Stanford University while doing part-time teaching and research, concerned mainly with uhf oscillators and apparatus for the measurement of ion densities in the atmosphere.

Dr. Kamphoefner received the Ph.D. degree in electrical engineering in 1949, and soon after became a staff member of the Stanford Research Institute. In his present position, his principal interest is industrial instrumentation. He is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.


F. J. Kamphoefner

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Joseph Leferson (A'48) was born in Point Pleasant, N. J., on August 1, 1914. He attended Monmouth College and graduated from the New York Electrical School. A licensed radio amateur since 1932, he now holds both amateur and commercial radio licenses. In 1937, he became associated with Machlett Laboratories, Inc., and is now project engineer on power and ultra-high-frequency tubes.


J. Leferson

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J. D. McGee was born on December 17, 1903, near Canberra, Australia. Following his graduation from St. Patrick's College, Goulburn, he entered Sydney University in 1923 and received the B.S. degree in mathematics and physics in 1927, and the master's degree in physics in 1928. At that time, he was awarded the University Medal in physics, as well as an 1851 Exhibition Scholarship to Clare College, Cambridge, England, where he worked in the Cavendish Laboratory from 1928 to 1931 on atomic physics. He received the Ph.D. degree in 1931.

In January, 1932, Dr. McGee joined the staff of the E.M.I.. Research Laboratories, and has since worked principally on the development of television pickup tubes and similar electronic devices. During the war, Dr. McGee was occupied with the development of electronic infrared image converters and the instruments in which they were used.


J. D. McGee

J. Z. Millar (A'30–SM'45) was born in Matton, Ill., on July 3, 1901. Following his graduation from the University of Illinois in 1923, he joined the Western Union Telegraph Company as an engineering apprentice, transferring in 1926 to the Water Mill laboratory to specialize in electronics. After doing research on shortwave equipment and audio-frequency apparatus for fifteen years, Mr. Millar was called to active duty with the Signal Corps, in which he attained the rank of colonel. He served as member and director of the Signal Corps Board in Ft. Monmouth, N. J., from March, 1941, to April, 1944, and was then assigned as Signal Officer of the Normandy Base section and as Signal Officer, Loire Section, European Theater.


J. Z. Millar

On February 15, 1945, Mr. Millar returned to Western Union, and was appointed radio research engineer. In this position he has organized the radio research division of the development and research department. On August 1, 1949, Mr. Millar was made director of research.

⁘

G. Franklin Montgomery (M'48) was born in Oakmont, Pa., on May 1, 1921. He graduated from Purdue University with the B.S.E.E. degree in 1941. From 1941 to 1944, he worked at the Naval Research Laboratory on radar and similar projects. During the period 1944 to 1946, he was in the U. S. Army Signal Corps, working on ionospheric research during part of that time. Since 1946, he has been employed by the Central Radio Propagation Laboratory, National Bureau of Standards.


G. F. Montgomery

Mr. Montgomery has been a licensed radio amateur since 1935. He is a member of Tau Beta Pi, Eta Kappa Nu, Sigma Pi Sigma, and Sigma Xi.

⁘

I. E. Mouromtseff (A'34–SM'45–F'47) was born on December 9, 1881, in St. Petersburg, Russia. After he received the Mil. Eng. degree he continued his study in electrical engineering and radio at the Institute of Technology, in Darmstadt, Germany, and was graduated in 1910 with the Dipl.-Eng. degree. Professor Mouromtseff was assistant chief of the Electrical School for Army Officers, in charge of academic and technical training, during the years 1910 to 1917. He was also a member of the Franco-Russian Radio Committee. He was presented with the Officer's Cross of the Legion of Honor by the French Army Command in 1916.


I. E. Mouromtseff

Professor Mouromtseff came to the United States in 1917 as a member of the Division of Supplies, attached to the Russian Embassy, in Washington, D. C. From 1923 until 1936 he was engaged as a research engineer in charge of transmitting electron-tube development for the Westinghouse Electric and Manufacturing Company, in East Pittsburgh, Pa. He later transferred to the lamp division of Westinghouse, where he was in charge of uhf and microwave electron-tube development in 1936, remaining until 1942, when he was made assistant manager of the electronics department. In 1947, Professor Mouromtseff was awarded the Westinghouse Order of Merit.

Since 1947, Professor Mouromtseff has been teaching in the department of physics at Upsala College, in East Orange, N. J. He was made a Fellow of the IRE in 1947, and has served on many IRE committees, including the Electronics Committee and Papers Review Committee. He is a member of the American Physical Society, the AIEE, and the New Jersey and National Society of Professional Engineers. Since 1943 he has been an active member of the Basic Science Division of the AIEE.

⁘

Joseph M. Pettit (S'39–A'40–M'45–SM'46) was born on July 15, 1916, at Rochester, Minn. He received the B.S. degree in 1938 from the University of California, the degree of Engineer in 1940 from Stanford University, and the Ph.D. degree in 1942 from the same institution. From 1938 to 1940, Dr. Pettit was a teaching and research assistant in electrical engineering at Stanford University. During the period 1940 to 1942 he was an instructor in electrical engineering at the University of California, and at the same time was a research associate in electrical engineering at Stanford University, in charge of a radio direction-finder project under the National Defense Research Committee. From 1942 through 1945, Dr. Pettit was engaged in radar countermeasures work on the staff of the Radio Research Laboratory, which operated under Harvard University in co-operation with the Office of Scientific Research and Development. He served progressively as receiver development engineer, group leader, and assistant executive engineer. During 1944, Dr. Pettit served in India and China as a technical observer with the Twentieth Air Force, and in 1945 he was in England as associate technical director of ABL-15, a laboratory associated with the Radio Research Laboratory. In 1945, he became supervising engineer at Airborne Instruments Laboratory, Inc Mineola, L. I., N. Y., in charge of receive development. Since January, 1947, he ha been at Stanford University, where he is a associate professor of electrical engineering


Joseph M. Pettit

Dr. Pettit is a member of the America Institute of Electrical Engineers, Sigma Xi Tau Beta Pi, and Eta Kappa Nu.

⁘

Andrew Tait was born in Bridgeport Conn., on April 22, 1920. Upon graduatio from high school, he went with the Rockefeller Institute fo Medical Research New York, N. Y., as a technical assistan in the physical chemistry laboratories There he was engaged in the design of specialized electronic equipment. He attended evening sessions at Brooklyn Polytechnic Institute, and was also engaged in experimental work in television receivers and preamplifiers for the "Fringe" operation.


Andrew Tait

In 1942 Mr. Tait joined the National Bureau of Standards staff, where he worked in the field of high-frequency direction finders. In 1944 he went on active duty as an officer in the Electronics Engineering Division of the U. S. Coast Guard, concerned with specialized military radio direction-finding equipment and associated radio communication equipment in this country and overseas.

In 1946 Mr. Tait rejoined the staff of the Central Radio Propagation Laboratory at the National Bureau of Standards, and is at present working in the field of navigation systems and the associated field of low-frequency propagation.

⁘

An Wang (A'45) was born in Shanghai, China, on February 7, 1920. He received the B.S. degree in electrical engineering from Chiao-tung University, China, in 1940. He attended Harvard University from 1945 to 1948, receiving the M.S. degree in 1946 and the Ph.D. degree in 1948.


An Wang

From 1940 to 1941, Dr. Wang was an assistant of Chiao-tung University. From 1941 to 1945, he was an engineer of the Central Radio Corporation in China. Since 1948, Dr. Wang has been a research fellow at Harvard University, working on the development of basic components of digital calculating machines. He is a member of Sigma Xi.

Institute News and Radio Notes

N.W.V. Hayes, President of Australian IRE, Is Dead

Norman William Victor Hayes, President of the Institution of Radio Engineers, Australia, died recently. Mr. Hayes was among the members of the Institution who were largely responsible for the close co-operation and the friendly spirit which exists between the Australian Institution and The Institute of Radio Engineers. His passing will be mourned by members of both organizations.


N. W. V. Hayes

Mr. Hayes commenced his engineering career in the Australian Postmaster General's Department as a Cadet Engineer. He advanced successively to the positions of Engineer, Divisional Engineer, Sydney and Central Administration, Superintending Engineer for Victoria, and Deputy Chief Engineer, Central Administration. At the time of his death he was Acting Assistant Director-General (Engineering Services), Engineer-in-Chief for Australia. He was also President of the Victorian Postal Institute which covers the welfare of all officers of that State.

As President of the Institution of Radio Engineers, Australia, he had continued his co-operative efforts for the development and expansion of the Institution, and had displayed the greatest zeal and enthusiasm during his seventeen years of membership. He was elected a Full Member in 1933, and at the invitation of the Council became a Fellow in 1940. Mr. Hayes had served on many committees. He was Chairman of the Melbourne Division for the period of the war.

He had served as Vice-President of the Institution from 1941 to 1943, as Deputy-President from 1948–1949, and as President from 1949 to 1950.

In commenting on his passing, representatives of the Institution paid tribute to Mr. Hayes as a "genial personality, an efficient engineer, and a wise administrator."

Technical Committee Notes

The Institute approved the appointment of the following as Chairmen of the IRE Technical Committees for the year May 1, 1950–April 30, 1951; Annual Review, Ralph Batcher; **Antennas and Waveguides,** A. Gardiner Fox; **Audio Techniques,** R. A. Miller; **Circuits,** W. N. Tuttle; **Electroacoustics,** E. S. Seeley; **Electron Tubes and Solid-State Devices,** L. S. Nergaard; **Electronic Computers,** Jay W. Forrester; **Facsimile,** J. V. L. Hogan; **Industrial Electronics,** D. E. Watts; **Measurements and Instrumentation,** Ernst Weber; **Mobile Communications,** F. T. Budelman; **Modulation Systems,** Bertram Trevor; **Navigation Aids,** Peter C. Sandretto; **Piezoelectric Crystals,** K. S. Van Dyke; **Radio Transmitters,** H. Tanck; **Receivers,** Richard F. Shea; **Sound Recording and Reproducing,** H. E. Roys; **Standards,** John G. Brainerd; **Symbols,** A. F. Pomeroy; **Television Systems,** Axel G. Jensen; **Video Techniques,** J. E. Keister; and **Wave Propagation,** H. G. Booker. Technical Committee Chairmen become members of the Standards Committee. This action facilitates direct representation of their Committees when prepared material comes before the Standards Committee for approval. In cases where a Technical Committee Chairman cannot attend a regular meeting of the Standards Committee, he will request his Vice-Chairman to attend in his place. The Institute wishes to extend its appreciation to the past Chairmen of Technical Committees for their excellent co-operation and diligent work during the past year. . . . **The IRE Technical Committee on Electron Tubes and Solid-State Devices** will sponsor, with AIEE, a Joint Conference on Electron Devices to be held at the University of Michigan, Ann Arbor, June 22 and 23. G. D. O'Neill has been appointed a member of the Committee on **Measurements and Instrumentation** as a representative of the Electron Tubes and Solid-State Devices Committee. . . . The Committee on **Antennas and Waveguides** held a meeting on March 21, under the Chairmanship of L. C. Van Atta. . . . **A meeting of the Receivers Committee** was held on March 10, under the Chairmanship of R. F. Shea. Jack Avins, Chairman of Subcommittee 17.2, reported that his group is working on the revision of Standards on Television: Methods of Testing Television Receivers, 1948, to incorporate single sideband receivers. A report on the activities of Subcommittee 17.3 was given by Chairman Jarvis. Mr. Shea, Chairman of Subcommittee 17.4, stated that the need for standard methods of measurement of spurious radiation has been accelerated by FCC action. This group will endeavor to write these methods on all types of receivers, placing the emphasis on television and FM equipment. Mr. Hodges gave a report of his Subcommittee 17.5. The activities of Subcommittee 17.6 were given by Mr. Mountjoy, Chairman. Mr. Mountjoy reported that work on the formulation of definitions is proceeding on schedule. Mr. Spielman, Chairman of Subcommittee 17.7, was not present to report for his Subcommittee, but had submitted a report at the last meeting. . . . **The Circuits Committee** held a meeting on February 24, under the Chairmanship of W. N. Tuttle. This Committee is preparing a glossary of definitions of Network and Circuit Theory. The glossary will be cumulative in the sense that it will include old definitions within its field, in addition to new definitions prepared by the Committee. It will also be complete so that it will have a maximum of utility as a reference. Mr. Grossman reported for Dr. Dietzold on the activities of Subcommittee 4.2, Linear Lumped-Constant Passive Circuits. The activities of Subcommittee 4.3, Circuit Topology, was reported by Professor Foster. A report was given by Dr. Bennett on Subcommittee 4.4, Linear Varying-Parameter and Nonlinear Circuits. Mr. Linvill reported for Mr. Huggins on the activities of Subcommittee 4.5, Time-Domain Network Analysis. Professor Weber, Chairman of Subcommittee 4.7, Linear Active Circuits, reported that a list of additional terms in the field of feedback amplifiers have been completed. Professor Krauss, Chairman of Subcommittee 4.9, Fundamental Quantities, reported that his Subcommittee is reviewing terms. Due to the voluminous return of applications for membership on Professional Groups, it is not possible to acknowledge receipt of applications from Headquarters. The Chairman of each Professional Group will communicate with each applicant as he is accepted. . . . L. C. Van Atta, Chairman of the IRE Professional Group on Antennas and Propagation, attended the San Diego Section Meeting on April 4 and spoke on "The Role of Professional Groups in the IRE." . . . The Annual Meeting of the **Audio Professional Group** was held on March 9, during the IRE National Convention at the Hotel Commodore. The Audio Group roster is in excess of 950 members. This Group sponsored a Symposium during the IRE Convention which proved highly successful. The attendance at the morning session of the Symposium of technical audio papers was over 450, while more than 300 were present at the afternoon session. **At the Annual Meeting an Administrative Committee** for the coming year was elected as follows: L. L. Beranek, Chairman; B. B. Bauer, Vice-Chairman; J. A. Green; O. L. Green; O. L. Angevine; J. K. Hilliard; and K. C. Morrical. It was agreed that "audio" would be treated in a professional sense rather than an avocational sense. There was a good deal of feeling that since the professional audio engineer is faced with very practical problems, this professional treatment would still be of great interest to the serious audio hobbyist. The first issue of the *Audio Professional Group Newsletter* has been distributed and will be circulated bimonthly. The newsletter will solicit papers so that at least one audio paper per month will appear in the Proceedings of the I.R.E., starting as soon as possible. The IRE Audio Professional Group plans to sponsor technical sessions at the fall meeting of the IRE/RMA in Syracuse, N. Y., and at the National Electronics Conference in September. **The Administrative Committee of the Professional Group on Broadcast Transmission Systems** held a meeting at IRE Headquarters on March 31. The new Administrative Committee for the coming year was elected as follows: Lewis Winner, Acting Chairman;

Orrin Towner, Earl Johnson, O. B. Hanson, R. M. Pierce, Les Bowman, W. B. Lodge, Frank Marx, H. A. Dorschug, Willard Hauser, George P. Hixenbaugh, Roland Hale, Roger Hodgkins, Arthur Stuarts, and Scott Helt. . . . **The Professional Group on Instrumentation** held its first meeting at the National Convention on March 9. Ernst Weber was appointed Acting Chairman of the Group and H. L. Byerlay, acting Vice-Chairman. The other members of this Group's Administrative Committee are Peter Janis, N. D. Saigeon, A. R. Satullo, S. C. Lawrence, E. P. Felch, E. H. Gamble, and Ivan Easton. . . . **The JTAC held a meeting on March 28,** under the Chairmanship of D. G. Fink. A report of the Committee of Consultants on Adjacent Channel Television Interference was prepared by the Chairman of the Committee on Consultants, W. T. Wintringham. This report is Volume V of JTAC, and has been presented to the Federal Communications Commission. . . . **The Planning Committee for the 1950 Joint IRE/AIEE Nucleonics Symposium** held a meeting on March 3, under the Chairmanship of G. W. Dunlap. This Symposium will be held in New York City on October 23, 24, and 25. Complete details will be announced later.

ASTE Establishes Fund For First Research Foundation

Establishment of a research foundation to carry on basic production research has been authorized by the Board of Directors of the American Society of Tool Engineers. An initial fund of $25,000 was appropriated at the same time.

Plans call for the use of existing production research facilities at various universities and colleges. The foundation will act, also, as an intermediary to assist small industries or companies interested in basic research but not in a position to finance such research.

The ASTE Board of Directors also adopted a resolution to work with the Army Ordnance Association on national preparedness plans.

Webster Is New Chairman of Research, Development Unit

William Webster, executive vice-president of the New England Electric System, Boston, Mass., since 1942, was sworn in as Chairman of the Research and Development Board by Secretary of Defense Louis Johnson on March 15. His first association with the Board was upon its inception in 1946 as the Joint Research and Development Board, when he served as consultant.

From September, 1948, to September, 1949, he was chairman of the Board's Committee on Atomic Energy, deputy to the Secretary of Defense on Atomic Energy Matters, and chairman of the Military Liaison Committee to the Atomic Energy Commission. Previously, he had been consultant to the Atomic Energy Commission from 1947 to 1948. Since October, 1949, Mr. Webster has been a consultant to the Office of the Secretary of Defense and the Atomic Energy Commission.

The Research and Development Board is composed of a civilian chairman and two representatives from each of the three military departments. Present Board members, in addition to Mr. Webster, are: Department of the Army: Archibald S. Alexander, Assistant Secretary, and General Mark W. Clark, Chief, Army Field Forces; Department of the Navy: Dan A. Kimball, Under Secretary, and Rear Admiral R. P. Briscoe, Director, Fleet Operational Readiness, Office, Chief of Naval Operations; and the Department of the Air Force: Harold C. Stuart, Assistant Secretary, and Lieutenant General Benjamin W. Chidlaw, Commanding General, Air Materiel Command.

NAB Surveying Manufacturers for Information on FM Sets

Edward L. Sellers, who is in charge of the NAB FM-Radio Division, is making a survey of all FM set manufacturers, asking for information on the characteristics of their FM receivers.

The NAB resolution expresses the belief that "a number of FM receivers now being offered the public do not operate satisfactorily" within a field intensity range of 50 microvolts per meter and that "some FM models are subject to excessive drift and are difficult to keep tuned."

He pointed out "that manufacture of FM home receivers with high sensitivity, effective limiting, and adequate antenna would go a long way in giving the country's some 700 FM broadcasters a new viewpoint on the role set manufacturers can play in providing the public with the full benefits of FM broadcasting on a nation-wide scale."

Curve Generator Developed at National Bureau of Standards

An instrument which gives an instantaneous visual display of electron tube characteristics has been developed at the National Bureau of Standards. The work of Milton L. Kuder, the curve generator plots directly on the screen of a cathode-ray oscilloscope the family of plate current versus plate voltage curves for any receiving tube. A standard rectangle is displayed along with the characteristic curves to provide a direct scale for voltage and current readings. In cases where the tube characteristics are not known or where an unusual combination of supply voltages is to be used, the curve generator can provide the necessary tube data at a great saving in time and labor.

In addition to producing plate-characteristic curves, the new instrument can provide a visual representation of plate current versus grid voltage. In this case the oscilloscope display is particularly convenient, since grid voltage increments are directly defined by calibrated vertical bars appearing on the oscilloscope screen; a standard current reference is given by a horizontal bar. All of the possible displays are produced by the curve generator without overloading the tube under test. Over-all accuracy of voltage and current readings from the oscilloscope screen is within plus or minus five per cent.

Attention!

Authors of Audio Papers

The Audio Professional Group of the IRE is planning a session at the National Electronics Conference to be held in Chicago, Ill., on September 25 to 27, and also at the Radio Fall Meeting to be held in Syracuse, N.Y., on October 30, 31, and November 1. In addition to these meetings, there will be a session at the IRE National Convention next year.

In order to avoid the possibility of omitting papers submitted too late for presentation at such meetings, all members who are interested in preparing audio papers should write to Dr. Leo L. Beranek, Chairman, IRE Audio Professional Group, Massachusetts Institute of Technology, Cambridge 39, Mass. Prompt notification to Dr. Beranek will assure authors that their papers can be considered for inclusion in the program of the proper meeting.

International Standardization Council Formed by Government

The co-ordination of matters involving national and international standardization in which Federal agencies are interested is now being accomplished through a newly established Interdepartmental Standards Council, composed of representatives from fifteen Federal agencies, including the FCC, Defense Department, and others.

One purpose of the Council, which meets monthly, is to provide machinery for the development of policy on national and international standardization matters of commercial significance as they concern the U. S. Government. The Council plans to assist recognized groups and technical organizations within industry concerned with standardization matters by providing a medium through which they may obtain the co-ordinated viewpoint of various Government agencies. S. P. Kaidanovsky, of the Federal Supply Service, is Chairman.

Calendar of COMING EVENTS

IRE–AIEE Conference on Electron Devices, University of Michigan, Ann Arbor, Mich., June 22–23

Conference on Ionospheric Physics, Pennsylvania State College, Pa., July 24, 25 and 26

IRE West Coast Convention of 1950, Municipal Auditorium, Long Beach, Calif., September 13–15

National Electronics Conference, **Edgewater Beach Hotel, Chicago, Ill., September 25–27**

IRE–AIEE Conference on Electronic Instrumentation in Nucleonics and Medicine, Hotel Sheraton, New York, N.Y., October 23–25

Radio Fall Meeting, Syracuse, N.Y., October 30, 31, November 1

American Radio Relay League Publishes World Radio Map

The American Radio Relay League of West Hartford, Conn., has published a new and completely revised edition of the Amateur Radio Map of the World. The four-color map, a special projection by Rand McNally, has been prepared for use by amateur radio stations primarily in their "DX" or long-distance international communications activities, and is completely different from any other map now on the market.

Published at a price of $2.00, the map measures 30 by 40 inches in size, and clearly shows the various countries of the world, together with the call sign prefixes used by the radio amateurs of those countries. Prefixes shown are not only those assigned by international agreement at Atlantic City, but also those used by the various military occupation forces throughout the world.

This map is a modified equidistant azimuthal projection, centered on Wichita, Kan., allowing distance measurements of reasonable accuracy between points in North America and the rest of the world. The map may also be used for determining great circle bearings from most points within the U.S.A.

In addition to the country boundaries, the map also shows time zones, principal cities, and International Amateur Radio Union continental subdivisions. Around the map's border are indexed the countries of the world according to the latest ARRL official countries list for amateur achievement awards, there being some 270 countries in the list. The border of the map also contains an explanation of the world's time zones and instructions on how to measure distance and bearings.

Radio Propagation Station Established in Virginia

A new radio propagation field station which will make continuous measurements of radio waves reflected from the upper atmosphere has been established at Fort Belvoir, Va., by the National Bureau of Standards. Made possible by the co-operation of the Army's Corps of Engineers and Signal Corps, the Belvoir Field Station is one of a system of fourteen stations operating under the supervision of the Bureau's Central Radio Propagation Laboratory as part of a world-wide network of over 50 radio observatories.

The station, 13 miles south of Washington, consists of four separate buildings designed for ionospheric and geophysical measurements. Equipment at the new field station includes the latest in field intensity recorders, ionospheric recorders, and visually-recording magnetographs. The new installation is continuing, on an improved basis, work begun by the National Bureau of Standards more than twenty years ago.

Data gathered at the new station will be used in the preparation of predictions three months in advance of the best frequencies for short-wave radio communication, as well as warnings of sudden radio disturbances.

Dr. Lester Field Is Honored By Eta Kappa Nu Association

Lester M. Field (S'39–M'48) was among the outstanding young electrical engineers who were honored by the Eta Kappa Nu Association in New York, N. Y., during the week of the Winter General Meeting of the American Institute of Electrical Engineers, January 30–February 3. Awards were made to Robert Chase Cheek, first prize winner, and to Dr. Field and Louis G. Gizendanner, who received honorable mention.

The awards were established in 1936 by Eta Kappa Nu, national honorary electrical engineering fraternity, to recognize young electrical engineers for "meritorious service in the interest of their fellow men."

To qualify for the award, the candidate must be not older than 35 years, nor be out of college for more than 10 years by May 1 of the year for which he is cited. After qualifying on these two counts, the candidate is judged on the basis of accomplishment in professional, social, and cultural fields.

Industrial Engineering Notes[1]

Television

The **Philco Corp. is working on a direct-view color television tube,** David B. Smith, vice-president, Engineering and Research, told the FCC, as he urged the Commission to give this development "careful consideration in the determination of color standards." With respect to a single-gun direct-view tube, the witness testified Philco was not in a position "at this time to report on this work other than to say that we have considerable confidence that single-gun all-electronic color tubes are not only possible but hold the greatest promise for future color receivers." . . . **Columbia Broadcasting System** has highlighted its testimony before the FCC on color television by revealing a new development which it claims doubles the resolution of pictures delivered by the CBS color TV system. The new developnent, Peter C. Goldmark said, involves addition of the horizontal interlace principle to the CBS field sequential system, making it possible to obtain greater definition in the horizontal direction. Dr. Goldmark said CBS has been working on the new principle for the past four months, but success was achieved only two days before he testified. Although it is purely a laboratory arrangement, he said he was confident that it established that the CBS pictures which the FCC and the public have already seen "by no means represent the maximum potential of the CBS system." . . . **Color Television, Inc.,** which has been troubled by technical difficulties in demonstrations of its color TV system before the FCC, apparently corrected this fault for showings staged for the press and important government officials. The company **will invite the FCC to take another look at the system in the near future. . . . Television receiver production** in February rose to the highest weekly average to date, RMA tabulations show, although the TV set output during the four-week period, January 30 to February 24, of 367,065 units was under that of last November, which included five weeks and production of 414,223 television receivers by RMA member companies. Radio receiver production also continued at a high level with the result that total radio and TV set output of 1,117,458 units reported to RMA was the highest for any four-week month since 1948. . . . A strong **hint of possible FCC controls in the manufacture of television receivers** was thrown out by Chairman Wayne Coy on April 5 during the cross examination of Donald G. Fink, Chairman of JTAC, at the FCC color inquiry. Mr. Fink also testified as an individual under subpoena so that the Commission could obtain his personal opinions on many question. Several times during the lengthy hearings Commission members and the Chairman, in particular, indicated they felt some government controls over the manufacture of receivers may be desirable and in the public interest. . . . **An official demonstration,** for the record, **of the RCA tri-color television tube** included the transmission of color pictures under conditions paralleling the coaxial cable, thus overcoming one of the hitherto criticisms of the RCA color system. The demonstration was made for the FCC, parties to the current TV hearing, and the press. Elmer W. Engstrom, Vice-President in Charge of Research, RCA Laboratories, explained that the New York-to-Washington coaxial cable was not available for the demonstration, but he explained that conditions had been created which corresponded to the coaxial cable in which the TV color broadcasts were reduced in width from 4 to 2.7 megacycles. . . . **The Bureau of Labor Statistics, summarizing a study of the radio and television industry, reported that "television,** heard about but not seen until after the war, **already has replaced radio as the chief product** of radio and television manufacturers." In 1949 television sales were responsible for about two-thirds of the industry's total dollar receipts, according to a study published by the U. S. Department of Labor's Bureau of Labor Statistics. The title of the study is "Radios and Television Sets," and is a part of BLS Detailed Report on Employment and Pay Rolls. Contrary to popular notion, the radio and television industry has provided relatively few additional jobs as a result of the television boom. Employment in January was only 4 per cent higher than in January, 1949, an approximate increase of 5,000 in the production worker force. Accelerated production of television sets was accomplished almost completely by intra-plant transfers of workers previously employed on radio set production. . . . **The Navy Department has demonstrated** some of the results of **a four-year research project in developing equipment and techniques for making motion picture recordings of black-and-white and color television broadcasts.** The demonstration included the first movies of a TV broadcast from under water and

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[1] The data on which these NOTES are based were selected, by permission, from *Industry Reports*, issues of March 17, March 24, March 31, and April 7, published by the Radio Manufacturers Association, whose helpful attitude is gladly acknowledged.

movies made of a color television receiver in operation. The Navy Bureau of Ships, which is investigating uses of television for all three armed services, is currently engaged in improving the use of television for aerial reconnaissance, salvage operations, research, engineering, pilotless aircraft, and training.

Radio and Television News Abroad

Foreign radio manufacturers established three assembly plants in Brazil in 1949, and two other plants are scheduled to be established in 1950, according to information received by the U.S. Department of Commerce. The assembly plants were producing between 1,000 and 2,000 sets per month each at the end of 1949. An estimated 150,000 receiving sets were assembled in 1949, compared with 70,000 and 55,000 sets during 1948 and 1947, respectively. . . . **Radio apparatus is in production by 21 firms in Norway,** and all but one make home-type radio receivers, according to information received by the U. S. Department of Commerce. . . . **Imports of radio receivers into Norway amounted to 2,012 units** from January to November, 1949. More than 1,200 of those sets were imported from the Netherlands. Approximately two million radio sets were in use in Argentina at the beginning of this year. Sixty per cent of the sets are designed to receive short-wave broadcasts. . . . **There were an estimated 148,000 radio sets in use in Greece** on December 31, 1949, compared with 43,000 at the time of liberation in 1944. . . . **Radio receivers of United States manufacture continue to be highly regarded in Egypt** and prices of sets are comparable to those of European suppliers, according to information received by the U. S. Department of Commerce. However, imports from the United States are limited because of the shortage of dollars. The Egyptian Import Permit Office did not issue licenses against dollar payment for the importation of radio receivers from the United States during 1949, and this policy is expected to be maintained during 1950. A limited number of receivers will continue to be imported from the United States by payment in dollars secured from foreign sources at the free market rate of exchange, the Department said. Radio receivers are not produced in Egypt. Imports during January to November, 1949, totaled 55,410 sets. Principal sources of supply were: The United Kingdom, 27,178 sets; The Netherlands, 19,148; and the United States, 4,817. During 1948 imports aggregated 47,930 units, of which 29,540 were from the United Kingdom, 12,545 from the Netherlands and 3,644 from the United States. . . . **An estimated 1,321,600 radio sets were in use in Denmark** in January, 1950, of which 1,310,900 were privately owned. The number of listeners per set was an estimated 3.5 persons. Almost all of the radios are designed to receive short-wave broadcasts. . . . At the end of January **there were 11,907,832 radios and 280,092 television receivers in operation in the British Isles,** according to information received by the U. S. Department of Commerce. . . . **The U. S. Embassy in the Netherlands reports** that there were **an estimated 200,000 radio sets manufactured** for the domestic market of that country in 1949. Imports of radio sets and parts are reported to consist chiefly of the products of foreign subsidiaries of the Philips Company. . . . **There were 180 concerns engaged in the manufacture of radio sets, tubes, and parts in Australia during 1949,** according to a report received by the U. S. Department of Commerce. Production during the year is estimated at 280,000 sets, or about seven per cent less than the 300,000 receivers reported in 1948. . . . **Production of radio receivers in Argentina** during 1949 is estimated at 80 per cent of the 1948 production due to a shortage of inventories of raw materials in the second half of the year.

FCC Actions

President Truman has submitted to Congress 21 reorganization plans based on the Hoover Commission reports, including one involving the Federal Communications Commission. He told Congress that when these plans become effective, half the proposals suggested by the Commission will have been acted on. One of the submitted plans would give the FCC Chairman additional power and make the Chairman responsible for the administration of that agency. . . . **The FCC has announced the appointment of several key officials,** including a new Chief Engineer, in connection with its functional reorganization. Curtis B. Plummer, present Chief of the Television Broadcast Division, Bureau of Engineering, was named Chief Engineer in charge of the new Office of Chief Engineer. Other appointments and their new titles follow: Benedict P. Cottone, General Counsel in charge of the new Office of General Counsel; William J. Norfleet, Chief Accountant in charge of the new Office of Chief Accountant; J. Fred Johnson, Jr., Chief Hearing Examiner in charge of the Hearing Division; William K. Holl, Executive Officer in charge of the Office of Administration; and Harold J. Cohen, Chief of the New Common Carrier Bureau. . . . The **FCC has denied an application of the Zenith Radio Corp. to change transmitter location, transmitter and antenna systems,** and to increase operating power output of its experimental TV broadcast station from 1-kw (visual and aural) to 5-kw visual and 2.5-kw aural, in connection with Zenith's Phonevision tests which had been authorized by the Commission.

Statistics

A total of 14,500,000 radio and television sets were purchased in 1949, it has been announced jointly by RMA and the National Association of Broadcasters. This first joint annual study of industry statistics was made under the direction of Kenneth H. Baker, director of research for the National Association of Broadcasters, and Frank W. Mansfield, chairman of the RMA Industry Statistics Committee. Home radio set sales account for 7,956,000 of the total number, and automobile radio sets for the 3,964,000 that went into 78 per cent of all cars manufactured last year. In addition to automobile sets, the total number of radio sets in the hands of the public at the end of 1949 was 70,436,000, of which 5,000,000 were in places of public assembly and 65,436,000 in homes. . . . **The total number of television sets in use at the end of 1949 was 3,764,000.** All told, there were in use by the end of last year 88,964,000 radio and television sets. . . . **Sales of both wholesale and retail radio dealers in January showed an increase over the corresponding month of 1949, but fell substantially below December,** the Department of Commerce has reported. January sales of appliance and specialties wholesalers, including radios, increased 23 per cent over January a year ago but dropped 33 per cent below December sales, the Department stated.

Signal Corps Develops Device for Millimicrosecond Pulses

Radio communications are expected to be advanced by the development of an electronic device capable of generating pulses of energy lasting only 1,000th of 1,000,000th of a second, according to an announcement by the U. S. Army Signal Corps. Circuits based on the millimicrosecond pulse technique are being embodied by the Signal Corps in experimental models of radio communications equipment, from short-range portable and vehicular sets, up through the trunk microwave radio relay stations. The development is said to be valuable as a tool in laboratory work. It will bear on interference reduction.

National Bureau of Standards Finds 'Round-the-World Signals

Very low-frequency radio signals traveling completely around the world have been detected by the National Bureau of Standards after a normal delay time of more than a tenth of a second. The signals, transmitted from the Naval Radio Station NSS at Annapolis, Md., on a frequency of 18 kilocycles with a power of 350 kilowatts, were received at the National Bureau of Standards radio propagation field station at Sterling, Va., about 50 miles away.

Recording Microwave Instrument

An instrument which measures and records small differences in frequencies between resonant cavities has been developed by the National Bureau of Standards. Full details of the new device are to be published by the Bureau in *The Technical News Bulletin* for April, monthly publication of NBS.

Virgil Graham Announces Time of Next Three Radio Fall Meetings

Virgil Graham, Associate Director of the RMA Engineering Department, has announced that the next three Radio Fall meetings, successors to the Rochester Fall Meetings, will be held as follows: 1950, Hotel Syracuse, Syracuse, N. Y., October 30, 31, November 1; 1951, King Edward Hotel, Toronto, Canada, October 29, 30, 31; 1952, Hotel Syracuse, Syracuse, N. Y., October 27, 28, 29.

IRE Awards, 1950

Medal of Honor

Frederick E. Terman

"For his many contributions to the radio and electronics industry as teacher, author, scientist and administrator."

Morris Liebmann Memorial Prize

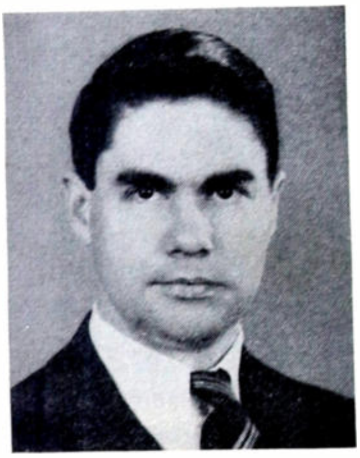
Otto H. Schade

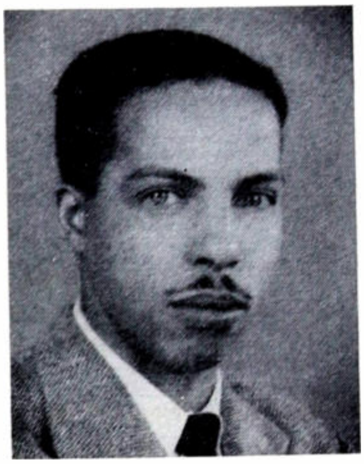
"For his outstanding contributions to the analysis, measurement technique, and system development in the field of television and related optics."

Harry Diamond Memorial Award

Andrew V. Haeff

"For his contribution to the study of the interaction of electrons and radiation, and for his contribution to the storage tube art."





Browder J. Thompson Memorial Award

Joseph F. Hull

Arthur W. Randals

"For their paper in the November, 1948, issue of the Proceedings of the I.R.E. entitled 'High-Power Interdigital Magnetrons.'"

Editor's Award

E. J. Barlow

"For his unusually clear presentation of a technical subject in his paper, 'Doppler Radar,' published in the April, 1949, issue of the Proceedings of the I.R.E."

Fellow Awards

Ralph R. Batcher

"For his pioneer work with cathode-ray instruments, and more recently for his development of precision variable frequency standards and meters."

Arthur L. Albert

"For his contribution to electronics as a teacher and writer."

Alda V. Bedford

"For his many contributions to sound recording and the development of many circuits of basic importance to present-day television."

Rawson Bennett

"For his contributions to programming, guiding and developing sonar systems for military use, and his contribution to the administration of military electronics laboratories."

Frank J. Bingley

"In recognition of his contributions in the field of television broadcast engineering."

K. H. Blomberg

"In recognition of his many contributions to development and engineering in the field of communications in Sweden."

John F. Byrne

"For his development of a system of polyphase broadcasting and for effective engineering administration in connection with radar countermeasures during the war."

William G. Dow

"For outstanding contributions to the teaching and understanding of electronics through the organization of educational material and the stimulation of students and others to critical thought."

Dudley E. Foster

"For his contributions and technical direction of work leading to better radio receiver design."

George W. Gilman

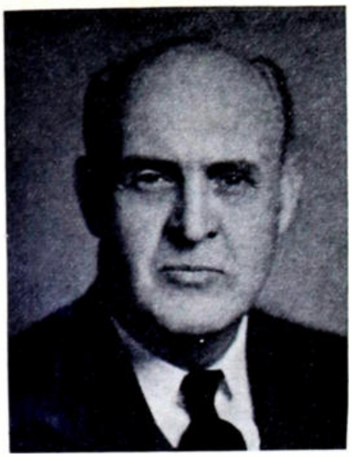
"For his contributions to the communication art and for his direction of important developments in the field of radio transmission systems."

George L. Haller

"For his work on aircraft antennas and for his diversified radio effort during the war."

Albert G. Hill

"For his work in the utilization of electronics to research in physics and his contribution in the conversion of wartime development laboratories to peacetime fundamental research."



Frederick S. Howes

"For his contributions as a teacher in the field of communication engineering."

Harley A. Iams

"For the development of electronic apparatus for converting images formed by electromagnetic waves to electrical signals, first in the television field, and later, using new principles, in the realm of short radio waves."

Willis Jackson

"For his service as an educator and his many contributions to the literature in both the radio and electrical fields."

Rudolph Kompfner

"For his research in electron tube theory and particularly for his original contributions to the concepts of the traveling-wave amplifier."

Harry B. Marvin

"For his outstanding contributions to the measurements art and pioneering work in FM, television, and allied fields."

Pierre Mertz

"In recognition of his important contributions to the fundamental concepts of television and reception."

Garrard Mountjoy

"For his contributions to the design of radio and television broadcast receivers."

John H. Miller

"For his long activity and many contributions in the field of electrical metering and measuring techniques."

Emanuel R. Piore

"For his many contributions in the fields of engineering and physical sciences, and for outstanding service in enhancing the national effort in basic research."

Jack R. Poppele

"For his long and continued leadership in the broadcasting field, and in particular for his recent contributions to television broadcasting."

Simon Ramo

"For his many contributions to the analysis of electromagnetic phenomena and for his leadership in research."

Claude E. Shannon

"For his contributions to the philosophy of new pulse methods and to the basic theory of communications."

W. Arthur Steel

"For his contributions in Canada in advancing development of military radio, broadcasting, and international communication."

Jerome R. Steen

"For his work in the introduction and development of statistical quality control techniques in electron tube manufacturing."

George R. Town

"For his contributions in radio receiver engineering and research."

Dayton Ulrey

"For pioneering research and for administrative and technical contributions to the development of special purpose and power tubes."

Robert R. Warnecke

"For his engineering and research contributions to vacuum-tube theory and design in France."

Harold A. Zahl

"For his guidance of the Army Signal Corps research program in the transition from war to peace and for his contribution to radar in its early development stages."

Report of the Secretary—1949

TABLE I.—TOTAL MEMBERSHIP DISTRIBUTION BY GRADES

Grade	As of Dec. 31, 1949 Number	As of Dec. 31, 1949 % of Total	As of Dec. 31, 1948 Number	As of Dec. 31, 1948 % of Total	As of Dec. 31, 1947 Number	As of Dec. 31, 1947 % of Total
Fellow	284	1.1	259	1.1	239	1.1
Senior Member	2,421	9.0	2,192	9.4	2,068	9.8
Member	3,559	13.3	3,334	14.2	3,017	14.4
Associate	12,309***	46.0	11,713**	50.0	12,079*	57.4
Student	8,196	30.6	5,939	25.3	3,634	17.3
Totals	26,769		23,437		21,037	

* Includes 1,490 Voting Associates.
** Includes 1,260 Voting Associates.
*** Includes 1,159 Voting Associates.

TABLE II.—FIVE-YEAR ANALYSIS OF U. S. AND FOREIGN MEMBERSHIP

	1949	1948	1947	1946	1945
TOTAL	26,769	23,437	21,037	18,154	15,779
U. S. and Possessions	24,434	21,048	18,723	15,898	14,053
Foreign (including Canada)	2,335	2,389	2,314	2,256	1,726
Per Cent Foreign	8.72	10.2	11.0	12.4	10.9

TO THE BOARD OF DIRECTORS,
THE INSTITUTE OF RADIO ENGINEERS

Gentlemen:

An account of the activities of the Institute for the year 1949 is submitted herewith, constituting the usual Report of the Secretary as required in the Bylaws.

Membership increased 14 per cent, a trend now in its sixteenth year. The large increase in Student Branches during 1948 is reflected in a 38 per cent increase of Student Members in 1949. Five additional Sections have been organized, making a total of 53, and the development of Subsections brings the number of these to 10.

The Professional Group plan has grown from three Groups at the end of 1948 to eight at the end of this past year, and others are in the process of organization. This advance in activity indicates much interest in this method of meeting the need of members having common technical interests.

Technical Committees have increased 20 per cent to a total of 24, and the record shows a significant increase in the number of meetings they have held.

The 1949 National Convention was attended by 15,710 persons, an increase of 8.7 per cent over the year before. Activities were greater, the number of technical papers being up 34 per cent to 172 covered by 34 technical sessions, and exhibits up 25 per cent to 225. Other meetings were: The Canadian Regional Conference, the National Electronics Conference, the New England Radio Engineering Meeting, the West Coast IRE Convention, the Rochester Fall Meeting, the Southwestern IRE Conference, and the URSI Meeting at Washington.

The Joint Technical Advisory Committee, created in June, 1948, by the Institute and the Radio Manufacturers Association, has been very active, and in addition to giving oral testimony at hearings before the Federal Communications Commission, has issued volumes three and four of its *Proceedings*, bringing the total number of pages containing technical information and data to 485 for the four volumes.

The Regional Director Plan functioned successfully during the year, confirming the usefulness of this method of broadening the geographical distribution of members of the Board of Directors. There were five Board meetings in 1949 and attendance by Regional Directors scored 92.5 per cent.

The accompanying statement indicates the condition of the Institute's fiscal affairs. By dint of very careful budgeting, the net income for the year has improved, but as the reserves are considered to be inadequate, and the sums going to them should be still further increased during these years of high economic activity.

Attention is also called to the growing cost of serving Student Members, the revenue from Student dues being insufficient to cover the basic costs of providing them with the PROCEEDINGS. At the end of 1949 such revenue deficiency on an annual basis was at the rate of approximately $15,000.

The year 1950 opens with the fiscal and business management of Institute affairs well attuned to the enlarged scope of its size and activities, a staff organization having sufficient experience with this broadened scope to more adequately cope with its problems, and having its governing groups and policy committees set up for better and more efficient coverage of the tasks to be performed.

Respectfully submitted,

Haraden Pratt

HARADEN PRATT, *Secretary*

February 28, 1950

Membership

At the end of the year 1949, the membership of the Institute, including all grades, was 26,769, an increase of 3,332, or 14 per cent over the previous year. The number of elections is the largest in any one year in the history of the Institute. The 3,332 member increase in 1949 compares favorably with 2,883 and 2,400, the increases for 1947 and 1948, respectively. The percentage increase was 15 per cent in 1947, and 11 per cent in 1948. The membership trend from 1912 to date is shown graphically in Fig. 2.

The distribution of members in the various grades for the years 1947, 1948, and 1949 is shown in the accompanying plot, Fig. 3. Actual figures for 1947, 1948, and 1949 are shown in Table I. Of the 11,150 nonvoting Associates, 3,443 have been in that grade for more than five years. The membership ratio of Associates to higher grades was 6 to 1 in 1944, 4 to 1 in 1945, less than 3 to 1 in 1946, and about 2 to 1 in 1947, 1948, and 1949, a very satisfactory trend. Monetary exchange difficulties and devaluation of the pound have contributed to the reduced increase in foreign membership during the past three years, shown in Table II.

It is with deep regret that this office records the death of the following members of the Institute during the year 1949:

Fellows

Edwin H. Colpitts (A'14–F'26)
William W. Hansen (A'39–F'47)
Frank B. Jewett (F'20)
Hendrik J. Van der Bijl (M'17–F'28)

Senior Members

Robert D. Avery (J'29–A'33–SM'44)
John H. Barron, Jr. (M'29–SM'43)
William A. Beatty (M'41–SM'43)
Arnold E. Bowen (SM'46)
Montague Ferry (SM'46)
William C. Hahn (A'36–SM'45)
Landon C. Herndon (M'36–SM'43)
Joseph W. Milnor (A'16–M'26–SM'43)
Newell R. Smith (M'41–SM'43)
Frank E. Spaulding, Jr. (A'35–VA'39–SM'48)
Andrew G. Tynan (M'43–SM'43)

Members

Clinton B. DeSoto (M'46)
Ralph E. Hartzsch (M'48)
Edwin R. Love (M'46)
Albert D. Martin, Jr. (M'45)
Frank E. Sessler (M'47)

Voting Associates

Frederick W. Townsley (A'32–VA'39)

Associates

Vincent A. Dolan (S'47–A-49)
Ambrose C. Kibler (A'44)
Wilson A. Maisel (A'46)
Cyril E. Maitland (A'47)
Martin Prager (A'46)
George Tompkins (A'45)

Students

Adney D. Collins (S'47)
William H. Wilder (S'46)

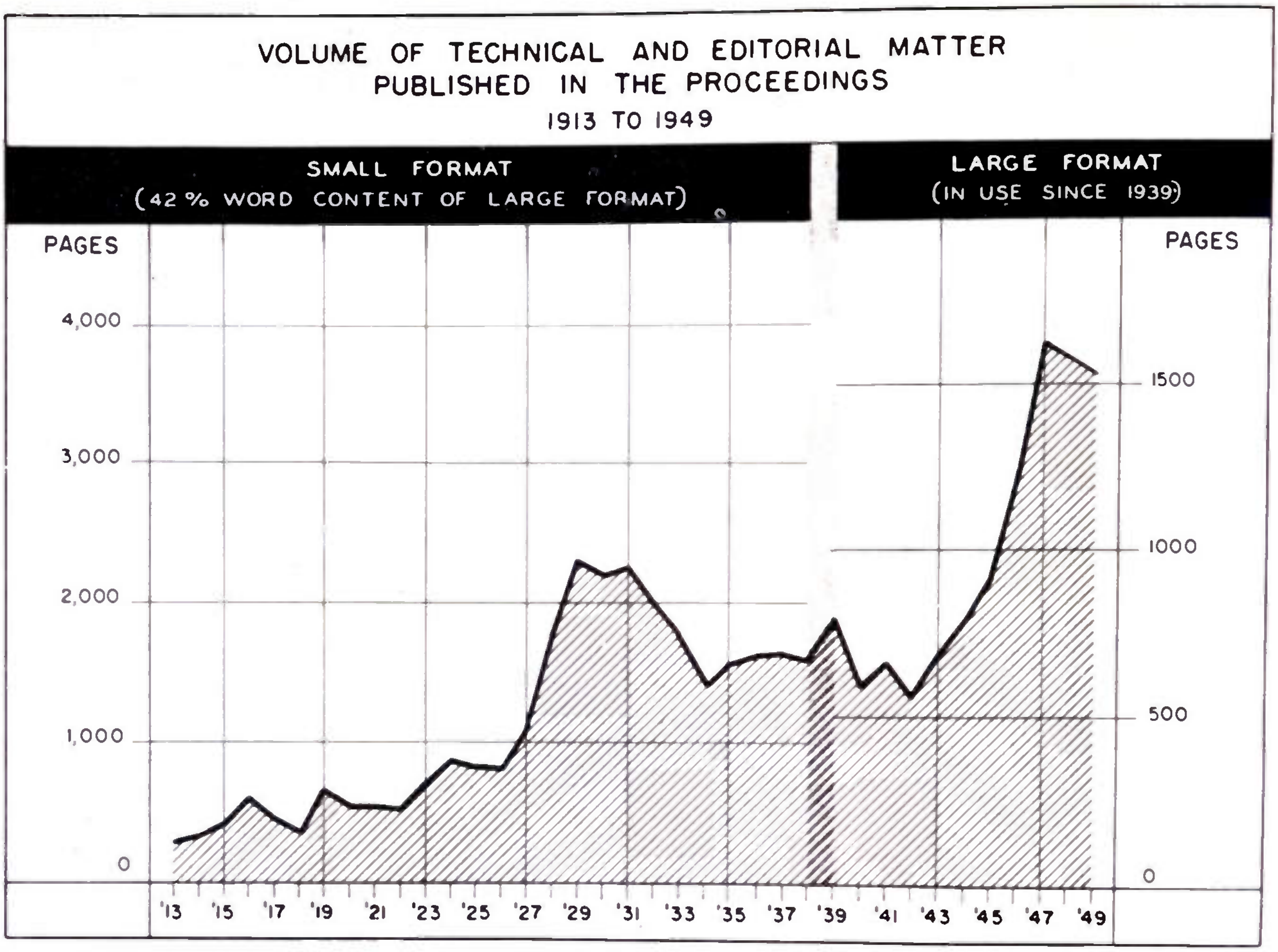

VOLUME OF TECHNICAL AND EDITORIAL MATTER PUBLISHED IN THE PROCEEDINGS
1913 TO 1949
SMALL FORMAT
(42 % WORD CONTENT OF LARGE FORMAT)
LARGE FORMAT
(IN USE SINCE 1939)
PAGES
4,000
3,000
2,000
1,000
0
PAGES
1500
1000
500
0
'13
'15
'17
'19
'21
'23
'25
'27
'29
'31
'33
'35
'37
'39
'41
'43
'45
'47
'49

Fig. 1

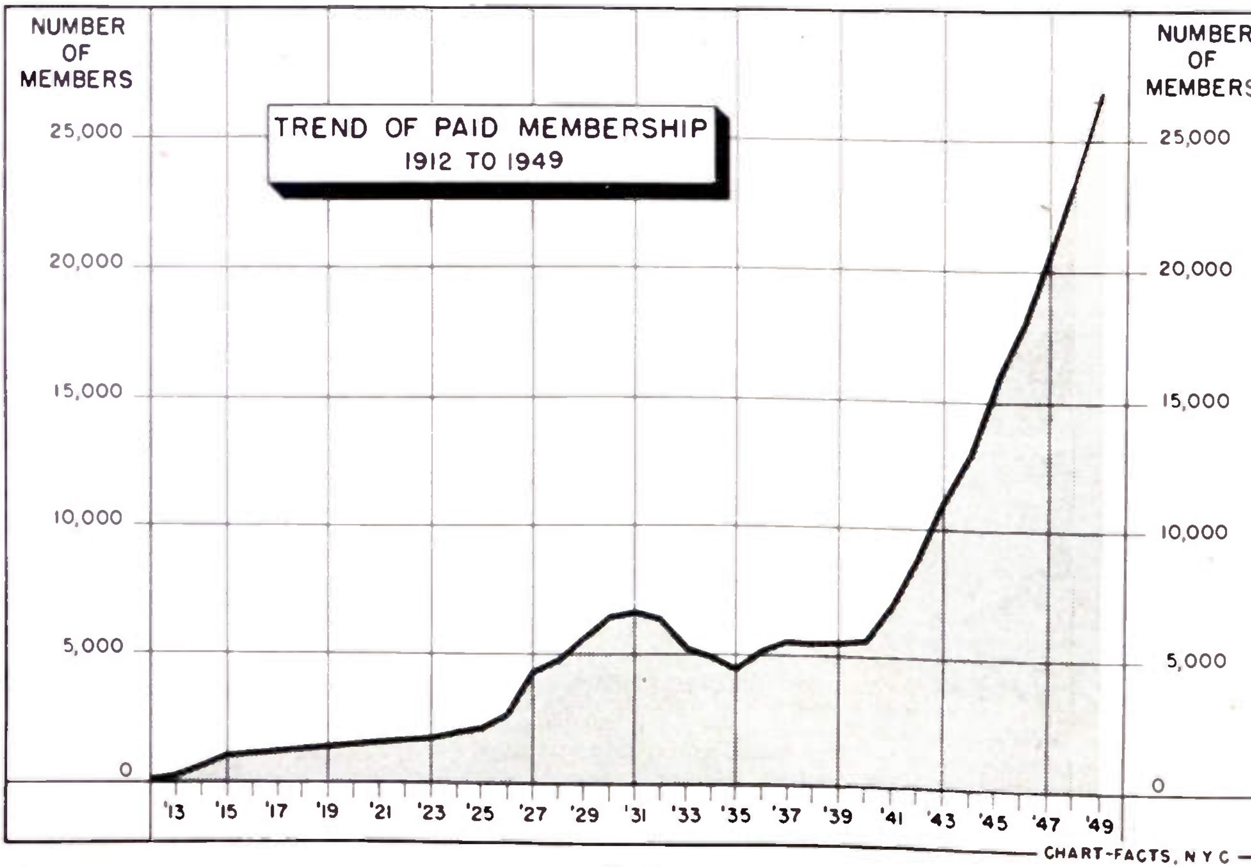

NUMBER OF MEMBERS
TREND OF PAID MEMBERSHIP
1912 TO 1949
25,000
20,000
15,000
10,000
5,000
0
NUMBER OF MEMBERS
25,000
20,000
15,000
10,000
5,000
0
'13
'15
'17
'19
'21
'23
'25
'27
'29
'31
'33
'35
'37
'39
'41
'43
'45
'47
'49
CHART-FACTS, N Y C

Fig. 2

Fiscal

A condensed summary of income and expenses for 1949 is shown in Table III, and a balance sheet for 1949 is shown in Table IV. Income and expenses for each year since 1914 are shown graphically in Fig. 4.

TABLE III—SUMMARY OF INCOME AND EXPENSES, 1949

Income		
Advertising	$173,366.61	
Member Dues and Conventions	393,420.75	
Subscriptions	42,435.90	
Sales Items, Binders Emblems, etc.	15,752.52	
Investments Income	9,881.50	
Miscellaneous Income	263.04	
TOTAL INCOME		$ 635,120.32
Expenses		
PROCEEDINGS Editorial Pages	$143,944.06	
Advertising Pages	98,042.74	
Yearbook	39,137.25	
Section and Student Branch Rebates	22,449.80	
Sales Items	11,301.34	
Miscellaneous Printing	5,481.31	
General Operations	160,321.64	
Convention Cost	102,792.62	
TOTAL EXPENSES		$ 583,470.76
SURPLUS		$ 51,649.56
Reserve for Depreciation		6,631.21
NET SURPLUS		$ 45,018.35

TABLE IV—BALANCE SHEET—DECEMBER 31, 1949

Assets		
Cash and Accounts Receivable	$265,742.75	
Inventory	13,260.11	
Total Current Assets		$ 279,002.86
Investments at Cost	372,655.16	
Building and Land at Cost	356,687.30	
Furniture and Fixtures at Cost	70,004.27	
Other Assets	18,029.70	
Total		817,376.43
TOTAL ASSETS		$1,096,379.29
Liabilities and Surplus		
Accounts Payable	$ 5,223.36	
Federal Taxes on Emblems, etc.	66.42	
Total Current Liabilities		$ 5,289.78
*Deferred Liabilities		239,248.58
Total Liabilities		$ 244,538.36
Reserve for Depreciation		20,827.52
Surplus—Donated	596,804.93	
Surplus—Earned	234,208.48	
Total Surplus		831,013.41
TOTAL LIABILITIES AND SURPLUS		$1,096,379.29

* 1950 Items, PROCEEDINGS for members and subscribers, Advertising, and Convention Service.

Editorial Department

The year 1949 was marked by several major changes in editorial policies and procedures which affected the format of the PROCEEDINGS OF THE I.R.E., the publication of Standards, the production of the Yearbook, the procurement of papers, and the size of the backlog of papers available for publication. Each of these changes was designed to increase the effectiveness with which the publications of the Institute might serve the membership.

The most pressing problem faced by the Editorial Department at the beginning of 1949 was the excessive size of the backlog, comprising 141 papers or 981 PROCEEDINGS pages. Of this amount, 69 papers representing 442 pages had been accepted, the remainder being under review. By the end of the year the backlog was reduced to 103 papers totaling 585 pages, of which 53 papers or 265 pages had been accepted, representing a normal backlog status for the first time since the end of 1945. This substantial decrease should result in eliminating excessive delays in the publication of papers.

During 1949 there were published in the PROCEEDINGS a total of 2,404 pages, including covers. Of these, 1,532 were editorial pages and 872 advertising pages. Of the editorial pages, 1,099 were devoted to technical papers (including discussions and correspondence), 32 to Standards, 163 to Abstracts and References, and 238 to nontechnical material (including covers, contents pages, and annual index). Of the advertising pages, 747⅛ represented paid advertising, while the remaining contained useful editorial material in the form of Institute Membership and Section Meeting lists, and news of new products and the industry in general.

The total of 2,404 pages published during the year compares with 2,452 in 1948, and 2,576 in 1947. The number of editorial pages (1,532) compares with 1,592 in 1948, and 1,636 in 1947. Advertising pages numbered 757⅓ in 1948 and 940 in 1947. The number of editorial pages published each year since 1913 is shown in Fig. 1.

Technical papers totaling 183 were published in 1949, as against 173 in 1948. Authorship of these papers was by 252 individuals of whom 183, or 73 per cent, were members of the Institute. In 1948, 176 of the 241 authors, or 73 per cent, were members.

The volume of papers submitted for publication continues at a high rate. During 1948, 289 papers totaling an estimated 2,043 PROCEEDINGS pages were submitted, or an average of 24 papers and 170 pages per month. During 1949, 287 papers of 1,894 pages were received, or 24 papers and 158 pages per month.

A major innovation in Institute publication policy was inaugurated with the appearance of four Standards in the December, 1949, issue of the PROCEEDINGS OF THE I.R.E. Because of high printing costs and limited circulation, it was not found feasible to continue publishing each Standard as a separate report. The publication of Standards in the PROCEEDINGS is regarded as a valuable new service to the membership as it makes available to all members each Standard that is published, thereby ensuring the widest practicable distribution of Standards without additional cost to the members. The four Standards published in 1949 are listed under *Technical Activities*.

In a move to consolidate the technical material in the PROCEEDINGS, it was decided to discontinue the Waves and Electrons Section beginning with the January, 1950, issue. This decision will not affect the contents of the journal, but will offer a more closely integrated arrangement of technical and editorial material.

With the reduction in backlog described above, it was found possible to embark on an active program of papers procurement. In order to co-ordinate procurement activities, the work of the Papers Procurement Committee was transferred to the Editorial Department. In carrying out these duties, the Editorial Department will be guided in part by the Technical Committees and Professional Groups of the Institute.

In an effort to reduce publication costs, the 1949 Yearbook was published by the varitype-offset method. This marked a return to the policy of publishing the Yearbook on an annual basis, a practice which was discontinued in 1932.

The Editorial Department was deeply saddened by the sudden and untimely passing on April 27 of Clinton B. DeSoto, Technical Editor, who for three years had effectively applied his unusual abilities to the efficient functioning of the Editorial Department staff. He was succeeded by E. K. Gannett, former Assistant Secretary of the Institute. The work of the Editorial Department was carried on without interruption under the able direction of Editor Alfred N. Goldsmith and Executive Secretary George W. Bailey, and with the effective assistance of Miss Mary L. Potter, Assistant Editor, and a competent staff. The Editorial Department is directed by the Editor in matters of editorial policy, content, and format, and by the Executive Secretary in matters of administration, both functioning through the Technical Editor. It has been greatly aided by the helpful advice and assistance unstintingly given by the members of the Board of Editors, the members of the Papers Review Committee, the Papers Procurement Committee, and the Editorial Administrative Committee.

Technical Activities

Technical Committees. During 1949, twenty-four technical committees with their subcommittees and task groups, held a total of 163 meetings at Institute Headquarters, eleven meetings at the Hotel Commodore during the National Convention, and ten meetings in other localities, a total of 184, a substantial increase over 1948.

The Board of Directors approved the formation of the Audio Techniques, Video Techniques, and the Sound Recording and Reproducing Committees. These three new committees assumed the expanded functions of the original Audio and Video Techniques Committee. Approval was also secured for the formation of the Committee on Measurements and Instrumentation.

The Board authorized the change in the name of the Railroad and Vehicular Communications Committee to the Committee on Mobile Communications, to correctly indicate this committee's activities. The name of the Radio Receivers Committee was also changed to the Receivers Committee. The Committee on Antennas was renamed the Committee on Antennas and Waveguides. The names of these committees were altered to more appropriately describe the fields in which the committees are working.

The *Standards on Radio Receivers: Methods of Testing Amplitude-Modulation Broadcast Receivers, 1948*, prepared by the Receivers Committee was published in January, 1949.

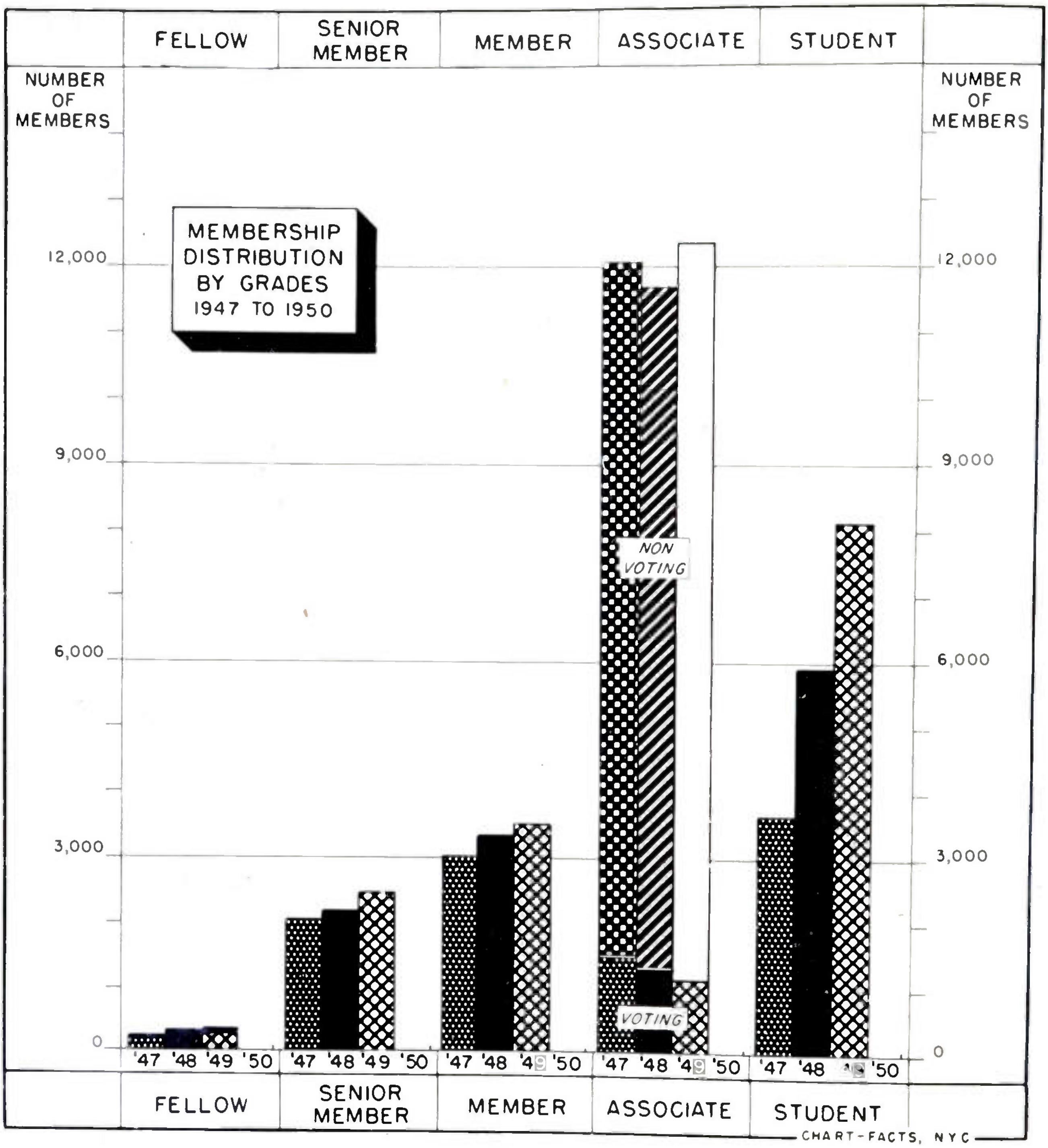


Fig. 3

The policy of publishing Standards in the PROCEEDINGS was adopted in late 1949. Issues of the PROCEEDINGS containing Standards are identified by a red band on the cover and spine.

The following Standards prepared by the IRE technical committees were published in the December, 1949, issue of the PROCEEDINGS:

Standards on Radio Aids to Navigation: Definition of Terms, 1949.
Standards on Railroad and Vehicular Communications: Methods of Testing, 1949.
Standards on Piezoelectric Crystals, 1949.
Tests for Effects of Mistuning and for Downward Modulation, 1949.

This was a supplement to *Standards on Radio Receivers: Methods of Testing Frequency-Modulation Broadcast Receivers, 1947.*

The following Standards prepared by the Video Techniques Committee have been approved and will be published in the PROCEEDINGS during 1950, namely: *Methods of Measurements of Television Signal Levels; Measurements of Timing of Video Switching Systems;* and *Measurements of Resolution in Television.* Approval was also secured for the *Standard Designations for Electrical, Electronic and Mechanical Parts, 1949,* and several sections of the *Standards on Electron Tubes: Methods of Testing.*

The Annual Review Committee prepared a survey, Radio Progress During 1949, which appeared in the April, 1950, issue of the PROCEEDINGS OF THE I.R.E.

The Electron Tubes and Solid-State Devices Committee sponsored its Annual Conference on Electron Devices at Princeton University. The attendance was over 300. The Conference was an outstanding contribution to electronic engineering.

The Definitions Co-ordinating Subcommittee of the Standards Committee, which correlates definitions of all terms under consideration within the technical committee structure of the IRE, has completed a Master-Index of all terms defined, or in the process of definition by IRE committees. The Index is available to all IRE committees, and upon request to other professional societies. A list of the terms and references contained in the Master Index was published in June. Pres-

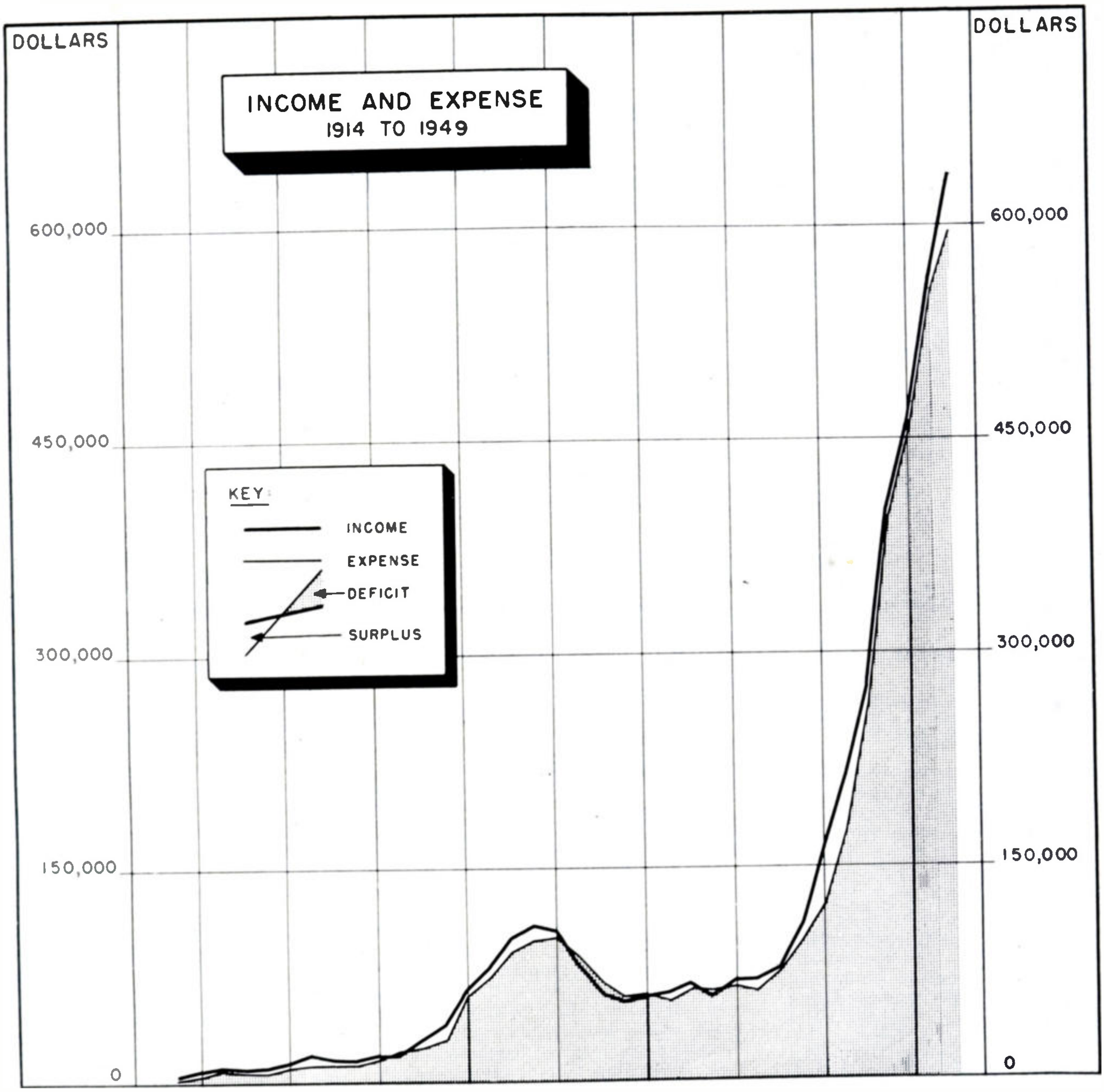


Fig. 4

ent plans call for a revised edition in 1950.

Professional Group System. There are eight Professional Groups presently in existence in the following fields: Antennas and Propagation, Audio, Broadcast and Television Receivers, Broadcast Transmission Systems, Circuit Theory, Nuclear Science, Quality Control, and Vehicular and Railroad Radio Communications. During 1949 two of these Groups co-sponsored symposia national in scope. The Nuclear Science Professional Group, jointly with AIEE, co-sponsored a three-day Conference on Electronic Instrumentation in Nucleonics and Medicine. Eight hundred and forty-seven registered for the Conference and 21 papers were presented. A round-table discussion was well attended and proved to be a most popular feature of the program.

The Antennas and Propagation Professional Group co-sponsored a meeting in cooperation with URSI in Washington. There was an average of approximately 340 people attending the sessions. Three other groups sponsored sessions at the 1949 RMA/IRE Fall Meeting in Syracuse. These were the Professional Groups on Quality Control, Broadcast and Television Receivers, and Audio. The Nuclear Science Group, since its inception, has issued a *Newsletter* at regular intervals. Groups are now in the process of being formed in the fields of Instrumentation, Telemetering, Basic Sciences, Electron Tubes, and Navigation Aids.

Joint Technical Advisory Committee. The JTAC held seven meetings at IRE Headquarters, one meeting in Baltimore, Md., and one during the March, 1949, National Convention, a total of nine meetings. JTAC participated in the RCA Television Observer Tests at Princeton University. JTAC also prepared two additional volumes. The first contained the "Official Correspondence between the Federal Communications Commission and JTAC," the second, "Comments on the Proposed Allocation of Television Broadcast Services," for presentation to the FCC.

The office of the IRE Technical Secretary prepared and compiled the RMA Color Television Report presented to the Federal Communications Commission in September, 1949. Material for inclusion in the JTAC Report to the FCC was obtained from this Report.

Technical activities were carried on under the guidance of the Standards Co-ordinator W. R. G. Baker, Technical Secretary L. G. Cumming, and a competent staff.

The Institute owes a debt of gratitude to those members who have given so unstintingly of their time and energy to the Technical Committees and Professional Groups activities during 1949.

(Continued on page 702)

Section Activities

We were glad to welcome five new Sections into the Institute during the past year. They are as follows:

Denver	(March) 1949
Ft. Wayne	(March) 1949
Inyokern	(March) 1949
Akron	(September) 1949
Schenectady	(September) 1949

The total number of Sections is now 53. There has been a substantial increase in membership of these Sections with a few exceptions. It should be borne in mind that most Sections with noticeable decreases in membership released substantial numbers of members to new Sections.

Student Branches

The Institute's program with respect to Student Branches continued to flourish in 1949. The number of Student Branches formed during 1949 was 29, 17 of which operate as joint IRE-AIEE Branches. The total number of Student Branches is now 88, 43 of which operate as joint IRE-AIEE Branches. This increase of Student interest was accompanied by a large increase in Student members.

Following is a list of Student Branches formed during 1949:

*Agricultural and Mechanical College of Texas, University of Colorado, *Cornell University, *University of Delaware, Fenn College, Georgia School of Technology, University of Kentucky, *Lafayette College, University of Louisville, University of Maine, *Marquette University, Mississippi State College, *University of Missouri, *University of Nebraska, *New York University (Evening Division), *Ohio State University, *University of Pennsylvania, *The Pennsylvania State College, University of Pittsburgh, *Rhode Island State College, San Diego State College, San Jose State College, Seattle University, *University of Southern California, *Southern Methodist University, *Tufts College, *Tulane University, Virginia Polytechnic Institute, and *Yale University.

* Joint IRE-AIEE Student Branches.

Books

Electronic Engineering Master Index, Edited by John F. Rider

Published (1950) by Electronics Research Publishing Co., Inc., 480 Canal St., New York 13, N. Y. 339 pages+10-page index+4-page bibliography+xiii pages. 6¾×9¾. $19.50.

This book is a comprehensive compilation of the literature in the field of electronics and related subjects. Covering the period 1947–1948, it contains 18,000 published paper titles and 5,500 electronic and allied subject patents. These references have been collected from 230 major scientific magazines from all over the world, as well as documents from government and university laboratories which have been declassified.

The reference and patents are arranged in terms of subject under headings such as Acoustics, Broadcasting, Instrumentation, Photoelectric Tubes, Radar, Television, etc. With an arrangement like this, it is possible to find quickly the references in a given field. On the other hand, since each reference appears only once in the listing, it is often necessary to look up several descriptive subject headings to exhaust the search. For instance, if the literature on the subject of Simultaneous Equation Solvers is desired, it is necessary to look up the subject of Computers, Analysers, and Mathematics. This is true especially in cases where the subject cannot be classified in any single item. To assist in this situation, a cumulative cross index of subjects is provided in this volume which covers not only the 1947–1948 edition, but also the 1925–1945 and 1946 editions. The list of patents is especially useful since this type of material is not normally available as classified in this book.

The Electronic Engineering Master Index is an invaluable reference book for anyone engaged in the research and development field, and is highly recommended by this reviewer.

John R. Ragazzini
Columbia University
New York 27, N. Y.

Giant Brains or Machines That Think, by Edmund C. Berkeley

Published (1949) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 208 pages+14-page index+xvi pages. 5½×8½. $4.00.

In "Giant Brains or Machines That Think," Edmund C. Berkeley has written a stimulating book on the various large-scale computers that have appeared since 1940. The book also mentions briefly the later machines, such as the Moore School's EDVAC, that were not completed at the time that the book was published.

This book is intended for everyone and, for that reason, suffers slightly from oversimplification. Each chapter can be read by itself, which makes the book very easy to read. The author suggests that the reader should skip what he finds uninteresting. There is a voluminous bibliography which is indexed by author and subject. The supplements on "Words and Ideas" and "Mathematics" are excellent.

Mr. Berkeley discusses the extent to which the action of the machine resembles thinking. No mechanical brain built so far can do intuitive thinking, make bright guesses and leaps to conclusions, determine all its own instructions, and perceive complex situations outside of itself and interpret them. He believes that these operations may eventually be performed by machines.

As an illustration of principles, he describes a very simple mechanical brain, including its control, method of transferring information, computing, and output. He next traces the history of punchcard machines.

The description of the MIT differential analyzer No. 2 is especially interesting because of the ingenious devices and artifices described, particularly the disk, wheel, and screw integrator originally devised by Lord Kelvin in 1879.

Harvard's IBM automatic sequence-controlled calculator, Moore School's ENIAC, and the Bell Laboratory's General-Purpose Relay Calculator are given a chapter each. The author describes their method of operation, speed, method of feeding data into the machine, how the information comes out of the machine, method of programming, and cost.

The Kalin-Burkhart Logical-Truth Calculator is especially challenging. Such machines may eventually make the decisions of civilization if the organization of mankind increases in complexity, as the welfare of billions of people seems to require.

Just as a major part of human thinking depends on memory, mechanical brains depend more and more on storing of information for rapid access.

Improved means are procured by magnetic wire and tape, by circulating information in a delay line, electrostatic storage tubes, etc. These devices allow the "program" itself to be stored, and make subroutines, such as the finding of square roots, available when needed.

The author describes the types of problems that may eventually be solved by machines. These are problems where large amounts of data must be correlated quickly; such as problems of industrial control and scheduling; weather information; psychological testing; and economics, such as "How will a rise in the price of steel affect the farming industry?"

The author discusses the responsibility of scientists for the devices they create. He discusses the type of social controls over robot machines that could be possible "if men were reasonable."

Mr. Berkeley is to be congratulated on having written a thought-provoking book that should be on every engineer's bookshelf.

Charles J. Hirsch
Hazeltine Electronics Corp.
58-25 Little Neck Parkway
Little Neck, L. I., N. Y.

Institute Committees—1950

Executive

R. F. Guy, *Chairman*
D. B. Sinclair, *Vice-Chairman*
Haraden Pratt, *Secretary*

S. L. Bailey
W. R. G. Baker
A. N. Goldsmith
J. W. McRae

Board of Editors

A. N. Goldsmith, *Chairman*

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J. S. Allen
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Ralph Bown
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O. H. Caldwell
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M. G. Crosby
R. B. Dome
W. G. Dow
E. W. Engstrom
W. L. Everitt
W. G. H. Finch
D. G. Fink
H. C. Forbes
I. A. Getting
P. C. Goldmark
A. W. Graf
F. W. Grover
L. B. Headrick
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J. A. Hutcheson
C. M. Jansky, Jr.
J. K. Johnson
L. F. Jones
H. S. Knowles
J. D. Kraus
J. B. H. Kuper
J. L. Lawson
D. G. Little
F. B. Llewellyn
S. S. Mackeown
Nathan Marchand
E. D. McArthur
Knox McIlwain
J. W. McRae
L. A. Meacham
G. F. Metcalf
E. L. Nelson
D. O. North
H. F. Olson
R. M. Page
H. O. Peterson
G. W. Pickard
Haraden Pratt
C. A. Priest
J. R. Ragazzini
Simon Ramo
H. J. Reich
J. D. Reid
F. X. Rettenmeyer
P. C. Sandretto
S. W. Seeley
V. W. Sherman
L. C. Smeby
C. E. Smith
J. A. Stratton
W. C. Tinus
K. S. Van Dyke
E. K. Van Tassel
E. C. Wente
H. A. Wheeler
J. R. Whinnery
W. C. White
L. E. Whittemore
Jerome Wiesner
G. W. Willard
I. G. Wolff
V. K. Zworykin

Awards

M. G. Crosby, *Chairman*

S. L. Bailey
D. E. Chambers
Keith Henney
F. S. Howes
A. V. Loughren
J. W. McRae
B. E. Shackelford
Karl Spangenberg

Education

H. J. Reich, *Chairman*

R. G. Anthes
R. E. Beam
L. J. Black
C. C. Chambers
R. M. Fano
G. H. Fett
R. A. Galbraith
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ACOUSTICS AND AUDIO FREQUENCIES

016:534 1047

References to Contemporary Papers on Acoustics—A. Taber Jones. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 66–70; January, 1950.) Continuation of 521 of April.

534:061.3 1048

Subject Matter of the Acoustics Conference Organized by the Acoustics Commission of the Academy of Sciences, U.S.S.R., Moscow, 26th February–1st March, 1949—(*Bull. Acad. Sci. URSS*, vol. 13, pp. 631–747; November and December, 1949. In Russian.) Report of 22 papers on various acoustical subjects presented at the conference. Many of the papers review Russian research in architectural acoustics, sound recording and reproduction, speech analysis, noise, etc.

534.131 1049

Acoustical Properties of Anisotropic Materials—W. J. Price and H. B. Huntington. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 32–37; January, 1950.) Investigation of ADP and KDP single crystals by pulsed ultrasonic methods. Results are analyzed.

534.231+538.566.2 1050

A New Method for Solving the Problem of the Field of a Point-Radiator in a Multilayer Nonuniform Medium—Brekhovskikh (*See* 1216.)

534.231 1051

The Sound Field near a Freely Oscillating Piston Membrane—J. Meixner and U. Fritze. (*Z. Angew. Phys.*, vol. 1, pp. 535–542; December, 1949.) The results of calculations for several wavelengths, of the order of magnitude of the membrane diameter, are presented graphically and compared with the sound field of an oscillating piston membrane in a rigid wall. The results are used to investigate the diffraction of a plane wave incident perpendicularly on a circular disk. For a wavelength about a third of the disk diameter, the sound fields of the incident and diffracted waves are determined, for both sides of the disk, and compared with approximations from Kirchhoff's diffraction theory. Comparison is also made between the sound field of a freely oscillating membrane and the lines of equal intensity behind a circular opening in a perfectly conducting plane on which a plane electromagnetic wave is incident perpendicularly.

534.26 1052

Variational Principles in Acoustic Diffraction Theory—H. Levine. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 48–55; January, 1950.) "The diffraction of a plane harmonic sound wave by an aperture in an infinitely thin rigid plane screen is investigated theoretically. Variational principles for the diffracted spherical wave amplitude at large distances from the aperture are derived. In the first of these, the stationary property is exhibited for the class of functions comprising the normal derivative of the aperture velocity potentials, whose distributions are governed by a generally insolvable integral equation. The second involves functions which characterize the discontinuity in velocity potentials at the screen (or their deviation from infinite screen distributions), and are specified by another integral equation. A comparison of the two variational principles is given, which indicates that their over-all agreement, following use of approximate functions, is a measure of the accuracy obtained. The plane wave transmission cross section of the aperture is related to the imaginary part of the diffracted amplitude in the direction of incidence and is cast in stationary forms. Particular attention is given to the low- and high-frequency behavior of the various forms of cross section, including comparison with Kirchhoff theory predictions."

534.26 1053

The Diffraction of Sound by Rigid Disks and Rigid Square Plates—F. M. Wiener. (*Jour. Acous. Soc. Amer.*, vol. 22, p. 47; January, 1950.) Correction to 3000 of 1949.

534.321.9:534.14 1054

Generation and Analysis of Ultrasonic Noise—S. C. Ghose. (*Nature* (London), vol. 165, pp. 66–67; January 14, 1950.) A brief description of apparatus used in the study of ultrasonic noise from turbo-jet engines. Special microphones and calibrating generators operating at frequencies up to 100 kc are described, and some details are given of a siren generator of ultrasonic noise of high intensity.

534.321.9:534.2 (26.03) 1055

Transmission of 24-kc/s Underwater Sound from a Deep Source—M. J. Sheehy. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 24–28; January, 1950.) Experimental methods are described. Preliminary results indicate that (a) the attenuation decreases with depth as both transmitter and hydrophone are lowered from 150 to 1,000 ft, (b) the average attenuation below 300 ft is about 6 db per 1,000 yd, (c) the magnitude of the amplitude fluctuations of the directly transmitted signals increases roughly as the square root of the range, from about 10 per cent at 100 yd to 45 per cent at 3,000 yd, and (d) surface-reflected signals show greater fluctuations than the direct signals.

534.321.9:534.373 1056

Ultrasonic Absorption in Liquids—G. W. Hazzard. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 29–32; January, 1950.) An experimental study of the effect of molecular symmetry on absorption.

534.373:534.6 1057

Attenuation in a Rectangular Slotted Tube of (1,0) Transverse Acoustic Waves—H. E. Hartig and R. F. Lambert. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 42–47; January, 1950.) Details are given of a pickup device for measuring the SWR of transverse waves. Results for the frequency range 2,400–4,200 cps are discussed.

534.6:621.395.6 1058

An Artificial Mouth for Acoustic Tests—P. Chavasse. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 436–438; January 30, 1950.) Description of an artificial head which when used with an artificial voice (3404 of 1947) produces sounds which can replace with advantage the human voice in many tests of communication equipment.

534.61+534.83 1059

The Objective Measurement of Noise: Its Possibilities and Limitations—P. Baron. (*Ann. Télécommun.*, vol. 4, pp. 330–340; October, 1949.) A critical examination of various methods of measurement of noise intensity and a discussion of the merits of proposed methods based on the spectral distribution of energy.

534.612.4 1060

A Method for the Absolute Calibration of Sound Receivers by the Reciprocity Method—M. V. Kazantseva. (*Compt. Rend. Acad. Sci. URSS*, vol. 58, pp. 1649–1651; December 11, 1947. In Russian.) Usually the absolute calibration of sound receivers is carried out in a free field and in chambers small in comparison with the sound wavelength. A method using a tube not shorter than half the sound wavelength is considered.

534.614:534.154 1061

A Crystal Pick-Up for Measuring Ultrasonic Wave Velocity and Dispersion in Solid Rods—A. E. Bakanowski and R. B. Lindsay. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 14–16; January, 1950.) A description of a pickup which has been successfully applied to determine the internodal distances in a solid metal rod vibrating in resonance with the applied oscillator frequency. Accuracy is similar to that obtainable by the powder method; the pickup can be applied to cases for which that method is unsuitable.

534.64 1062

Reciprocity Pressure Response Formula which includes the Effect of the Chamber Load on the Motion of the Transducer Diaphragms—M. S. Hawley. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 56–58; January, 1950.)

534.75+534.792 1063

The Masking of Pure Tones and of Speech by White Noise—J. E. Hawkins, Jr. and S. S. Stevens. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 6–13; January, 1950.) The results of a study of monaural masking at eight sensation levels in the range 20–90 db.

534.771 1064

The Watch, as an Apparatus for Measuring Acuteness of Hearing—P. Chavasse, R. Caussé, and R. Lehmann. (*Ann. Télécommun.*, vol. 4, pp. 413–424; December, 1949.) The ticking of a watch is submitted to spectral analysis and its sound intensity and the principal components of the spectrum are considered. Different methods of noise measurement are compared. A watch affords a ready means for rapid measurements of hearing acuity; the physiological deductions which can be made from such measurements are discussed.

534.771 1065

A Method of Calculating Hearing Loss for Speech from an Audiogram—H. Fletcher. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 1–5; January, 1950.)

534.83/.84:061.3 1066

Symposium on Noise, London, 14th–19th July 1948—P. Chavasse. (*Ann. Télécommun.*, vol. 4, pp. 377–382; November, 1949.) General comment on and review of the proceedings. Thirty-four of the papers presented are listed with authors' names. Twenty-six of these are summarized. They concern architectural acoustics; noise in aircraft, ships, and public halls; noise analysis; loudness measurement, etc. See also 2115 of 1949 (Beranek).

534.85:621.396.645 1067

Phonograph Reproduction: Part 1—McProud. (*See* 1117.)

621.395.61 1068

On the Application of Negative Feedback to Electroacoustic Systems—P. Chavasse and P. Poincelot. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 529–530; February 6, 1950.) The response curve of a reversible electrodynamic microphone was improved by the application of up to 32 db negative feedback.

621.395.623.7 1069

Duo-Cone Loud Speaker—H. F. Olson, J. Preston, and D. H. Cunningham. (*RCA Rev.*, vol. 10, pp. 490–503; December, 1949.) Full account of modifications of an earlier model which were noted in 17 of February.

621.396.645.37:621.395.623.7:621.3.018.8 1070

Output Impedance Control—T. Roddam. (*Wireless World*, vol. 56, pp. 48–49; February, 1950.) A method is described for varying the damping of a loudspeaker without alteration of the output level. Negative voltage feedback is used to keep the amplifier gain constant while the output impedance is varied by positive current feedback.

621.395.623.7+534.85 1071

Sound Reproduction [Book Review]—G. A. Briggs. Publishers: Wharfedale Wireless Works, Bradford, England and British Industries Corporation, 315 Broadway, New York 7, 1949, 143 pp., $2.95. (*Electronics*, vol. 23, pp. 236–237; February, 1950.) "The book consists of two major parts: Part I: Loudspeakers; Part II: Records. The part on loudspeakers extends the discussion of the previous book, Loudspeakers: The Why and How of Good Reproduction' (3333 of 1948) . . . The part on records surveys recording techniques and characteristics and discusses the various ills the art is heir to, such as tracking error, surface noise and motor rumble. The most interesting feature of the book from the reviewer's viewpoint is the series of photomicrographs, taken by C. E. Watts, of needles and grooves. The 200X pictures show the effects of wear after various numbers of playings with different types of needles. They provide an excellent objective argument for using as hard a material for the tip of a pickup needle as possible."

ANTENNAS AND TRANSMISSION LINES

621.315.212 1072

Theory of the Coaxial Cable: Part 1—F. Pollaczek. (*Jour. Phys. Radium*, vol. 8, pp. 215–224; July, 1947.) By means of the ideas and methods of classical mathematical physics, making use of Green's functions, integral equations, and developments in series, a rigorous theory of the coaxial cable in the periodic regime is established which is directly based on Maxwell's equations. A particular study is made of that part of the field which does not undergo exponential attenuation and which is completely neglected in the elementary theory of cables based on Kirchhoff's equations.

621.315.212 1073

Statistical Study of Irregularities in Coaxial Cables—A. Sarti. (*Alta Frequenza*, vol. 18, pp. 195–204; October, 1949.)

621.392.21+621.392.26† 1074

Impedance of Resonant Transmission Lines and Wave Guides—W. W. Harman. (*Jour. Appl. Phys.*, vol. 20, pp. 1252–1255; December, 1949.) Universal curves relating the Q value and the resonance impedance of capacitively terminated transmission-line and waveguide sections are given and discussed. Their use in the design of cavity resonators and for determining resonator shunt resistance is considered.

621.392.26† 1075

The Excitation of a Circular Waveguide by a Ring Aerial—A. Gaponov-Grekhov and M. Miller. (*Zh. Tekh. Fiz.*, vol. 19, pp. 1260–1270; November, 1949. In Russian.)

621.392.26† 1076

Waveguide Field Patterns in Evanescent Modes—A. L. Cullen (*Wireless Eng.*, vol. 26, pp. 317–322; October, 1949.) Field patterns in a waveguide at frequencies below cutoff are calculated, both for a single evanescent field, and for two oppositely attentuated ("incident" and "reflected") fields. Isometric projection of the surface depicting the electric field in the guide is found to provide a useful picture of the field distribution.

621.392.26† 1077

Application of Picard's Method of Successive Approximations to the Study of Discontinuities in Waveguides—T. Kahan. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 527–529; February 6, 1950.) A general solution of the problem of the echo pattern in a waveguide having a slight nonuniformity. The solution is given in the form of an infinite series converging more rapidly the smaller the nonuniformity.

621.392.26†:538.566 1078

On the Theory of the Propagation of Electromagnetic Waves in Tubes with Abruptly Varying Cross-Sections—G. V. Kisun'ko. (*Compt. Rend. Acad. Sci.* (*URSS*), vol. 58, pp. 1653–1656; December 11, 1947. In Russian.) A mathematical investigation of the propagation of electromagnetic waves, without reflection, through joints between two waveguides or through a combination of elements such as diaphragms, transformers, and other similar devices.

621.392.26†:621.315.611 1079

An Investigation of Dielectric Rod as Wave Guide—C. H. Chandler. (*Jour. Appl. Phys.*, vol. 20, pp. 1188–1192; December, 1949.) An experimental investigation indicates that the guiding effect is retained even when the rod is only a fraction of a wavelength in diameter. The greater part of the guided energy is then outside the dielectric, so that a very-low-loss waveguide results. Measurements at a wavelength of 1.25 cm indicate attenuations down to 0.004 db/m in polystyrene rod and show good agreement with theoretical predictions. A resonator using a dielectric-rod waveguide is described; this has a maximum Q of about 53,000.

621.392.26†:621.315.611 1080

Attenuation in a Dielectric Circular Rod—W. M. Elsasser. (*Jour. Appl. Phys.*, vol. 20, pp. 1193–1196; December, 1949.) Analysis is given of the nonradiating modes propagated along a cylindrical dielectric rod, with discussion of dielectric loss. The attenuation for the three lowest modes in a rod of polystyrene is computed; the results agree with the values found experimentally by Chandler (1079 above).

621.392.5:621.392.26† 1081

Microwave Directional Couplers—S. Freedman. (*Radio and Telev. News, Radio-Electronic Eng. Supplement*, vol. 14, pp. 11–13. . . 25; February, 1950.) An illustrated description of various types of waveguide coupling and their use in microwave measurements.

621.396.67 1082

The Concept of an Angular Spectrum of Plane Waves, and Its Relation to that of Polar Diagram and Aperture Distribution—H. G. Booker and P. C. Clemmow. (*Proc. IEE* (London), Part III, vol. 97, pp. 11–17; January, 1950.) "A critical examination is made of the somewhat loose and incomplete statement that a polar diagram is the Fourier transform of an aperture distribution. By aperture distribution it is necessary to understand, in the two-dimensional case, distribution across the aperture of the component along the aperture plane of the electromagnetic field in the plane of propagation. Furthermore, the concept of the polar diagram has to be replaced by that of an angular spectrum, except in the common case when the aperture may be considered more or less limited in width, and the field is being evaluated at a point whose distance from the aperture is large compared with the width of the aperture (and the wavelength). For example, it is convenient for some purposes to regard the problem of diffraction of a plane wave by a emi-infinite plane screen, with a straight edge, s a problem about an aperture distribution in he plane of the screen. This is a case for which he concept of a polar diagram is not, in general, pplicable, and has to be replaced by that of n angular spectrum. The field at all points in ront of a plane aperture of any distribution may be regarded as arising from an aggregate of plane waves traveling in various directions. The amplitude and phase of the waves, as a unction of their direction of travel, constitutes an angular spectrum, and this angular spectrum, appropriately expressed, is, without approximation, the Fourier transform of the aperture distribution. If the aperture distribution is of such a nature that the concept of the polar diagram is applicable at sufficiently great distances, then the polar diagram is equal to the angular spectrum. But the angular spectrum is a concept that is always applicable, whereas the polar diagram is one that is liable to be invalid for example, in the Sommerfeld theory of propagation over a plane, imperfectly reflecting earth)."

621.396.67 1083

Constructing Helical Antennas—E. D. Smith. (*Electronics*, vol. 23, pp. 72–75; February, 1950.) Construction details are given for several helical antennas suitable for the 435-Mc amateur band and for citizens' radio on 465 Mc. A simple impedance-matching unit for coaxial-cable feeder is also described.

621.396.67:530.1 1084

Electromagnetic Similitude: General Principles and Application to the Study of the Aerial—S. Colombo. (*Ann. Télécommun.*, vol. 4, pp. 370–376; November, 1949.)

621.396.67:621.397.62 1085

Trends in TV Receiver Antenna Design—I. Kamen. (*TV Eng.*, vol. 1, pp. 8–11; January, 1950.) A critical analysis, with illustrations, of eight types of antenna particularly suitable for vision and sound reception in the U.S.A.

621.396.671 1086

Gain of Aerial Systems—D. A. Bell. (*Wireless Eng.*, vol. 26, p. 380; November, 1949.) Correction to 3057 of 1949.

621.396.671 1087

Impedance/Frequency Characteristics of Some Slot Aerials—J. W. Crompton. (*Proc. IEE* (London), Part III, vol. 97, pp. 39–44; January, 1950.) Measurements were made by the earth-plane method, using a coaxial standing-wave indicator, of the input impedance near half-wave resonance of (a) simple center-fed slot antennas of various widths, (b) two dumb-bell slot antennas, and (c) an end-fire slot antenna, over the frequency range 130–370 Mc.

621.396.677 1088

The Design of Metallic Delay Dielectrics.—J. Brown. (*Proc. IEE* (London), Part III, vol. 97, pp. 45–48; January, 1950.) "A theory of metallic delay dielectrics, which is more accurate than existing ones and is based on an analogy with shunt-loaded transmission lines, has been developed for the simplest case, when the delay medium consists of an array of infinitely long conducting strips. Formulas have been obtained for the refractive index, impedance and cut-off wavelength of such a medium, and the more important results are presented graphically. A procedure for designing such dielectrics to have a desired refractive index is outlined, and some limitations in the use of metallic delay lenses are pointed out."

621.396.67 1089

Radio Aerials [Book Review]—E. B. Moullin. Publishers: Oxford University Press, London, 514 pp., 50s. (*Wireless World*, vol. 56, p. 70; February, 1950.) "The first section of the book is theoretical. The Lorentz vector and scalar retarded potential functions are first established and then applied to specific problems. These include the fields due to filaments, the effect of flat-sheet and V-shaped reflectors, and problems relating to cylinders immersed in electromagnetic fields.

Some hypothetical problems are solved rigorously, usually in terms of Bessel functions, and practical problems are considered as approximations—usually very close ones—to the hypothetical cases. The power gain of typical arrays with various current distributions is calculated, and methods of suppressing the side-lobes discussed. A short section is devoted to the isolated antenna.

The remaining one-third of the book describes experimental procedure, and the results of measurements made on some of the antennas described in the first section. Results for V antennas are given in great detail, and include the radiation patterns for various V angles and sizes of sheet."

Attention is drawn to many practical design considerations, such as the permissible tolerance on reflector shape, the use of netting and rods instead of continuous sheet, and similar problems. "The treatment is mainly mathematical" and the book "will therefore appeal more to the antenna specialist or post-graduate student." It "will be found valuable not only as a book of reference on the types of antenna covered, but also for the clear and logical development of the general theory."

CIRCUITS AND CIRCUIT ELEMENTS

621.318.572 1090

Direct-Reading Electronic Timer—R. R. Freas. (*RCA Rev.*, vol. 10, pp. 554–566; December, 1949.) "The application of resistance-coupled multivibrators to binary and decade counter chains is described, with suggestions for improved stability. Several independent counts may be selected and registered on counter dials from one continuously operating electronic counter with the coincidence indicator circuits shown. The electronic counter may be automatically reset to zero or any advance count with the reset circuit presented." Applications to loran receivers are mentioned.

621.318.572:621.317.755.087 1091

A Beam-Switching Circuit—R. Milne. (*Electronic Eng.*, vol. 22, pp. 54–56; February, 1950.) A trigger circuit with particular application to the recording of transient effects. The beam of a cro is switched on for a single sweep of the timebase.

621.318.572:621.385.3:621.315.59 1092

Counter Circuits using Transistors—E. Eberhard, R. O. Endres, and R. P. Moore. (*RCA Rev.*, vol. 10, pp. 459–476; December, 1949.) A flip-flop circuit using two transistors is compared with the Eccles-Jordan circuit and its use in a decade counter is described. A flip-flop circuit using a single transistor is analyzed and stability limits with respect to load resistance and transistor parameters are indicated; practical component values are given for this circuit, which can be operated up to 500 kc. Transistor relaxation oscillators have been synchronized in a frequency divider which converts from 100 kc to 25 cps, and operates from a 45-v supply, from which it takes 675 mw. Present transistors are unstable with time and temperature and differ widely in characteristics and hence cannot yet be used for reliable counting.

621.392.4 1093

Constant-Phase-Shift Networks—W. Saraga. (*Wireless Eng.*, vol. 26, p. 380; November, 1949.) Comment on 3078 of 1949 (Rowlands).

621.392.43 1094

Theoretical Limitations on the Broadband Matching of Arbitrary Impedances—R. M. Fano. (*Jour. Frank. Inst.*, vol. 249, pp. 57–83; January, 1950.) Technical Report No. 41, Massachusetts Institute of Technology, Research Laboratory of Electronics.

621.392.43 1095

Impedance Matching using a Special Type of Exponential Line—A. Niutta. (*Poste e Telecomun.*, vol. 17, pp. 417–423; August, 1949.) For matching low-impedance lines to antennas with high input impedance. Formulas are derived from which the dimensions of a suitable 4-wire exponential line can be calculated. This line is inserted between half-section high-pass filters. Measurements show that the impedance remains satisfactorily uniform over the frequency range 4–24 Mc. A numerical example illustrates the method.

621.392.5 1096

The Solution of Linear-Network Problems by the Method of Nodal Equations—M. Soldi. (*Alta Frequenza*, vol. 18, pp. 213–231; October, 1949.) The practical advantages of the method are illustrated. Suitable equivalent networks are shown for tubes, transformers, and, in particular, the ideal transformer.

621.392.5 1097

Contribution to the Theory of Networks—M. Prudhon and P. M. Prache. (*Ann. Télécommun.*, vol. 3, pp. 24–30; January, 1948.) Relations are established between the impedances of the various branches and the coefficients of the link and nodal equations. Formulas are then derived from which the nodal equations may be simply obtained when the parts of the network are connected by mutual-impedance coupling. Simple applications of the formulas to a transformer and to a tube are given.

621.392.5.018.1:621.396.615 1098

The Effect of Valve Impedance on Phase-Shift Oscillators—R. Townsend. (*Electronic Eng.*, vol. 22, pp. 116–117; March, 1950.) Comment on 840 of May. Curves are given for the frequency and attenuation constants in 3-section and 4-section *RC* networks and also *CR* networks.

621.392.52 1099

Dissipative Band-Pass Filters—F. Juster. (*Radio Prof.* (Paris), vol. 19, pp. 28–30; January, 1950.) Formulas are given for practical application of the methods of design given by Dishal (3369 of 1949).

621.392.52 1100

Filters: Part 2—"Cathode Ray." (*Wireless World*, vol. 56, pp. 61–65; February, 1950.) Discussion of the properties of filters, using the simplest possible mathematics. Part 1, 578 of April.

621.392.53:621.3.015.3 1101

The Exact Solution for Transient Distortion in Networks—D. C. Espley. (*Electronic Eng.* (London), vol. 22, pp. 82–87; March, 1950.) Paper presented at the 1948 Television Congress, Paris. Summary noted in 652 of 1949. A rigorous and systematic method of distortion correction in electrical networks. The distorting system is represented by a hypothetical simulating network defined entirely in terms of input and output waveforms. The input waveform is considered as a pulse with an energy content appropriate to the expected frequency range of the transmission path, and the output waveform as the desired signal appearing after the nominal propagation time and followed by a finite series of echoes defined by amplitude, time, and sign

The correction network is derived by inversion from the distortion simulating network and the passage of the distorted output current through it will produce across its terminals a voltage of the original undistorted form.

621.396.611.1.015.3:621.3.012 1102

Detuned Resonant Circuits—H. Elger. (*Wireless Eng.*, vol. 26, pp. 360–364; November, 1949.) A method of calculating or graphically tracing the transients in such circuits is shown in connection with a simple memory rule. The application of the method to similar mechanical and supersonic problems is considered. The real response of resonant circuits when using PM or any other kind of shock excitation (such as static), the change of time constant due to detuning, and the signal-to-noise ratio are considered.

621.396.611.3:517.2/.3 1103

Time-Constant Selection in the Application of *RC* Differentiating and Integrating Circuits—R. J. Jeffries. (*Instruments*, vol. 22, p. 1106; December, 1949.) The performance of the basic circuit for both operations is presented graphically in terms of M (the product of R in ohms, C in farads, and f in cps) and the phase characteristic. The useful operating ranges of the circuit are found for values of M less than about 0.01 for differentiation and greater than about 3 for integration. Numerical examples are given.

621.396.611.4 1104

On the Characteristics of Electromagnetic Resonant Cavities formed by Two Concentric Spheres—J. Broc. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 198–199; January 9, 1950.) For resonance on a fixed wavelength in the magnetic mode, the outer radius (R_2) of such a cavity is greater than that of a completely hollow sphere resonant on the same wavelength, and as the ratio $s=(R_2/R_2)$ approaches unity, $R_2\rightarrow\infty$. For the electric mode, R_2 is less than the radius of the equivalent hollow sphere and approaches a finite limit, not zero, as the ratio s tends to unity. Calculated values of R_2 and Q for λ 10 cm are tabulated for values of s ranging from 0 to 1.

621.396.611.4 1105

On the Variation of the Natural Frequency of Cavity Resonators formed by Concentric Spheres, for Small Displacements of the Inner Sphere—J. Broc. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 285–286; January 16, 1950.) For the H_{110} mode, a displacement e produces a reduction Δf of the natural frequency f_o, such that $\Delta f/f_o=-2(e/R_2)/(1-s^2)^2$, where $s=R_1/R_2$ and R_1 and R_2 are the radii of the inner and outer spheres respectively. For larger displacements $(e/R_2>0-01)$ the approximations used in the derivation of this expression do not hold. See also 1104 above.

621.396.615 1106

Operating Conditions for Cathode-Coupled Flip Flops—D. W. Thomasson and J. D. Storer. (*Jour. Brit. I.R.E.*, vol. 10, pp. 12–14; January, 1950.) Comment on 3083 of 1949 (Storer) and Storer's reply. The conditions for oscillation and the effects of using different component values are discussed.

621.396.615:621.396.611.21 1107

A High-Stability Quartz Oscillator—M. Indjoudjian. (*Onde Elec.*, vol. 29, pp. 76–78; February, 1950.) The oscillator is basically a 2-stage amplifier with negative feedback. A Y-cut 100-kc crystal vibrating longitudinally couples the anode circuit of the second tube to the grid of the first tube. In this grid circuit is a thermistor indirectly heated by the feedback current from the output transformer secondary; its resistance varies in value from 50 kΩ when cold to 500 Ω during normal working. Variations of oscillation amplitude and of supply voltages produce negligible phase change. The short-period stability during a year's test was $<\pm 0.5\times 10^{-7}$ and the monthly drift of the order of 1×10^{-7}. The test included deliberate tube changing and stoppages. The method of frequency checking is described.

621.396.615.14:621.385.18 1108

Plasma Oscillator—G. Wehner. (*Jour. Appl. Phys.*, vol. 21, pp. 62–63; January, 1950.) Rf oscillations are found to occur in low-pressure arc-discharge tubes in which a repeller electrode is inserted in the electron beam between the grid and anode. The resonant circuits are formed by thin oscillation layers inside the plasma. The oscillations are of well-defined frequency, tunable between 1kMc and 4kMc by changing the anode voltage and current. They occur in different modes; a relation is given between beam velocity, frequency and mode number for a typical tube construction.

621.396.615.14.029.63 1109

New U.H.F. Oscillator—D. H. Preist. (*Electronics*, vol. 23, pp. 120..162; February, 1950.) Advantages claimed for this circuit, which largely resembles a grounded-grid amplifier, are that a wide range of adjustment of phase and amplitude of the feedback is available, adjustments are simple to make, and power-handling capacity is high. Satisfactory results are obtained at frequencies from below 100 Mc to over 2kMc at power levels up to many kilowatts.

621.396.615.17 1110

Asymmetrical Multivibrators—R. Feinberg. (*Wireless Eng.*, vol. 26, pp. 325–330; October, 1949.) "The frequency and the waveform of oscillation of an asymmetrical multivibrator with pentodes can be computed with the help of simple formulas provided the circuit is so designed that one of the two anode voltages is of rectangular or near-rectangular shape when the multivibrator is in the steady state of operation. The waveform of the other anode voltage may be rectangular or triangular and may be varied within a fairly wide range without interfering with the frequency of oscillation. The frequency of oscillation may be altered without disturbing the waveform by changing the capacitance values but keeping their ratio constant. Predicted frequencies and waveforms are verified by experiment."

621.396.615.17 1111

The Miller Circuit as a Low-Speed Precision Integrator—I. A. D. Lewis. (*Electronic Eng.* (London), vol. 22, pp. 66–68; February, 1950.) Errors caused by drifts, by nonlinearity in time (assuming all the circuit elements to be linear) and by the operation of resetting between cycles of integration, are considered in some detail. The analysis includes the effects of stray capacitance in the circuit and is therefore also applicable to high-speed integrators.

621.396.615.17:621.317.755 1112

A Self-Adjusting Time-Base Circuit—H. Asher. (*Electronic Eng.* (London), vol. 22, pp. 61–65; February, 1950.) A type of time-base in which the only frequency control is a range switch which selects one of three wide frequency ranges. Inside these ranges, the frequency of the timebase will be rigidly locked to that of the signal, and at the same time the amplitude of the timebase sweep voltage will remain constant.

621.396.615.18.029.64 1113

Microwave Frequency Dividers—H. Lyons. (*Jour. Appl. Phys.*, vol. 21, pp. 59–60; January, 1950.) Methods of division for frequencies up to 20 kMc or more are of importance for applications such as atomic clocks. A regenerative-modulator type of circuit dividing from 9.3 kMc to 3.1 kMc is described; tests show that the output is one-third of the input frequency to an accuracy within one part in 10^{10}. This circuit uses klystron multiplier and amplifiers, and the amplifiers are needed to work only at the output frequency and not at the higher input frequency. A modified circuit is proposed in which the regenerative-modulator divider is provided with a second regenerative channel and thus provides its own input. A suggested arrangement for an atomic oscillator and clock is described. Suitable equipment for this method, using the ammonia line at 24 kMc, is being investigated, but alternatively a convenient line at about 9 kMc could be used.

621.396.645 1114

Stagger-Tuned Low-Pass Amplifiers—W. E. Thomson. (*Wireless Eng.*, vol. 26, pp. 357–359; November, 1949.) The method of design of band-pass amplifiers by stagger-tuning, in its wider sense, may be simply extended to low-pass amplifiers to specify the frequency response. The determination of networks with the desired frequency response presents a fresh problem. Suitable networks are given for a multistage low-pass amplifier with a gain, as a voltage ratio, of g_m/ω_oC per stage, a bandwidth of $\omega_o/2\Pi$ c/s, or slightly more, and a reasonably flat pass-band; g_m is the mutual conductance of the tubes used, ω_o the nominal bandwidth, and C the interstage capacitance.

621.396.645 1115

The Cathode-Follower Output Stage—R. M. Mitchell. (*Audio Eng.*, vol. 34, pp. 12–13..32; February, 1950.) Advantages and disadvantages are discussed.

621.396.645 1116

On the Input and Output Admittances of Amplifiers—L. Vallese. (*Alta Frequenza*, vol. 18, pp. 205–212; October, 1949.) The input and output admittances of linear amplifiers can be represented by means of networks consisting of the passive quadripole equivalent circuit with the addition of appropriate series or parallel elements, the admittances of which are functions of the amplification factor. The consideration of such networks gives a clear picture of the working of the amplifier.

621.396.645:534.85 1117

Phonograph Reproduction. Part 1—C. G. McProud (*Audio Eng.*, vol. 34, pp. 24..31; February, 1950.) Discussion of the design of a control unit for use with the musician's amplifier (70 of February). Needle scratch is effectively reduced.

621.396.645.029.3 1118

Push-Pull A. F. Amplifiers—K. R. Sturley. (*Wireless Eng.*, vol. 26, pp. 338–343; October, 1949.) "Problems of the position of the composite load line and the construction of the tube-load curves for classes A, B, and C audio-frequency push-pull amplifiers are discussed and it is shown that B. J. Thompson's original work on class-B amplifiers with matched tubes can be extended to cover all classes of push-pull amplification under matched or unmatched conditions. Deductions regarding composite and tube-load curves as a result of voltage and current measurements are confirmed by photographs of tube-load curves obtained on the screen of a cathode-ray tube."

621.396.645.029.3 1119

Study of a High-Quality Voltage Amplifier—J. Mey. (*Ann. Télécommun.*, vol. 4, pp. 383–384; November, 1949.) Description of a 5-stage audio amplifier, with circuit diagram and performance figures. It has low distortion and background noise, high and linear gain, and satisfactory stability with normal supply-voltage variations.

621.396.645.4029./.52:621.396.822 1120
Selective Amplification at Low Frequencies—L. de Queiroz Orsini. (*Onde Élect.*, vol. 29, pp. 91–102; February, 1950.) For parts 1 and 2 see 596 of April. Part 3 describes the apparatus, method, and results of measurement of flicker effect. The following conclusions are drawn: (a) the current due to shot effect in saturated diodes is well represented by Macfarlane's equation (4087 of 1947); (b) at lf tungsten filaments are much better than oxide-coated filaments; (c) the presence of space charge does not affect the spectral distribution of noise; and (d) the equivalent resistance of the lf noise in triodes and the noise voltage at the grid are given by simple formulas and can be calculated approximately from the values given for certain parameters.

621.396.645.35 1121
A D.C. Amplifier Using an Electrometer Valve—D. H. Peirson. (*Electronic Eng.* (London), vol. 22, pp. 48–53; February, 1950.) Description, with circuit details, of a balanced dc amplifier of the bridge type in which the controls are reduced to the minimum required for zero-balancing by use of a highly stable power pack. Under these conditions external temperature variation exerts a limiting effect upon amplifier stability. The effect is minimized by using a new indirectly heated double electrometer tube, Type DBM8A, which has a temperature coefficient not greater than 0.2 mv/°C, a grid current of about 3×10^{-15} A and grid insulation, when carefully cleaned, $>10^{16}\Omega$.

621.396.645.3:621.317.39 1122
A D.C. Amplifier with Cross-Coupled Input—J. N. Van Scoyoc and G. F. Warnke. (*Electronics*, vol. 23, pp. 104–107; February, 1950.) A compact unit with cross-coupled input stage for use with either single-ended or push-pull circuits. Two amplification stages, with a cathode-follower output and feedback amplifier, provide an output of 310 v peak-to-peak to feed a cr tube. Gain is variable in 20 steps of 2 db from a maximum of 50,000. Input impedance is 100 MΩ and frequency response extends from zero to 60 kc. A preamplifier has a maximum usable gain of about 250,000.

621.396.645.371 1123
Negative-Feedback Amplifiers—C. F. Brockelsby. (*Wireless Eng.*, vol. 26, p. 380; November, 1949.) Reply to comment by McLeod (3111 of 1949) on 2768 of 199.

621.396.69 1124
New Radio Components in the World Market—M. Alixant. (*Radio Tech. Dig.* (Franc), vol. 4, pp. 3–41; February, 1950.) New edition of 2770 of 1949 and including, in addition, resistors, potentiometers, *RC* subassemblies, and materials. Most of the novelties brought out since the last Paris Salon de la Pièce détachée are noted.

GENERAL PHYSICS

53.081+621.3.081 1125
The Introduction of the Giorgi System of Units—H. König, M. Krondl, and M. Landolt. (*Tech. Mitt. Schweiz. Telegr.-Teleph. Verw.*, vol. 27, pp. 257–278; December 1, 1949. In French and German.) A general discussion of the system and its practical application. Tables are included giving the corresponding values of all the principal electrical quantities in the cgs, em, and es systems, and in the Giorgi system, with conversion factors.

530.1:621.396.67 1126
Electromagnetic Similitude: General Principles and Application to the Study of the Aerial—S. Colombo. (*Ann. Télécommun.*, vol. 4, pp. 370–376; November, 1949.)

537.212 1127
Effect of an Arbitrary Field on a Sphere or on a Circular Cylinder—É. Durand. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 188–190; January 9, 1950.) A solution of the Laplace equation which is constant over the conductor. Analogous solutions for a dielectric sphere or cylinder are obtained.

537.533 1128
Production and Application of Directional Electron Beams—A. A. Rusterholz. (*Bull. Schweiz. Electrotech. Ver.*, vol. 41, pp. 65–77; February 4, 1950. In German, with French summary.) The development of high-power electron guns is principally a question of design. The efficiency of such systems is limited by certain factors, of which the effect of space charge is of major importance. Approximate methods of investigating the effect of these factors are examined. The best solution of the problem at present appears to be that proposed by Pierce (1419 and 4275 of 1940), which greatly simplifies calculation. As an example, the design of the electron injector for a betatron is considered.

538.114 1129
Physical Theory of Ferromagnetic Domains—C. Kittel. (*Rev. Mod. Phys.*, vol. 21, pp. 541–583; October, 1949.) Survey of domain theory and of the crucial experiments relating to it. The interpretation of coercive force, reversible permeability, hysteresis, and the Barkhausen effect is outlined in terms of domains. The theoretical analysis of domain energy is discussed and expressions are given for the exchange, magnetocrystalline, magnetoelastic, and magnetostatic energy components. The properties of the Bloch wall, across which the spin direction in adjacent domains reverses, are reviewed. Various domain structures are considered theoretically and compared with experimental data derived from magnetic powder pattern technique in which a thin liquid layer containing a colloidal suspension of a fine ferromagnetic powder is applied to the surface of the sample. The single domain behavior of small particles is described; the dependence of initial permeability and coercive force on nonuniformities of the material is outlined.

538.566 1130
The Symmetrical Excitation of an Infinitely Long Dielectric Cylinder—B. Z. Katsenelenbaum. (*Zh. Tekh. Fiz.*, vol. 19, pp. 1168–1181; October, 1949. In Russian.) A mathematical investigation of the em oscillations due to excitation of the cylinder by an elementary dipole mounted on the axis.

538.566 1131
The Asymmetrical Oscillations of an Infinitely Long Dielectric Cylinder—B. Z. Katsenelenbaum. (*Zh. Tekh. Fiz.*, vol. 19, pp. 1182–1191; October, 1949. In Russian.) A mathematical investigation of free and forced em oscillations due to excitation by an elementary dipole.

538.566 1132
On the Focusing of a Wave—J. Ortusi and J. C. Simon. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 521–523; February 6, 1950.) The conditions under which focusing occurs are discussed. The focusing obtained by phase displacement at surfaces with circular holes is contrasted with optical focusing by means of lenses. Experiments with 8-cm waves are briefly described.

539 1133
On the Asymptotic Distribution of Eigenvalues—J. S. de Wet and F. Mandl. (*Proc. Roy. Soc. A*, vol. 200, pp. 572–580; February 22, 1950.) Application of the methods of Courant and Hilbert leads very simply to a proof that the asymptotic distribution of the eigenvalues of the one-dimensional Schrödinger equation is given by the WKB formula. Analogous formulas are also found for the 2- and 3-dimensional equations.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.7 1134
A Tentative Model of the Sun—R. H. Woodward. (*Jour. Geophys. Res.*, vol. 54, pp. 387–396; December, 1949.)

523.72 1135
Investigations of Solar Radiation by the Brazil Expedition of the Academy of Sciences of the U.S.S.R. for Observing the Solar Eclipse of 20th May 1947—S. E. Khaykin and B. M. Chikhachev. (*Compt. Rend. Acad. Sci.* (URSS), vol. 58, pp. 1923–1926; December 21, 1947. In Russian.) A report on observations made in a ship off the Brazilian coast. The results are plotted and discussed; they show (a) that the minimum rf radiation intensity of the eclipsed sun is about 40 per cent of that when the sun is not eclipsed, and (b) that the variation of the intensity is displaced with regard to the geometrical eclipse. It is concluded that the observed rf radiation is generated in the upper layers of the sun's atmosphere not covered by the moon's shadow, and that the radiation intensity is not uniformly distributed over the surface of the radiating sphere.

523.72:621.396.81 1136
Solar Noise and Ionospheric Fading—Smith-Rose. (*See* 1228.)

523.746 1137
On the Mathematical Characteristics of Sunspot-Variations—A. F. Cook. (*Jour. Geophys. Res.*, vol. 54, pp. 347–354; December, 1949.) New relations among the monthly relative numbers and their maxima and minima are applied to Stewart and Panofsky's formula to compute a new family of curves. See also 3454 of 1939 (Stewart and Eggleston).

523.746"1949.07/.09" 1138
Provisional Sunspot-Numbers for July to September, 1949—M. Waldmeier. (*Jour. Geophys. Res.*, vol. 54, p. 398; December, 1949.)

538.12:521.15 1139
The Fundamental Relation between the Magnetic Moment and the Structure of Rotating Celestial Bodies—H. P. Berlage. (*Nature* (London), vol. 165, pp. 242–243; February 11, 1950.) Simple derivation of Blackett's formula, assuming that celestial bodies possess a nucleus from which free electrons migrate into the outer shell more easily than positive ions.

550.38"1949.04/.06" 1140
International Data on Magnetic Disturbances, Second Quarter, 1949—J. Bartels and J. Veldkamp. (*Jour. Geophys. Res.*, vol. 54, pp. 399–400; December, 1949.)

550.38"1949.07/09" 1141
Cheltenham [Maryland] Three-Hour Range Indices *K* for July to September, 1949—R. R. Bodle. (*Jour. Geophys. Res.*, vol. 54, p. 398; December, 1949.)

550.384 1142
Sudden Commencements in Geomagnetism: Their Dependence on Local Time and Geomagnetic Longitude—V. C. A. Ferraro and W. C. Parkinson. (*Nature* (London), vol. 165,

pp. 243–244; February 11, 1950.) Preliminary analysis of data from six observatories.

550.385"1949.04/.09" 1143
Principal Magnetic Storms [April-Sept. 1949]—(*Jour. Geophys. Res.*, vol. 54, pp. 401–402; December, 1949.)

551.508.1:551.594.11 1144
A Radiosonde Method for Potential-Gradient Measurements in the Atmosphere—Kreielsheimer. (*See* 1204.)

551.510.535 1145
Ionosphere Review: 1949—T. W. Bennington. (*Wireless World*, vol. 56, pp. 53–56; February, 1950.) Sunspot cycle and sw propagation survey, with forecast for 1950.

551.510.535 1146
The Problem of the Ionospheric Regions—M. Nicolet. (*Jour. Geophys. Res.*, vol. 54, pp. 373–381; December, 1949. In French, with English summary.) The dissociation and ionization processes in the upper atmosphere due to solar radiation are discussed. Assumptions are made as to the scale height and molecular density at different heights and the most likely processes assigned to each layer. Molecular and atomic oxygen are the main sources of ions in the E and F_1 regions respectively. In the F_2 region atomic nitrogen and oxygen are both important and their relative concentrations are deduced. The conditions under which molecular nitrogen is important are discussed.

551.510.535 1147
On Investigations of the F_2 Layer of the Ionosphere during the Total Solar Eclipse of 20th May, 1947—Ya. L. Al'pert. (*Compt. Rend. Acad. Sci.* (URSS), vol. 58, pp. 1919–1922; December 21, 1947. In Russian.) Using pulse equipment previously described (2893 of 1946), observations of the F_2 layer of the ionosphere were carried out in Brazil from May 17 to 24 inclusive. Measurements of the effective height of the layer were made at a frequency of 9.5 Mc. In addition, measurements were made at various times at frequencies from 5 to 15 Mc. The results obtained are plotted and discussed. One of the conclusions reached is that in addition to variations in the ionization, a considerable expansion of the layer took place during the eclipse. It is considered that measurements of critical frequencies only are not sufficient for studying the properties of the layer.

551.510.535 1148
Nocturnal Ionization in the F_2 Ionospheric Region—N. C. Gerson. (*Rev. Mod. Phys.*, vol. 21, pp. 606–624; October, 1949.) The structure and composition of the earth's atmosphere to a height of 300 km are reviewed and also the photochemical and photoionization processes which occur in sunlight. Whereas the Chapman model accounts well for the E region and roughly for the F_1 region, it is not relevant to the F_2 region. A study of the various oxygen and nitrogen reactions shows that the F_2 region should dissipate during darkness at a greater rate than is observed; consideration is given to possible nonsolar sources of ionization and of these only meteoric bombardment is likely to be significant. Examination of the vertical, horizontal, and turbulent transportation of ions suggests that these may be important in maintaining the nocturnal ion concentration in the F_2 layer.

551.510.535 1149
The Distribution of Atomic and Molecular Oxygen in the Upper Atmosphere—R. Penndorf. (*Phys. Rev.*, vol. 77, pp. 561–562; February 15, 1950.)

551.510.535 1150
Electron Diffusion in the Ionosphere—R. Seeliger. (*Ann. Phys.* (Lpz.), vol. 3, pp. 297–304; August 1, 1948.) Discussion of ionosphere structure and the layer theory. The lack of sufficient reliable observations prevents definite conclusions being drawn.

551.510.535 1151
Stratification of the F_2-Layer of the Ionosphere over Singapore—Hon Yung Sen. (*Jour. Geophys. Res.*, vol. 54, pp. 363–366; December, 1949.) Measurements during January and February, 1946, indicated that the F_2 layer was daily stratified into three discrete layers during the daylight hours. Typical daytime h'/f and diurnal-variation curves are reproduced.

551.510.535 1152
Analysis of Virtual-Height/Frequency Records—U. C. Guha. (*Jour. Geophys. Res.*, vol. 54, pp. 355–362; December, 1949.) For a Chapman layer an approximate expression is derived for virtual height h' at frequency f as a function of f/f_c, where f_c is the critical frequency. The fit is very close both for the theoretical Chapman distribution and also for the theoretical Chapman distribution and also for experimental h'/f curves for the F_2 region with and without a retarding E region. The extension to non-Chapman distributions is discussed.

551.510.535:551.506.2 1153
Ionosphere and Weather—E. Gherzi. (*Nature* (London), vol. 165, p. 38; January 7, 1950.) Discussion of weather forecasting based on a correlation, as yet unexplained, between the usual ionosphere echoes and the movement of polar, maritime, and equatorial air masses. See also 2771 of 1947.

551.510.535:621.396.11 1154
Calculation of the Absorption Decrement for a Parabolic Ionospheric Layer in the Case of Oblique Incidence—É. Argence and K. Rawer. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 69–70; January 2, 1950.) A treatment of the problem by geometrical optics to derive a final expression for the case of reflection. See also 627 of April.

551.510.535:621.396.11 1155
Manifestations of the Sporadic-E Layer on Metre Waves in 1949, and Observation of a Concomitant Noise—Revirieux. (*See* 1222.)

551.510.535:621.396.11 1156
Sporadic Ionization at High Latitudes—J. H. Meek. (*J. Geophys. Res.*, vol. 54, no. 4, pp. 339–345; December, 1949.) See 725 of March.

551.510.535:621.396.11 1157
Effect of the D-Ionospheric Layer on Very Low Frequency Radio Waves—W. Pfister. (*J. Geophys. Res.*, vol. 54, no. 4, pp. 315–337; December, 1949.) See 723 of March.

551.510.535:621.396.812.3 1158
The Analysis of Observations on Spaced Receivers of the Fading of Radio Signals—Briggs, Phillips, and Shinn. (*See* 1233.)

LOCATION AND AIDS TO NAVIGATION

621.396.93 1159
Rotating Aerial for an Automatic Radiogoniometer—P. Bodez. (*Ann. Télécommun.*, vol. 4, pp. 341–346; October, 1949.) Theory and principle of operation of a system comprising three vertical λ/4 dipoles asymmetrically spaced at the corners of a horizontal triangle and turning together on a vertical axis. A field-strength pattern with a single sharp minimum is obtained. An experimental model operating on 120 Mc gave satisfactory results. Sensitivity is about the same as with Adcock systems.

621.396.932.1/.2 1160
A New Direct-Reading Loran Indicator for Marine Service—F. E. Spaulding, Jr., and R. L. Rod. (*RCA Rev.*, vol. 10, pp. 567–585; December, 1949.) The pulse rate of the receiver-indicator is automatically controlled by means of a pulse-phase discriminator and a reactance tube which correct the frequency of the receiver's 100-kc cyrstal oscillator. Frequency division from 100 kc is performed by binary and decade counters using double-triode trigger circuits (see 1090 above); any one of 24 station pulse rates may be selected. After matching the pulses on the screen of a 3-in. cr tube, time-difference readings are directly displayed on a five-digit dial. An extra sweep is available for self-checking the performance and accuracy; the error due to signals differing in amplitude by as much as 1000:1, with 10-v maximum peak signal, does not exceed 0.5 μsecond.

621.396.933 1161
Electronic Aids to Air Navigation—A. A. McK. (*Electronics*, vol. 23, pp. 66–71; February, 1950.) Elementary and brief accounts of the principal features of various navigation aids now in use or about to be used in U.S. civil aircraft. These include compass systems, direction finders, radio ranges, distance measuring equipment, approach and landing equipment, and markers.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.37 1162
On the Kinetics of the Luminescence of Sulphide Phosphors with Several Activators—M. Schön. (*Ann. Phys.* (Lpz.), vol. 3, pp. 333–342; August 1, 1948.)

535.37 1163
Note on the Width of the Emission Bands of Sulphide Phosphors—M. Schön. (*Ann. Phys.* (Lpz.), vol. 3, pp. 343–344; August 1, 1948.)

535.37 1164
Effect of Temperature on the Extinction of the Luminescence of Sulphides by Electric Fields—J. Mattler. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 76–77; January 2, 1950.)

537.58:539.234 1165
Thermionic Emission of Thin Films of Alkaline Earth Oxide Deposited by Evaporation—G. E. Moore and H. W. Allison. (*Phys. Rev.*, vol. 77, pp. 246–257; January 15, 1950.) Monomolecular films of BaO and SrO on tungsten and molybdenum surfaces are shown to give thermionic emissions of the same magnitude as commercial oxide-coated cathodes. The effect is qualitatively explained by considering the adsorbed molecules as oriented dipoles, and the theory is compared with the semiconductor theory for bulk oxide coatings. Full experimental details are given.

538.221 1166
Saturation Magnetization of Certain Ferrites—L. Néel. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 190–192; January 9, 1950.) A study of the different affinities of the divalent M ion in ferrites of the type Fe_2O_3MO, from consideration of the energy expended in movement of the M ion from an octahedric to a tetrahedric position. A formula for the magnetic moment at saturation at absolute zero of the Fe_2O_3MO molecule gives results in good agreement with experiment for ferrites in which M is replaced by Mn, Fe, Co, Ni, Cu, or Zn. See also 3159 of 1949.

38.221 **1167**

Spontaneous Magnetization of Ferromag- etic Ferrites of Spinel Structure—E. W. orter. (*Compt. Rend. Acad. Sci.* (Paris), ol. 230, pp. 192–194; January 9, 1950.) xperimental results which verify Néel's heory (1166 above).

38.221 **1168**

Magnetization Curves and Energies, Coer- ive Field and Magnetostriction of a Ferrite f Cobalt [vectolite]—C. Guillaud, R. Vautier, nd S. Medvedieff. (*Compt. Rend. Acad. Sci.* Paris), vol. 230, pp. 60–62; January 2, 1950.)

538.221 **1169**

On the Decrease of the Thermo-Remanent and Isothermally-Remanent Magnetizations of Fired Earths by Reheating at Successively Higher Temperatures—J. Roquet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 282–285; January 16, 1950.) See also 902 of May.

538.221 **1170**

The Molecular-Field Coefficients for Mixed Ferrites of Nickel and Zinc—L. Néel and P. Brochet. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 280–282; January 16, 1950.)

538.221 **1171**

Saturation Magnetization of Mixed Ferrites of Nickel and Zinc—L. Néel. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 375–377; January 23, 1950.) Some experimental results are interpreted satisfactorily by the author's theory (1166 above and 3159 of 1949).

538.221:621.318.323.2.042.15 **1172**

Effective Permeability and *Q* Factor of Magnetic Powders—A. Colombani. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 523–525; February 6, 1950.) A derivation of general formulas.

538.221.029.64 **1173**

The Properties of Ferromagnetic Compounds at Centimetre Wavelengths—J. B. Birks. (*Proc. Phys. Soc.*, vol. 63, pp. 65–74; February 1, 1950.) Measurements of the magnetic and dielectric properties of λ-Fe_2O_3, Fe_3O_4, Mn/Zn, and Ni/Zn ferrites at wavelengths in the range 60–1.23 cm are described and discussed.

538.222 **1174**

Magnetic Properties of Palladium Alloys—J. Wucher. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 1309–1310; December 14, 1949.) Results of measurements of the paramagnetism of alloys of Pd with Cu, Ag, Au, Al, Sn, Pb, and Al, and comparison with results for Ni alloys.

539.23:537.311.1 **1175**

The Resistivity of Thin Metallic Films—R. A. Weale. (*Proc. Phys. Soc.*, vol. 62, pp. 135–136; February 1, 1949.) A formula is derived for the effective temperature coefficient (α') of resistance of thin films; this shows that α' is zero for a particular thickness of any given material and that it should be negative for smaller thicknesses. It is to be expected that α' will be negative for comparatively thick films of Bi, in which metal the mean free path of the electrons is exceptionally long. The results of van Itterbeek and de Greve on Ni films (3305 of 1946) are in good agreement with the formula given.

546.287:621.315.61 **1176**

Silicones: Composition; Properties; Applications—O. Regert-Monod and P. Seguin. (*Ann. Télécommun.*, vol. 4, pp. 431–440; December, 1949.) A comprehensive study of commercial silicones.

546.821:[537.533+621.3.011.2 **1177**

The Physical Properties of Titanium: Part 1—Emissivity and Resistivity of the Commercial Metal—W. C. Michels and S. Wilford. (*Jour. Appl. Phys.*, vol. 20, pp. 1223–1226; December, 1949.)

549.514.51:621.924 **1178**

Mechanical Production of Very Thin Oscillator Plates—L. T. Sogn and W. J. Howard. (*Bur. Stand. Jour. Res.*, vol. 43, pp. 459–464; November, 1949.) The thickness at which the crystal carrier stretches or buckles is the limiting factor in conventional apparatus. To produce quartz oscillator plates thinner than 0.005 inch (i.e., for operation above 14 and 20 Mc for AT and BT cuts respectively) the usual lapping methods and machinery have been modified by replacing the conventional top lapping plate and changing the crystal carrier correspondingly. The development of different methods is traced from the simple pressure-block method to the more elaborate forms using automatic truing. The improved equipment is capable of producing quartz crystals 0.001 inch thick, and equally thin wafers of other materials. For a shorter account see *Radio and Telev. News, Radio-Electronic Eng. Supplement*, vol. 14, pp. 7, 30; February, 1950.

620.197:621.314.2 **1179**

Some Mechanical Aspects of Hermetically Sealed Transformer Technique—C. Evans. (*Jour. Brit. IRE*, vol. 10, pp. 20–36; January, 1950.) A discussion of the problems of sealing oil-filled and air-filled transformers, and of practical methods recently used.

621.319.4:620.197.119 **1180**

Hermetic Sealing of Capacitors—P. R. Coursey. (*Proc. IEE* (London), Part III, vol. 97, pp. 56–64; January, 1950.) A review of the development of modern types of terminal insulation.

669.15.26:666.1.037.5:621.385.832 **1181**

The Evaluation of Chromium-Iron Alloys for Metal Kinescope Cones—A. S. Rose and J. C. Turnbull. (*RCA Rev.*, vol. 10, pp. 593–599; December, 1949.) Description of tests to determine the suitability of Cr/Fe alloys for sealing to the glasses used in the production of cr tubes with metal cones.

MATHEMATICS

517.432.1 **1182**

Justification of Heaviside Methods—J. J. Smith and P. L. Alger. (*Elec. Eng.* vol. 69, p. 116; February, 1950.) Digest of AIEE 1949 Summer Meeting paper. The Schwartz method of representing a discontinuous function as a distribution in mass-space, gives equations exactly parallel to those obtained by use of Heaviside's method.

681.142 **1183**

Static Magnetic Storage and Delay Line—An Wang and Way Dong Woo. (*Jour. Appl. Phys.*, vol. 21, pp. 49–54; January, 1950.) "Magnetic cores with a rectangular hysteresis loop are used in a storage system which requires no mechanical motion and is permanent. The binary digit '1' is stored as a positive residual flux, and the binary digit '0' as a residual flux in the opposite direction. When a negative probing field is applied to the core, a large voltage is induced in another winding if the digit stored has been a '1,' and very small voltage if it has been a '0.' The induced voltage in the former case is large enough to magnetize another core of identical construction. Binary digits can thus be transferred from one core into another. Many cores are arranged in tandem to form an information delay line. Binary digits can be advanced along the line step by step. The present upper limit of the speed of propagation is about 35,000 digits per second, and there is no lower limit."

681.142 **1184**

Use of the Relay Digital Computer—E. G. Andrews and H. W. Bode. (*Elec. Eng.* vol. 69, pp. 158–163; February, 1950.) The machine whose use is described is the Bell Telephone Laboratories Model V computer. Control is by perforations on teletype tapes. Illustrative applications include binomial expansions in probability theory, and the solution of ordinary and partial differential equations. A simple example of coding is given. For efficient operation, problems should be repetitive.

MEASUREMENTS AND TEST GEAR

531.76 **1185**

On the Use of Crystal Controlled Synchronous Motors for the Accurate Measurement of Time.—V. E. Hollinsworth. (*Canad. Jour. Res.*, vol. 27, pp. 470–478; December, 1949.) An illustrated description of an electronic frequency converter which uses a 1 kc supply controlled by the 50-kc quartz frequency standard of the Dominion Observatory to derive sufficient power for operating several standard 60-cps motors. These are used for driving drum chronographs and other recorders. Circuit diagrams are given of the 100-cps and the 10-cps frequency dividers, the 60-cps frequency multiplier, the 60-cps power amplifier, and the supply unit.

621.317.2.089.6 **1186**

A Royal Air Force Calibration Centre—W. H. Ward and the staff of the Measurements Division, TRE (*Proc. IEE* (London), Part III, vol. 97, pp. 49–55; January, 1950.) Discussion of the general principles governing the planning of a center for maintaining the calibration of RAF radio and radar test equipment, and some details of the methods and apparatus used.

621.317.3.029.5/.6 **1187**

Measurements at Radio Frequencies—G. A. Day. (*Aust. Jour. Instr. Tech.*, vol. 5, pp. 190–210; September, 1949.) Outline of rf methods used for the measurement of various electrical quantities.

621.317.73.029.62./.63 **1188**

Discontinuities in Concentric-Line Impedance-Measuring Apparatus—M. H. Oliver. (*Proc. IEE* (London), Part III, vol. 97, pp. 29–38; January, 1950.) Three methods of investigating such discontinuities are described: (a) the narrow-band frequency-variation method; (b) the wide-band frequency-variation method; and (c) the reactance-variation method. The last method has certain advantages. The effect of discontinuities on the measurement of resistance and reactance, of conductance and susceptance, and of cable characteristics, is discussed and practical measurement techniques are suggested.

621.317.733.083.4 **1189**

Differential-Amplifier Null Detector—M. Conrad. (*Electronics*, vol. 23, pp. 96–97; February, 1950.) The advantages of this instrument are reduction in weight and cost, very high input impedance, guarded shielding, and adaptability to very low frequencies where transformers are not advantageous. On the other hand, it requires balancing, and in general it is more complex than a transformer. When used as a null-detector for an equal-arm ac resistance bridge, the instrument will detect an unbalance of less than 0.1 per cent, and the discrimination against common-mode signals is greater than 70 db.

621.317.77 1190

An Improved Audio-Frequency Phase Meter—O. E. Kruse and R. B. Watson. (*Audio Eng.* vol. 34, pp. 9–11, 46; February, 1950.) The instrument described can be used to measure without ambiguity the phase angle between two sinusoidal signals in the frequency range 40 cps to 29 kc with an error <2 per cent. Up to 20 kc the error is <1 per cent. An alternative circuit arrangement overcomes the instability of reading which normally occurs in a phase meter when the phase angle is very near 0° or 360°.

621.317.78.029.65 1191

Discrepancies in the Measurement of Microwave power at wavelengths below 3 cm—J. Collard, G. R. Nicoll, and A. W. Lines. (*Proc. Phys. Soc.*, vol. 63, pp. 215–216; March 1, 1950.) Measurements of 20-mw power were made at TRE for the wavelength range 7.5–12.5 mm, using (a) a water calorimeter, (b) an enthrakometer (2020 of 1948), and (c) a thermistor bead. More accurate comparisons between enthrakometer and thermistor for a power of 1 mw and wavelength 9 mm, and between enthrakometer and calorimeter for a power of 50 mw and the same wavelength, were consistent with the direct comparison of all three instruments. The results obtained with the enthrakometer and the calorimeter agreed to within a few per cent, but the readings given by the thermistor were only about half those obtained by the other two methods. Thermistor beads of a particular type give fairly consistent results, so that it may be possible to dispense with individual calibration and apply a simple correction factor.

621.396:621.3.018.4(083.74) 1192

[British] Standard-Frequency Transmissions—(*RSGB Bull.*, vol. 25, p. 264; February, 1950.) Daily transmissions, arranged by the Department of Scientific and Industrial Research, commenced on February 1, 1950, from the GPO station at Rugby on frequencies of 60 kc (1029 to 1045 GMT), 5 Mc (0544 to 0615 GMT), and 10 Mc (0629 to 0700 GMT). The transmitter power in each case is 10 kw. The frequencies are maintained within 2 parts in 10^8 of the nominal frequencies and are monitored at the National Physical Laboratory. The transmissions are arranged in cycles of 15 minutes, each starting with a 1-minute transmission of the Morse call sign MSF, with spoken announcements; the carrier wave is then modulated at 1,000 cps for 5 minutes and is unmodulated for the remaining 9 minutes. 1-cps pulses will at some future date be added during the first 5 minutes of the unmodulated transmission. See also 134 of 1949.

621.396.662.029.6 1193

Calibrated Piston Attenuator—A. C. Gordon-Smith. (*Wireless Eng.*, vol. 26, pp. 322–324; October, 1949.) The attenuator, which operates in the evanescent H_{11} or TE_{11} circular mode, consists of a pair of telescopic brass tubes. "The outer tube has a diameter such that when filled with air the attenuation is about 7 db per mm. The inner tube is filled with polystyrene and acts as a nonattenuating waveguide for the H_{11} mode but does not transmit other modes. The attenuation per mm of movement has been measured, after frequency conversion, in terms of a standard piston attenuator and agreement obtained between the measured and theoretical values within the limits of experimental accuracy. It has also been confirmed that the crystal mixer and its associated circuit is a linear converter at a wavelength of 6 mm."

621.397.62.001.4 1194

Measurement of Transient Response of TV Receivers—J. Van Duyne. (*TV Eng.*, vol. 1, pp. 14–18; January, 1950.) Discussion of the requirements of measurement equipment for studying the electrical fidelity of the picture channel, and description of suitable apparatus. 60 cps, 2.4 kc, and 94.5 kc are selected as the repetition rates for the rectangular test signals.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.767 1195

Gas-Flow Speedometer—G. L. Mellen. (*Electronics*, vol. 23, pp. 80–81; February, 1950.) Measurement of the transit time of an ion cloud in the gas over a known distance gives rate of flow without introducing impurities. Range of speeds is 20–400 mph.

534.321.9:616.36-073 1196

Lupam—an Ultrasonoscope Locator for Medical Applications—R. P. McLoughlin and G. N. Guastavino. (*Rev. Telegr.* (Buenos Aires), vol. 38, pp. 507–517, 543; September, 1949.) Basic principles for such units are reviewed, and equipment is described in which the reflection of ultrasonic pulses is applied to locate gallstones or other small objects. Cathode-ray-oscillograph presentation similar to type-A radar display is used. Calibration methods are indicated.

538.569.2.047:621.38 1197

Dielectric Properties of Some Animal Tissues at Meter and Centimeter Wave Lengths—E. R. Laird and K. Ferguson. (*Canad. Jour. Res.*, vol. 27, pp. 218–230; November, 1949.) Measurements of the dielectric constant and the absorption coefficient of different tissues, using wavelengths of 1.72 m, 9.5 cm, and 3.2 cm. For most of the substances tested the dielectric constant decreased considerably with increasing frequency; the absorption, on the other hand, increased largely.

539.16.08 1198

Photomultipliers for Scintillation Counting—G. A. Morton. (*RCA Rev.*, vol. 10, pp. 525–553; December, 1949.) A review of various commercially available multipliers and discussion of their characteristics.

539.16.08 1199

A Stabilizer for Proportional Counters—D. H. Wilkinson. (*Jour. Sci. Instr.*, vol. 27, pp. 36–38; February, 1950.) The introduction of an α-particle group into a counter enables high precision to be attained over long periods by deriving control of the voltage applied to the counter from this group. Theory of the method is given and a suitable circuit is described.

539.16.08 1200

A Portable Geiger Counter—M. Michaelis and R. O. Jenkins. (*Electronic Eng.*, vol. 22, pp. 112–115; March, 1950.) Detailed description of equipment including a standard GEC type GM2 or GM4 self-quenching tube and suitable for either β-, λ-, or X-ray detection.

539.16.08 1201

Temperature Effects in the Spurious Discharge Mechanism of Parallel-Plate Counters—F. L. Hereford. (*Phys. Rev.*, vol. 77, pp. 559–560; February 15, 1950.)

539.16.08:549.211 1202

Effect of Light on a Diamond Conduction Counter—R.K. Willardson and G. C. Danielson. (*Phys. Rev.*, vol. 77, pp. 300–301; January 15, 1950.) The accumulation of space charge field resulting from the trapping of charge carriers reduces the rate of counting. The effect may be completely eliminated in some diamonds by irradiation with violet light. After treatment the diamond is in an activated condition and the counting rate is maintained indefinitely. The effect is thus different from the ordinary release of space charge by red light, which has only a temporary effect on the counting rate.

539.16.08:615.849 1203

Geiger-Müller Tubes in Industrial Radiography—O. J. Russell. (*Electronic Eng.*, vol. 22, pp. 94–98; March, 1950.)

551.508.1:551.594.11 1204

A Radiosonde Method for Potential-Gradient Measurements in the Atmosphere—K. Kreielsheimer. (*The New Zealand Science Congress*, pp. 91–98; 1947.) Details of modifications of the Bureau of Standards radiosonde (315 of 1941) to adapt it for recording potential gradients, and discussion of results obtained. A preliminary account was noted in 1955 of 1946 (Kreielsheimer and Belin).

621.318.572 1205

An Electrical Transient Display System—J. A. Lyddiard and J. W. Osselton. (*Jour. Sci. Instr.*, vol. 27, pp. 38–41; February, 1950.) Description of apparatus designed to facilitate the observation of transients in networks including iron-cored reactors operated at 50 cps, by providing a succession of identical transients which can be viewed for any length of time on the screen of a cro. Electronic methods are used. Switching cycles are accurately controlled. The equipment can also be used to generate square waves of very low frequency.

621.38.001.8:578.088.7 1206

An Electronic Stimulator for Biological Research—V. H. Attree. (*Jour. Sci. Instr.*, vol. 27, pp. 43–47; February, 1950.) Circuit details and operation of a 0.005–500-cps generator with three outputs simultaneously available, one providing exponential pulses, the other two giving square pulses.

621.38.001.8:616-07 1207

Electronic Instruments in Diagnostic Medicine—H. A. Hughes. (*Electronic Eng.* (London), vol. 22, pp. 43–47 and 88–93; February and March, 1950.) Discussion of equipment for electroencephalography, electrical stimulation, electrocardiography, etc.

621.384.6 1208

Phase Focusing in Linear and in Spiral Accelerators—W. Dällenbach. (*Ann. Phys.* (Lpz.), vol. 3, pp. 89–100; August 1, 1948.)

621.384.611.1 1209

Betatrons with and without Iron Yoke—A. Bierman and H. A. Oele. (*Philips Tech. Rev.*, vol. 11, pp. 65–78; September, 1949.) The fundamental principles of the betatron are outlined, with a description of two recently constructed machines. One, on conventional lines, used an iron circuit for the magnetic field, with 500-cps ac excitation, and accelerates electrons to 5 Mev. In the second apparatus the magnetic field is obtained by means of coils through which the discharge current from a battery of capacitors flows. A small iron core weighing only 5 kg is used. The flux, being obtained from damped oscillations, the frequency being 2,500 cps, reaches the required intensity only during the first few cycles of each discharge and thus the output from the apparatus is pulsed. Electron energies of 9 Mev are obtained. The practical elimination of iron in this model results in a very great saving in weight. See also 2015 of 1949 (Bierman).

621.385.833 **1210**

The Field in an Electron-Optical Immersion Objective—L. Jacob. (*Proc. Phys. Soc.*, vol. 63, pp. 75–83; February 1, 1950.)

621.385.833 **1211**

Electron Optical Mapping of Electromagnetic Fields—L. Marton and S. H. Lachenbruch. (*Jour. Appl. Phys.*, vol. 20, pp. 1171–1182; December, 1949.) See also 967 of May.

621.385.833 **1212**

Fringe Fields of Ferromagnetic Domains—L. Marton, S. H. Lachenbruch, J. A. Simpson, and A. Van Bronkhorst. (*Jour. Appl. Phys.*, vol. 20, p. 1258, December, 1949.) Photographs showing (a) the distorted mesh pattern due to the fringe field along the edge of a cobalt single crystal, and (b) the pattern obtained with a small $BaTiO_3$ crystal, using the new "shadow" technique (199 of February), are briefly discussed.

621.385.833 **1213**

On the Theory of the Independent Electrostatic Lens with Three Diaphragms—É. Regenstreif. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 1311–1313; December 14, 1949.) Formulas are derived which give results in good agreement with experiment and permit accurate determination of the voltage which must be applied to the central diaphragm to convert the lens into a reflector.

621.385.833:669 **1214**

Metallurgical Applications of the Electron Microscope—(*Engineering* (London), vol. 168, pp. 625 and 652; December 9 and 16, 1949.) Report of the meeting organized by the Institute of Metals in November, 1949, with summaries of 13 papers presented and of the discussions on them. See also *Nature* (London), vol. 165, pp. 390–393; March 11, 1950.

621.398 **1215**

Remote Control by A.F. Discrimination—R. B. McNeil. (*Electronics*, vol. 23, pp. 142. 150; February, 1950.) The equipment is designed to have a bandwidth no greater than that used for normal AM communications, and considerably simplified control equipment, which comprises a tone-modulated transmitter at each master station, and an amplifier-relay unit at each of the relay stations, operating in conjunction with a receiver and constant-output amplifier. The modulator unit may be used for voice modulation of the transmitter. The equipment operates well through interfering signals and noise.

PROPAGATION OF WAVES

538.566.2+534.231 **1216**

A New Method for Solving the Problem of the Field of a Point-Radiator in a Multilayer Non-Uniform Medium—L. M. Brekhovskikh. (*Bull. Acad. Sci.* (URSS), vol. 13, pp. 409–420; July and August, 1949. In Russian.) The theory of propagation of waves in multilayer media is important not only for electromagnetic but also for sound waves, especially when the latter are propagated in the sea, where ducts may be formed. In this paper the field due to a point-radiator of either electromagnetic or sound waves is investigated mathematically. The spherical wave is resolved into a number of plane waves which are integrated in a complex plane.

538.566.2 **1217**

The Reflection of Plane Waves from Multilayer Non-Uniform Media—L. M. Brekhovskikh. (*Zh. Tekh. Fiz.*, vol. 19, pp. 1126–1135; October, 1949. In Russian.) Mathematical analysis for multilayer media with arbitrary variation of the parameters. The theory proposed is not based on the wave equation but on an equation of the first order of the reflection coefficient. Two methods of successive approximations are indicated for expressing the reflection coefficient in the form of convergent series. An example shows the rate of convergence of these series.

538.566.2 **1218**

Application of Huyghens' Principle to Refraction. Expressions for the Reflected and Refracted Waves—J. Brodin. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 67–69; January 2, 1950.)

621.396.11:535.42 **1219**

On the Diffraction of Radar Waves by a Semi-Infinite Conducting Screen—C. W. Horton and R. B. Watson. (*Jour. Appl. Phys.*, vol. 21, pp. 16–21; January, 1950.) Measurements of the diffraction patterns of cm waves at the edge of a semi-infinite copper screen are described. A receiving horn was revolved about the center of the edge at a radius of 15" or 30" and angles of incidence upon the screen of 0° and—222° were investigated. The results are shown as polar diagrams, which agree well with theoretical patterns.

621.396.11:551.510.535 **1220**

Ionospheric Cross-Modulation at Oblique Incidence—L. G. H. Huxley. (*Proc. Roy. Soc. A*, vol. 200, pp. 486–511; February 22, 1950.) Systematic investigation of ionospheric cross-modulation, using selected pairs of BBC transmitters in special night-time tests. Evidence supports the assumption that the seat of cross-modulation is highly localized in the *E* region, in one or two regions near the lower of the surfaces of reflection of the wanted and disturbing waves. The quantities measured are $G\nu$, (ν is the mean electron collisional frequency and G a constant for the gas at the seat of cross-modulation) and T_0, the percentage transferred modulation depth at "zero" modulation frequency. Nocturnal and seasonal changes in these quantities are discussed. An estimate is made of the gradients of electron concentration and collision frequency at a height of about 86 km.

621.396.11:551.510.535 **1221**

Discussion of the Ionosphere Reflection Factor. Scattering of Ground-Wave Energy and Its Conversion into Space-Wave Energy—C. Glinz. (*Tech. Mitt. Schweiz. Telegr.-Teleph. Verw.*, vol. 27, pp. 279–283; December 1, 1949. In German.) Methods of calculating the reflection coefficient are discussed. Vilbig (887 of 1939) considers that a relation exists between this coefficient and ground conductance. It seems that such a relation can only be apparent, since in the wavelength range 200–2,000 m ionospheric absorption cannot be neglected. From the CCIR field-strength curves it can only be deduced with certainty that the indirect radiation reaches a maximum for distances of 400–600 km. In the area covered by the direct radiation of a transmitter, part of the energy is scattered by irregularities of the ground. Such scattering increases the ratio of space-wave to ground-wave energy, so that modifications of a transmitting antenna designed to reduce the ratio do not result in the calculated improvement in reception. The ground wave undergoes no appreciable dispersion when propagation is over the sea. This may partly explain the dependence of the indirect radiation on the ground conductance.

621.396.11:551.510.535 **1222**

Manifestations of the Sporadic-*E* Layer on Metre Waves in 1949, and Observation of a Concomitant Noise—P. Revirieux. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, p. 200; January 9, 1950.) Manifestations were frequent in June and July, on 50–52 Mc. June 18 is cited as a favorable example. After 1,700 GMT, a 100-w transmitter at Oslo produced a field strength at Meudon of above 100 μv/m on 50.1 Mc. Other distant stations were heard. During this time a characteristic noise was received on 145 Mc, using a Yagi antenna with 4 horizontal elements. The noise level was some 5–10 db above that of the receiver. The direction of the noise source coincided with that of the E_s layer causing the long-range reception on 50 Mc.

621.396.11:551.510.535 **1223**

Effect of the *D* Ionospheric Layer on Very-Low-Frequency Radio Waves—W. Pfister. (*Jour. Geophys. Res.*, vol. 54, pp. 315–337; December, 1949.) See 723 of April.

621.396.11:551.510.535 **1224**

Sporadic Ionization at High Latitudes—J. H. Meek. (*Jour. Geophys. Res.*, vol. 54, pp. 339–345; December, 1949.) See 725 of April.

621.396.11:551.10.535 **1225**

Calculation of the Absorption Decrement for a Parabolic Ionospheric Layer in the Case of Oblique Incidence—Argence and Rawer. (*See* 1154.)

621.396.11:621.396.81 **1226**

A Relation between the Sommerfeld Theory of Radio Propagation over a Flat Earth and the Theory of Diffraction at a Straight Edge—H. G. Booker and P. C. Clemmow. (*Proc. IEE* (London), Part III, vol. 97, pp. 18–27; January, 1950.) The field strength at a point above an imperfectly reflecting flat earth, due to vertically polarized radiation from a line source parallel to the earth's surface, is considered as that due to reflection by a perfectly conducting earth added to that due to the Zenneck wave diffracted, in the two-dimensional case, under the image line. For the three-dimensional case, the diffraction would be through a slot extending downwards from the image point. For points well above the earth's surface this diffracted wave is the edge-wave from the diffracting edge. The field strength is then the same as calculated on the ray theory with the ordinary Fresnel image of the source in the imperfectly conducting source. For points close to the earth's surface, the diffracted wave must be calculated using the Fresnel integral. This gives the Sommerfeld formula for propagation over a flat earth. The advantages of this derivation of Sommerfeld's formula in considering propagation over a land/sea boundary are briefly indicated.

621.396.11.029.55 **1227**

New Observations on the Doppler Effect in the Propagation of Decametre Radio Waves—B. Decaux and M. Crouzard. (*Compt. Rend. Acad. Sci.* (Paris), vol. 227, pp. 116–117; July 12, 1948.) From observations in Paris of the standard-frequency transmissions from Washington on 5, 10, 15, and 20 Mc from April 12 to 15, 1948, between 0800 and 2200 GMT, and from former observations 1725 of 1948), the mean values of frequency variation during 24 hours are plotted. A positive maximum appears during sunrise over the propagation path; a corresponding frequency minimum occurs during sunset. The maximum frequency changes range from 5×10^{-8} at 20 Mc to 20×10^{-2} at 5 Mc. Between 1,400 and 1500 GMT and at 0200 GMT no variation is observed.

621.396.81:523.72 **1228**

Solar Noise and Ionospheric Fading—R. L. Smith-Rose. (*Nature* (London), vol. 165, pp. 37–38; January 7, 1950.) Solar noise measurements at Slough on 30, 42, 73, and 155 Mc, made simultaneously with field-strength measurements of signals from South Africa and Germany on 18.89 Mc and 191 kc show that bursts of noise are frequently accompanied by a fade-out of the hf signal and a marked fluctuation in level of the lf signal.

An outstanding example, which occurred on May 21, 1948, is discussed in detail; the variation of the lf signal is characteristic of interference between ground and ionosphere waves caused by a change in phase and amplitude of the latter. This can be explained by a change in the equivalent height of reflection, together with an increase in the effective reflection coefficient of the ionosphere.

621.396.81:551.510.535 1229

Field Intensity at the Receiver as a Function of Distance—P. Lejay and D. Lepechinsky. (*Nature* (London), vol. 165, pp. 306–307; February 25, 1950.) A graphical method is described for finding, from a vertical-incidence $h'f$ trace, the virtual path followed by a wave reflected obliquely from the ionosphere. A reflectrix, or locus of the reflection point for the virtual ray as the angle of incidence is varied, can be found by a simple graphical construction if a plane earth is assumed. For a curved earth and curved ionosphere a more laborious method, not described, must be used. From the shape of the reflectrix the field set up at a distant point on the earth's surface can be found. Focusing and interference effects produce a high field, rapidly fluctuating with distance, just beyond the skip distance, and a high field at the limiting distance for the tangential ray.

621.396.81:551.510.535 1230

'*M*-Mode' Propagation Possibilities—D. Lepechinsky (*Nature* (London), vol. 165, p. 307; February 25, 1950.) The ionization density of the *E* layer being greatest at the subsolar point, a two-hop ray passing near this point may be unable to penetrate the *E* layer to reach the ground, and be reflected instead back to the *F* layer, with resulting lower absorption. Outstandingly good reception has been observed on frequencies of 19 Mc and 14 Mc when this type of propagation had been predicted.

621.396.81.029.45 1231

Round-the-World Signals at Very Low Frequency—J. N. Brown. (*Jour. Geophys. Res.*, vol. 54, pp. 367–372; December, 1949.) Pulses were radiated from an omnidirectional antenna at Annapolis, Maryland, and were received on a vertical loop 50 miles away. Transmitter power was 350 kw and frequency 18 kc. In March, 1949, the delay time was 0.1373 sec, corresponding to about 55 hops between an ionosphere layer at a height of 65 km and the ground. During an ionospheric storm the time fell to 0.1365 sec. The maximum attenuation of round-the-world signals is about 70 db; near local sunset it falls to a sharp minimum of about 56 db.

621.396.81.029.58 1232

Study of the Propagation of Decametre Waves by Means of Standard-Frequency Transmissions—B. Decaux, M. Barré, and G. Bertaux. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 378–380; January 23, 1950.) A résumé of the measurements of the frequencies and field strengths at Bagneux of the transmissions from Washington WWV and Hawaii WWVH. During breaks in the WWVH transmission, frequency stability of the WWV signal clearly improved. Permanent recordings were made at Bagneux of the 15 and 20-Mc carriers and of the 440-cps modulation. Better reception of WWVH on 15 Mc, despite greater distances and lower power, was particularly evident in the early morning. Identical conclusions can be drawn from records obtained on the S. S. *Commandant Charcot* during her voyage between South Africa and Australia, especially in the late afternoon and at nightfall. Typical records are reproduced. See also 1227 above and 1725 of 1948 (Decaux).

621.396.812.3:551.510.535 1233

The Analysis of Observations on Spaced Receivers of the Fading of Radio Signals—B. H. Briggs, G. J. Phillips, and D. H. Shinn. (*Proc. Phys. Soc.*, vol. 63, pp. 106–121; February 1, 1950.) The fading of a radio wave once reflected from an irregular ionosphere is discussed in terms of the variable diffraction pattern produced at the ground. Fading may arise either from a drift of the pattern past a receiver, or from irregular time variations in the pattern, or both. Observations at three receiving points can be used to deduce the rate at which the pattern changes and the drift velocity with respect to the ground. For practical results see 96 of February (Mitra).

RECEPTION

621.396.621 1234

Denco Model DCR19 Communications Receiver—(*Wireless World*, vol. 56, pp. 50–52; February, 1950.) Test report of a receiver with six overlapping ranges covering the band 175 kc–36Mc. Special features include a crystal calibrator and a well-made rotary coil-turret which removes all idle coils from the circuit and short-circuits any that might act as absorbers for the coils in use. Circuit details are given of the interstage coupling following the first intermediate-frequency tube, and also of the first amplitude-frequency amplifier, which includes a negative-feedback circuit acting as a filter.

621.396.621 1235

Communication Receiver Design—D. W. Heightman. (*RSGB Bull.*, vol. 25, pp. 253–258; February, 1950.) Technical and economic factors governing the efforts of the set designer and manufacturer to meet the requirements of the average buyer and operator are reviewed. Only AM telephony and telegraphy reception in the range 150 kc to 35 Mc is considered. Emphasis is on amateur requirements. The design of radio-frequency mixer, and intermediate-frequency stages is discussed and certain refinements are recommended.

621.396.621:621.385 1236

A Gated Beam Tube—Adler (*See* 1292.)

621.396.621.22 1237

Marine Communal Aerials—(*Wireless World*, vol. 56, pp. 73–74; February, 1950.) Description of two systems providing interference-free reception on a large number of receivers. Filter circuits in the communal antenna reject the interference due to the ship's transmitters, and in one system feed three basic output stages, which in turn may feed other supplementary units whose number depends on the size of the ship. In the second system the filtered output from the communal antenna is fed via a cathode follower to two wide-band amplifiers, one covering the medium- and long-wave bands and the other the short-wave broadcasting bands. The output from each chain is then distributed, after further amplification, by low-impedance lines throughout the ship.

621.396.664 1238

An Aerial Comparator and Monitor Unit—J. D. Storer and S. Southgate. (*Jour. Brit. IRE* vol. 10, pp. 4–9; January, 1950.) An improved system for antenna selection and signal monitoring at receiving centers. Any available antenna may be applied, via a coaxial switch, to a monitor receiver the intermediate-frequency output of which is displayed on the screen of a cathode-ray tube. Any two antennas may be switched alternately to the monitor receiver, either manually or at high speed by electronic means, to obtain a visual comparison on the cathode-ray tube, the time-base being locked to the antenna switching speed.

621.396.822 1239

On the Spectral Distribution of Energy of Voltage Fluctuations at the Output of a Radio-Noise Receiver—J. Mosnier and J. L. Steinberg. *Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 438–440; January 30, 1950.) There conclusions are drawn from the results of experiments in which the noise spectrum of a receiver was compared with the uniform spectrum of current fluctuations of a saturated diode: (a) there exists in the receiver itself a hitherto unnoticed source of fluctuations which must be looked for in the tubes themselves; (b) the spectrum of these fluctuations has a hyperbolic trend and the amplitude becomes negligible with respect to normal fluctuations only towards 500 cps; (c) a system of periodic commutation at af proposed by Dicke (475 of 1947) must be operated at above 500 cps to eliminate these parasitic fluctuations.

621.396.828 1240

Radio Interference: The Work of the E.R.A. on Suppression of Interference from Electrical Equipment—(*Beama Jour.*, vol. 56, pp. 412–413; December, 1949.) Brief outline of investigations of radio interference from domestic and industrial electrical equipment and from combustion engines. The relevant British Standards publications are listed.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 1241

Theory of Communication—P. Aigrain. (*Ann. Télécommun.*, vol. 4, pp. 406–411; December, 1949.) Statement of the recently developed theory of Shannon (1361 of 1949) showing the relations existing between the bandwidth of a communication channel, the type of signal transmitted, and the signal-to-noise ratios at the input and the output of the receiver used.

621.394.5+621.396.71 1242

The Italcable Radio-Receiving and Submarine-Cable Station at Acilia—A. Niutta. (*Poste e Telecomun.*, vol. 17, pp. 373–386; July, 1949.) General description of a new station, some 24 km from Rome, to replace stations lost during the war. World-wide radio and cable links are provided. The space-diversity system is used for reception, the rhombic antennas simultaneously receiving signals from two opposite directions. To avoid interference, all cable equipment is enclosed in a Faraday cage.

621.395.44:621.315.052.63]:621.395.658.076.7
1243

Voice-Operated Switching of Carrier Systems—R. C. Fox, F. S. Beale, and G. W. Symonds. (*Electronics*, vol. 23, pp. 92–95; February, 1950.) An all-electronic transfer unit provides two-way or party-line communication over power line or radio carrier systems using a single frequency. The equipment functions at a high speed and makes it possible for the speaker to be interrupted by the listener. Oscillograms show negligible speech clipping at the start.

621.395.5:621.315.212 1244

Coxial-Cable Telephony Transmission Systems—L. Albanese and P. Schiaffino. (*Poste e Telecomun.*, vol. 17, pp. 141–167; March, 1949.)

comprehensive survey covering problems of esign peculiar to coaxial-cable telephony, nd the solutions arrived at in current practice r cable dimensions, channel frequency alloca-ion, transmitting equipment, supervisory ear, and power supply. Line-amplifier prob-ems, in particular interstage tube coupling, istortion, and output-impedance matching, re treated in some detail.

21.396.1 **1245**

The Provisional Frequency Board—P. F. iling. (*RCA Rev.*, vol. 10, pp. 600–607; ecember, 1949.) Short account of the func-ions of the Board as laid down at the Atlantic ity conference, and of its work up to the resent.

21.396.5 **1246**

Citizens Radio Range—(*Electronics*, vol. 3, pp. 136, 138; February, 1950.) A short ccount of tests carried out by the Hallicrafters ompany with experimental equipment oper-ted from batteries. The transmitters used grid nodulation of about 30 per cent on peaks, nd λ/2 folded dipoles served as antennas. he maximum ranges for two hand-held trans-eivers varied from 7 miles, both sets being 3 ft above ground level and intervening ter-ain flat and clear, to 0.2 mile with both sets t a height of 6 ft and intervening terrain ntirely wooded. When one dipole antenna was eplaced by a highly directional corner-reflector ype of antenna, the maximum ranges were hree times greater.

Using a mobile unit with a ground-plane ntenna mounted on top of a car and another nit which fed 15-w rf power into a corner-eflector antenna mounted on top of a 75-ft ower, reasonable communication was possible t distances up to 3 miles in a typical residential ection, and up to 9 miles in less-populated reas.

521.396.931 **1247**

V.H.F. Radio Equipment for Mobile Serv-ices—D. H. Hughes. (*Radio Tech. Dig.* (Franç.), vol. 3, pp. 365–368; December, 1949; and vol. 4, pp. 43–49; February, 1950.) French version of 1797 of 1949, with added bibliography.

621.396.931 **1248**

Planning V.H.F. Mobile Systems—E. R. Burroughes. (*Marconi Rev.*, vol. 13, pp. 37–46; January–March, 1950.) Discussion of the advantages obtainable from a correctly planned vhf system, and information on the type of equipment available for mobile services.

621.396.933:621–526 **1249**

Electrical Remote-Control and Indicating Systems in Airborne Radio Equipment—Gamlen. (*See* 1254.)

621.396.97:654.191 **1250**

Economies in the Planning, Design and Operation of a Sound Broadcasting System—R. T. B. Wynn. (*Proc. IEE* (London), Part III, vol. 97, pp. 1–10; January, 1950. Summaries in *Proc. IEE* (London), Part I, vol. 97, pp. 7–8; January, 1950. and *Engineer* (London), vol. 188, pp. 636–640; December 2, 1949.) Inaugural address as Chairman of the Radio Section of the IEE The development of continuity working and program input equipment in studio centers is considered. Automatic moni-tors designed by BBC engineers are described and economies effected by the use of such monitors are discussed. The problem of economic balance between station installation and running costs and freedom from break-down is considered; details are tabulated of capital costs (1943) for the construction of the Skelton sw transmitting station, and also comparative running costs of the Droitwich medium- and long-wave Slant Point medium-wave, and Skelton short-wave stations Un-attended transmitting stations are also dis-cussed and the arrangements at present in use at various BBC stations of this type for automatic frequency adjustment, routine checking, and maintenance are outlined.

SUBSIDIARY APPARATUS

621–526 **1251**

Analysis of the Operation of Discontinuous Physical Systems and Its Application to Servo-mechanisms—F. H. Raymond. (*Ann. Télé-commun.*, vol. 4, pp. 250–256, 307–314, and 347–357; July–October, 1949.)

621–526 **1252**

An Analysis of Relay Servomechanisms—D. A. Kahn. (*Elec. Eng.*, vol. 69, p. 155; February, 1950.) Summary of AIEE 1949 San Francisco meeting paper. The differential equa-tions governing a relay-controlled system are nonlinear and cannot be solved by conventional methods. Application of Laplace-transform theory to such a system yields a response ex-pressed in the form of a series from which the behavior of the system may be determined. The method is applicable to any controlled system with a unilateral transfer function, such as a motor, tube circuit, or a complete servomechanism. Information concerning peri-odic oscillations and stability of such systems can be derived.

621.–526 **1253**

Servomechanisms and Telecontrol—Y. Rocard and J. Loeb. (*Ann. Télécommun.*, vol. 4, pp. 397–404; November, 1949.)

621.–526:621.396.933 **1254**

Electrical Remote-Control and Indicating Systems in Airborne Radio Equipment—D. R. Gamlen. (*Marconi Rev.*, vol. 13, pp. 1–20; January–March, 1950.) Remote-control sys-tems for receivers, transmitters, and radio compasses, which are small and adapted to con-ventional aircraft voltage supplies, are dis-cussed. Devices used in these systems are de-scribed and include a controller switch drive, a preselector unit and ac and dc selsyns, both the self-synchronous and "Motaysynn" ac systems being treated.

621.355.8 **1255**

New Lightweight Accumulator—C. L. Chapman. (*Elec. Rev.* (London), vol. 146, pp. 345–347; February 24, 1950.) A special method of construction, using Ag and Zn electrodes in KOH electrolyte, provides an unspillable, rechargeable battery of $\frac{1}{2}$ the size and $\frac{1}{3}$ the weight of the lead/acid accumu-lator of equivalent capacity. It requires normal maintenance and its characteristics are similar to those of the Ni-Fe cell. Nominal voltage during discharge is 1.5 v. Types to be made available comprise 0.5, 1, 3, 5, 15, 20, and 40 A hr units. A 1-A hr cell only weighs 1 oz. Specifications of various types are tabulated. 40 w hr/lb weight is a standard production achievement. When being charged, only minute quantities of hydrogen are evolved.

621.396.68:621.397.5 **1256**

Metal-Rectifier Voltage Multipliers for E.H.T. [extra-high-tension] Supplies—A. H. B. Walker, (*Jour. Telev. Soc.*, vol. 5, pp. 311–317; September, 1949.) Discussion of the operating principles and characteristics of cascade vol-tage multipliers of the Cockcroft type, and description of a multiplier giving an output of 10 kv and operated from the 350–0–350 v secondary of a center-tapped transformer. Pulse-operated multipliers are also considered briefly.

621.396.682.027.5/.6 **1257**

Variable High-Voltage Power Source—W. S. Ramsey. (*Electronics*, vol. 23, pp. 98–101; February, 1950.) Two separate units for the ranges 5–10 kv and 10–30 kv, with regulation better than 0.05 per cent. In series they will give a current of 2ma at 40 kv with <0.1 per cent ripple voltage.

621.398 **1258**

Radio Synchro-Motor [selsyn]—J. Loeb, J. R. Duthil, and A. Jeuden. (*Ann. Télécom-mun.*, vol. 4, pp. 87–102; March, 1949.) A study of different transmission systems in which angular motion or intelligence is repre-sented by phase modulation.

TELEVISION AND PHOTO-TELEGRAPHY

621.397.331 **1259**

Improved Television Modulator—J. Haughawout. (*Electronics*, vol. 23, pp. 86–88; February, 1950.) A circuit is described, with complete component details, which gives a constant synchronization signal and holds the output black level at a predetermined voltage. When this modulator is used, on switching from one program source to another, the re-ceiver picture fades out and does not return till the resynchronization process is completed. Provision is also made for emergency operation in case the keyed clamping circuit fails.

621.397.331.2 **1260**

A New Image Orthicon—R. B. Janes, R. E. Johnson, and R.R. Handel. (*RCA Rev.*, vol. 10, pp. 586–592; December, 1949.) "The design of a new panchromatic high-sensitivity photo-surface has resulted in the development of a new image orthicon, RCA-5820, which permits the televising of low-level illuminated scenes with a faithful gray-scale rendition of colors. Performance results of this new tube in com-parison with other types for both remote and studio pickup are given."

621.397.331.2 **1261**

Some Considerations on a Scanning Tube for Film—N. Schaetti. (*Le Vide* (Paris), vol. 5, pp. 739–747; January, 1950.) A general dis-cussion of cathode coatings for photoelectric and electron multiplier tubes, and a descrip-tion of a tube designed for 729-line scanning of cinematograph film. This operates on the principle of Farnsworth's image dissector, transmitting 25 frames per second. An Sb/Cs cathode is used. A 1-v signal is obtained, so that additional amplification is not required. Investigations are to be made with a concave cathode and with interlaced scanning.

621.397.331.2 **1262**

Manufacturing Metallized Picture Tubes—E. R. Ewald. (*Electronics*, vol. 23, pp. 78–79; February, 1950.) Details of techniques by which the phosphor coating of a television picture-tube screen is covered by a film on which a thin coating of Al is evaporated, the film being removed by heat prior to the final evacuation.

621.397.5 **1263**

A Six-Mc/s Compatible High-Definition Color Television System—R.C.A. Labora-tories Division. (*RCA Rev.*, vol. 10, pp. 504–524; December, 1949.) Color-picture sampling and time-multiplex transmission together with the "mixed-highs" principle enable the color transmission to be compressed into a 4-Mc band suitable for a total channel assignment of 6 Mc. Each primary color signal passes through a low-pass filter (0–2 Mc) and is then passed to an electronic commutator. The three-color signals are also sent to a band-pass filter (2–4 Mc) the output being the "mixed-high" signal which is added to the signals from the commutator and then passed through a low-pass filter (0–4 Mc). Various receivers and color converters are illustrated and described. The standard scanning speeds are used so that a monochrome picture can be received on existing receivers.

621.397.5 1264

Dot Systems of Color Television—W. Boothroyd. (*Electronics*, vol. 22, pp.88–92; December, 1949; and vol. 23, pp. 96–99; January, 1950.) Short accounts of several sequential sampling and multiplex methods that permit reception either in color or monochrome with the appropriate receiver.

621.397.5 1265

Geometrical Study of the Optimum Number of Channels for Television—Payen. (*Onde Élec.*, vol. 29, pp. 398–401; November, 1949.) Discussion of the problem of providing satisfactory service over extensive territory by means of a number of suitably located stations transmitting the same program on selected frequencies.

621.397.5:535.88 1266

Stereoscopic Television—F. Lachner. (*Radio Tech.* (Vienna), vol. 25, pp. 699–701; December, 1949.) Different methods of stereoscopic viewing and projection are reviewed, particularly those requiring the viewer to wear special eye-pieces. The application of a special cr tube to stereoscopic color television is described. This tube is divided into six chambers for left and right scanning in the three primary colors; a combinaation of prisms and reflectors is used for composition of the colored image.

621.397.5:621.315.212 1267

Television Terminals for Coaxial Systems—L. W. Morrison, Jr. (*Elec. Eng.*, vol. 69, pp. 109–115; February, 1950.) By double modulation the L-1 coaxial-line terminal equipment transforms the video frequencies to frequencies 200 kc higher, which can be transmitted efficiently over the line (transmission band 64–3,100 kc). The original frequencies are restored on the receiver by a double demodulation process. Vestigial-sideband transmission is used. The equipment and its operation are described.

621.397.5:621.396.68 1268

Metal-Rectifier Voltage Multipliers for E.H.T. [extra-high-tension] Supplies—Walker. (*See* 1256.)

621.397.5:621.396.813 1269

Artificial Lines for Video Distribution and Delay—A. H. Turner. (*RCA Rev.*, vol. 10, pp. 477–489; December, 1949.) Delay distortion may be reduced by the use of mutual inductance between adjacent half-sections of a low-pass filter. A combination of such T sections and bridged-T sections gives small delay distortion up to 0.85 of the cut-off frequency. The quality of picture signals transmitted through several experimental lines of 20 sections each confirmed this theory. Each section provides a low-impedance feed point and the branch lines from these points must behave as lumped capacitors; they must therefore be unterminated and short compared with the signal wavelength. Input and output voltages are sketched for various values of mutual coupling. The work of other investigators is briefly reviewed and 14 relevant publications are noted.

621.397.61 1270

Low-Power Television Transmitter—L. Voorhees. (*Elec. Eng.*, vol. 69, pp. 151–154; February, 1950.) Description, with block diagrams, of equipment designed for local coverage. The sound transmitter has an output of 250 w; that of the video transmitter is 500 w. Separate transmitters are used for the low and the high television frequency bands.

621.397.62 1271

The Simplification of Television Receivers—W. B. Whalley. (*Sylvania Technologists*, vol. 3, pp. 9–12; January, 1950.) An investigation of general means of simplifying television receivers so as to reduce the costs of production and servicing. The operation of a television receiver is analyzed from the standpoint of the minimum number of necessary functions, and then the simplest possible circuit to perform each function is planned, taking into consideration the possibility of combining two functions in the same circuit.

621.397.62 1272

Fixed-Tuned Broad-Band Television Booster—A. Newton. (*Electronics*, vol. 23, pp 116 . . . 134; February, 1950.) The booster is intended to improve the over-all noise figure and raise the useful gain of a receiver. The first rf stage, a grounded-grid triode amplifier, is coupled to the second stage by a double-tuned inductively coupled circuit, the output of which is fed to the receiver through a 300 Ω input resistance. Separate high-band and low-band amplifiers are used and the respective inputs and outputs are connected through a crossover network consisting of λ/4 sections. Total gain is about 8.

621.397.62:621.396.67 1273

Trends in TV Receiver Antenna Design—Kamen. (*See* 1085.)

621.397.62.001.4 1274

Measurement of Transient Response of TV Receivers—Van Duyne. (*See* 1194.)

621.397.62.029.63:621.396.622.63:546.289 1275

Germanium Diodes for U.H.F. TV—J. Lingel. (*See* 1307.)

621.397.822 1276

Perception of Television Random Noise—P. Mertz. (*Jour. Soc. Mot. Pic. Eng.*, vol. 54, pp. 8–34; January, 1950.) The problem is studied by analogy with graininess in a photographic image. Effective random noise power is obtained by cumulating and weighting actual noise powers over the video frequencies with a weighting function diminishing with increasing frequency. Values obtained check reasonably well with preliminary experiments. The effect of changing the tone rendering and contrast of the television image is analyzed.

621.397.828 1277

Television Interference Seldom comes from Power Systems—F. L. Greene. (*Elec. World*, vol. 133, pp. 55–59 . . . 128; January 16, 1950.) The principal causes of interference with television reception so far encountered are listed and their effects on the television screen illustrated. Methods of eliminating the interference are suggested. Ignition systems cause most interference; domestic and industrial apparatus are less troublesome. Interference from power lines is slight.

621.397.828 1278

Some Devices for Reducing the Effects of Fading and Interference: Part 1—D. McMullan (*Jour. Telev. Soc.*, vol. 5, pp. 318–328; September, 1949.) The various types of interference encountered in the reception of television signals are reviewed and methods of reducing such interference are discussed, with special reference to circuit techniques which may be used in the design of the receiver. Future problems are considered, and the respective merits of positive and negative modulation are discussed from the point of view of their susceptibility to interference.

621.397.828 1279

Suppression of TVI—F. T. Wilson. (*Short Wave Mag.*, vol. 7, pp. 740–745 and 828–832; December, 1949 and January, 1950.) A detailed account of an investigation of the source of television interference in an amateur transmitter and of its elimination by means of low-pass filters, effective screening, and suitable modification of the power amplifier. A circuit diagram, with component values, is given of a push-pull amplifier including harmonic suppression arrangements. Final tests of the modified transmitter when radiating 100 W on a frequency of 14 Mc showed only a trace of interference with a television receiver only 2 ft away from the transmitter.

Reception tests on local amateur stations, with the receiver tuned to frequencies of the order of 42–43 Mc, showed that the strongest harmonics were received from transmitters using single-ended output stages with beam tetrodes, while transmitters using variable-frequency oscillators had very weak harmonics.

TRANSMISSION

621.316.726 1280

Carrier Frequency Control: Automatic System for Unattended Transmitting Stations—J. C. Gallagher. (*BBC Quart.*, vol. 4, pp. 249–256; Winter, 1949–50.) The Droitwich standard-frequency transmission of 200 kc is used to effect the near synchronization of unattended transmitters. Frequency dividers and pulse-shaping circuits are used to generate a train of 0.6-μs pulses with a recurrence frequency of 1,000 per sec, which are fed together with the locally generated carrier into two integrating circuits; from these pulses are derived, each corresponding to a frequency change of ±2 parts in 10^8. The pulses drive an impulse motor coupled to a series capacitor in the crystal-oscillator circuit.

621.396:621.394.611.2 1281

An Electronic Keyer—B. Bröndum-Nielsen. (*RSGB Bull.*, vol. 25, pp. 259–262; February, 1950.) Description of an easily constructed semi-automatic keying unit. Dash/dot and mark/space ratios can be adjusted to be correct for any sending speed within the range of the unit. The resulting Morse signals compare favorably with those of an automatic tape transmitter.

621.393.61 1282

Aircraft Communications Transmitter Type AD.107—W. R. Bitcheno. (*Marconi Rev.*, vol. 13, pp. 21–36; January–March, 1950.) Transmitters are described for hf and mf, each comprising four units: driver and modulator, amplifier, antenna, and power units. The hf transmitter, weighing 85 lb, with a maximum power of 150 W, operates on 10 crystal-controlled spot frequencies in the range 2–18.5 Mc. The 80-lb mf transmitter, with maximum output of 120 W in the range 320 to 520 kc, has identical controls and the same number of crystal-controlled frequencies as the hf equipment.

621.396.61 1283

Band-Switch QRO Transmitter—J. N. Walker. (*Short Wave Mag.*, vol. 7, pp. 818–827; January, 1950.) Full circuit and construction details of a 3-stage transmitter for full-power operation in the range 3.5–28 Mc. Switched coil turrets are used for both driver and power-amplifier. Only a keyed vfo is needed for cw working. The set is adaptable for either screen or anode-and-screen modulation of the power amplifier for telephony. Standard British components are used throughout.

621.396.615 1284

3-kW M.F. Transmitter Design—I. F.

Deise and L. W. Gregory. (*Communications*, vol. 29, pp. 12–14, 35, 12–13, and 30–31; October–December, 1949.) In order to reduce the size of the transmitter, hf iron-cored transformers are used; the construction and the operational characteristics of these are described. The cores are of hipersil iron. An autotransformer with 5 per cent tappings is used for matching the reflected antenna impedance to that of the power amplifier. The working frequency can be anywhere in the range 250–540 kc and arrangements are provided for immediate switching, when desired, to a second frequency in the same range.

621.396.619.15 1285

Square-Wave Keying of Oscillators—J. C. Seddon. (*Electronics*, vol. 23, pp. 162. . . 172; February, 1950.) A low-power circuit is described which makes possible square-wave grid-modulation of oscillators over a wide range of pulse widths and duty cycles. 750-v positive pulses can be produced from 15-v negative pulses with a maximum current of 15 ma, the rise and decay times being $<1\mu s$. 15-w average power will easily control an oscillator giving 7.5 kw peak power output.

TUBES AND THERMIONICS

537.122:531.112:536.525.92 1286

Electron Transit Time in Space-Charge-Limited Current between Coaxial Cylinders—P. L. Copeland and D. N. Eggenberger. (*Jour. Appl. Phys.*, vol. 20, pp. 1148–1151; December, 1949.) The transit time of electrons between a cylindrical cathode and an external concentric anode can be determined if the potential distribution between the cylinders is known. Taylor series expansions are developed for this in a form which enables the transit times to be readily calculated to within 0.1 per cent.

537.58:539.234 1287

Thermionic Emission of Thin Films of Alkaline Earth Oxide Deposited by Evaporation—Moore and Allison. (*See* 1165.)

621.383 1288

Design Features of a New Photocell—J. H. Crow and V. C. Rideout. (*Radio and Telev. News, Radio-Electronic Eng. Supplement*, vol. 14, pp. 8–9, 28; February, 1950.) The photocell Type CE-70V is a high-vacuum, end-on type of tube, with two ring anodes and a flat disk-type cathode, which may be used as a combination modulator and photocell at carrier frequencies up to at least 200 kc. The inner ring is the control anode and serves to vary the emission current reaching the load anode. Light striking the cathode controls the emission. Its application in an electro-optical pyrometer is described, where the comparison of the light intensities at two wavelengths is used to determine the absolute temperature of a luminous flame.

621.385 1289

New Valves—J. Steiger. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 14, pp. 112–121; February, 1950. In French.) An illustrated review of the most recent thermionic tubes, camera tubes, magnetrons, memory tubes, electron multipliers, electron couplers, and of the "selectron" developed in the RCA laboratories.

621.385 1290

New Subminiature Valves—C. G. Gee (*Wireless World*, vol. 56, pp. 46–47; February, 1950.) These are flat types with filament currents of only 15ma. Performance is comparable with that of the corresponding Mullard 10-mm tubes (2410 of 1948), but anode and filament currents are both much lower and the reduced size and flat shape results in a saving of space of about one-third compared with the 10-mm tubes, when three tubes are fitted in a hearing aid. The 10-mm range is also being extended for both mains and battery operation, for use in miniature equipment.

621.385:621.396.621 1291

A Survey of Modern Radio Valves: Part 3—Receiving Valves for Use Below 30 Mc/s—K. D. Bomford. (*P.O Elec. Eng. Jour.*, vol. 42, pp. 201–208; January, 1950.) The modes of operation of the more widely used types of receiving tube are discussed and contrasted and the various factors limiting their performance are considered. Data are included on the physical form and the characteristics of modern tubes and the operating conditions conducive to long life are discussed. Parts 1 and 2: 482 and 483 of March.

621.385:621.396.621 1292

A Gated Beam Tube—R. Adler. (*Electronics*, vol. 23, pp. 82–85; February, 1950.) This tube, Type 6BN6, uses a sharply-focused electron beam passing through two control grids, each of which has an unusually steep and linear transfer characteristic. It is suitable for use in FM discriminator circuits, as a synchronization-pulse separator, or as a square-wave generator. The assembly fits into a miniature-tube envelope with a 7-pin base.

621.385:621.396.621 1293

Contour Analysis of Mixer Valves—N. E. Goddard. (*Wireless Eng.*, vol. 26, pp. 350–356; November, 1949.) "The characteristics of a mixer tube are completely defined by a series of three-dimensional surfaces. Two of the co-ordinates, grid voltage, and heterodyne-oscillator voltage, determine the operating conditions for small signal voltages. The third co-ordinate is one of a number of tube parameters: fundamental or harmonic cónversion conductance, cathode current, or grid current. Each surface is described by a contour map on which load lines are drawn for several automatic-bias circuits. The method is illustrated by Fourier analysis of a theoretical mutual-conductance curve and by experimental measurements on an EF42 pentode.

621.385.029.63/.64+621.396.615.14 1294

The Amplification of Centimetre Waves: Travelling-Wave Valves—G. Goudet. (*Onde Élec.*, vol. 29, pp. 8–12; January, 1950.) Some typical performance figures are given for klystrons and triodes at 3,000 Mc and the development of the traveling-wave tube in different laboratories is reviewed. Details are given of the design and construction of the tubes produced in the Laboratoire Central de Télécommunications. Their mean operating frequency is 2,600 Mc; pass band, 400 Mc; output power, 50 mw; noise factor, 18 db; gain, 35 db for a useful current of 3 ma. The bibliography includes 42 references.

621.396.615.14+621.385.029.63/.64 1295

Valves for Communication on Frequencies above 1000 Mc/s: Part 2—H. Schnitger. (*Fernmeldetech. Z.*, vol. 3, pp. 13–22; January, 1950.) An illustrated description and comparison of three types of microwave amplifiers: (a) the disk-seal triode; (b) the klystron; and (c) the traveling-wave tube. The disk-seal triode is very useful for wavelengths near 10 cm, where its noise factor is about 8, rising to about 60 for 3-cm wavelength. The noise factor of the traveling-wave tube is appreciably higher and the klystron has both higher noise factor and relatively small bandwidth. Part 1: 2105 to 1949.

621.385.029.63/.64 1296

The Traveling-Wave Tube (Discussion of Waves for Large Amplitudes)—L. Brillouin. (*Jour. Appl. Phys.*, vol. 20, pp. 1196–1206; December, 1949.) Equations are developed for the interaction of an electron beam with an em wave in a simplified linear model with infinitely short sections. Hence the usual theory for waves of small amplitude, assuming a strong beam with a weak signal, is derived and the precise limits of the validity of this theory are defined. For waves of large amplitude, the amplified wave is progressively distorted until a final stage is reached, dependent essentially upon the details of the tube structure, when no further amplification is possible. In some cases, a type of shock wave results, with a complete bunching of the space charge. A similar solution applies also to linear accelerators and to synchrotons.

621.385.029.63/.64 1297

Travelling-Wave Valve T.P.O.85—(*Ann. Radioélec.*, vol. 5, pp. 62–63; January, 1950.) A linear amplifier tube made by the Compagnie Générale de TSF for operation between 1,500 and 5,000 Mc. Max. output power, 1 w; gain, 15–19 db; pass band, 80–100 Mc; total length, about 460 mm, max. diameter, about 41 mm.

621.385.029.63/.64 1298

Recent Developments in Traveling-Wave Tubes—L. M. Field. (*Electronics*, vol. 23, pp. 100–104; January, 1950.)

621.385.029.63/.64:621.396.822 1299

Transit-Time Effects in U.H.F. Valves—A. H. Beck and J. Thomson. (*Wireless Eng.*, vol. 26, pp. 379–380; November, 1949.) Comment on 3312 of 1949 (Thomson) and Thomson's reply.

621.385.032.213 1300

Activation Phenomena with Thoria Cathodes—O. A. Weinreich. (*Jour. Appl. Phys.*, vol. 20, p. 1256; December, 1949.) An investigation has been carried out into the activation of thoria-coated tungsten cathodes by (a) reverse electron current, and (b) exposure to evaporation products from a nearby operating thoria cathode. The highest emission at temperatures at which thermal activation is negligible is given by the reverse electron current method. In some cases the emission passes through a maximum when plotted against the time during which reverse electron current is drawn. Procedure (b) was used for the activation of a pulsed thoria cathode, from which high emission could be maintained if a small dc emission was drawn from an auxiliary thoria cathode to the common anode.

621.385.032.213.2:621.317.336.1 1301

Change of Mutual Conductance with Frequency—W. Raudorf. (*Wireless Eng.*, vol. 26, pp. 331–337; October, 1949.) This phenomenon occurs in receiving tubes with indirectly-heated cathodes after operation for about 1,000 hours. It is "due to the deterioration, during operation, of the contact between the oxide coating and the metal sleeve which forms the core of the cathode. It can be greatly reduced by giving the metal sleeve an appropriate shape. By applying Holm's theory of electric contacts it is possible to estimate the number and size of the contact spots (a-spots) between coating and core metal. For an average size of 3μ for the oxide grain, for instance, and a contact resistance of about 40Ω per cm^2 of the cathode surface, the number of a-spots is about 3×10^4 per cm^2 and their average diameter about 1λ. The contact resistance and the liability to cathode sparking seem to be connected. In general, cathode sparking is unlikely up to a specific emission of 1 a/cm^2. To prevent sparking at pulse currents of 10 a/cm^2 of <1 ms duration, the contact resistance between coating and core must be $<10\Omega/\text{cm}^2$."

621.385.032.216 1302

Distribution of Potential in the Coating of an Oxide Cathode during a Pulse of Great Current Density—R. Loosjes and H. J. Vink. (*Le Vide* (Paris), vol. 5, pp. 731–738; January, 1950.) The distribution was investigated by applying the method previously described (2414 of 1948) to specially constructed diodes having up to 3 metal probes inserted in the cathode coating. No barrier layer at the metal/oxide interface was found, but it was established that a great part of the potential drop occurs close to the emitting surface.

621.385.032.216 1303

Comparison between the Electronic and the Thermodynamic Temperature of Oxide Cathodes—R. Champeix. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 64–65; January 2, 1950.) From results of experimental tests on some ten tubes whose thermodynamic temperature Θ was varied between 850° and 1,200°K, the following conclusions are drawn: (a) the electronic temperature T is from 2 per cent to 40 per cent greater than Θ; (b) the difference decreases with increasing temperature; (c) the difference is about halved if a plate electrode is used instead of a grid; and (d) within the above temperature range $T \approx a\Theta + b$, where $a > 1$. A theoretical explanation of the results will be given later.

621.385.032.216:537.583 1304

Thermionic Emission from Oxide Cathodes: Retarding and Accelerating Fields—C. S. Hung. (*Jour. Appl. Phys.*, vol. 21, pp. 37–44; January, 1950.) The effects of applied fields ranging from retarding fields up to accelerating fields of 50kv/cm were studied. For strong retarding fields, excellent agreement with theory was obtained. Near zero field, cathode inhomogeneity or patch effect may be responsible for the deviations found; with the large cathode used, the existence or non-existence of a reflection effect could not be determined. For accelerating fields, deviations of the results from those predicted by the Schottky mirror-image theory are explainable by patch effect and by field intensification at sharp points of the rough cathode surface.

621.385.2:621.315.59 1305

Backward Current of Germanium Diodes—P. Aigrain. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 62–63; January 2, 1950.) Theory is given which leads to values of backward current in agreement with experimental values.

621.385.2:621.315.59 1306

Backward Current and Capacitance of Germanium Diodes—P. Aigrain. (*Compt. Rend. Aacd. Sci.* (Paris), vol. 230, pp. 194–196; January, 1950.) An experimental verification of the theory (see 1305 above) for the extreme cases of low and high back resistance, represented respectively by a transistor with the emitter disconnected and a Type-1N54 diode. Calculated and observed values of current for different voltages are tabulated. The calculated lf shunt capacitance for the transistor is 0.46 pF at 10 V; for the Type-1N54 diode, 0.1 pF.

621.385.2:621.315.59:621.397.62.029.63 1307

Germanium Diodes for U.H.F. TV—F. J. Lingel. (*TV Eng.*, vol. 1, pp. 12–13, 39; January, 1950.) The construction and application of mixer diodes for frequency conversion in the 475–890-Mc television band. The diodes can withstand microsecond pulses of 500 ma in the forward direction and 1 ma in the backward direction. The contact point is welded to the Ge pellet.

621.385.2:621.396.822 1308

Valve Noise and Electron Transit Time—D. A. Bell. (*Wireless Eng.* vol. 26, p. 379; November, 1949.) Comment on 2097 of 1949 (Fraser).

621.385.2/.3].029.64 1309

Electronic Admittances of Parallel-Plane Electron Tubes at 4000 Megacycles—S. D. Robertson. (*Bell Sys. Tech. Jour.*, vol. 28, pp. 619–646; October, 1949.) The general features of the mechanism of electron transit in close-spaced diodes are briefly reviewed. Measurements were made of the electronic admittance of close-spaced parallel-plane diodes and BTL1553 triodes (2964 of 1949) at 4,060 Mc, rf power being fed to the tube from a waveguide source through a waveguide-cavity transformer. The theory and practice of the method are described in detail. The diode conductance is found to be much greater than the lf value, and increases with decreasing electrode spacing. The susceptance decreases with increasing current. For the triode, the input short-circuit admittance departs considerably from values predicted by single-velocity theory, but the transadmittance is only slightly lower than the lf value. See also 1315 below.

621.385.3:621.315.59 1310

Characteristics of Transistors—P. Aigrain and C. Dugas. (*Compt. Rend. Acad. Sci.* (Paris), vol. 230, pp. 377–378; January 23, 1950.)

621.385.3:621.315.59 1311

Physical Interpretation of Type A Transistor Characteristics—L. P. Hunter. (*Phys. Rev.*, vol. 77, pp. 558–559; February 15, 1950.) The relative efficiencies of the processes taking place in a transistor are discussed. For saturation conditions it appears that the poor performance of some transistors is due to low hole-injection efficiency. Below saturation the emitter may remove some electrons from the crystal, or the collector voltage may be insufficient for all the holes to reach the collector; either effect would cause a reduction in efficiency.

621.385.3:621.315.59 1312

The Germanium Crystal Triode—H. Heins. (*Sylvania Technologist*, vol. 3, pp. 13–18; January, 1950.) The electrical properties of semiconductors are briefly discussed. The construction and characteristics of the crystal triode are then described and a few typical circuits are shown.

621.385.3.011.4 1313

Interelectrode Capacitance of Valves—E. E. Zepler. (*Wireless Eng.*, vol. 26, pp. 378–379; November, 1949.) Further discussion. See 2101 (Zepler and Hekner), 2102 (Humphrey and James), and 2962 (Booth) of 1949.

621.385.3.029.63 1314

Two Triodes for Reception of Decimetric Waves—K. Rodenhuis. (*Philips Tech. Rev.*, vol. 11, pp. 79–89; September, 1949.) A description of two receiving tubes for frequencies higher than 300 Mc: a triode, Type EC80, for amplifying and mixing at frequencies up to about 600 Mc, and an oscillator, Type EC81, with an upper frequency limit of about 1,500 Mc. The nine copper-plated pins sealed through the glass base have a very low hf resistance. In contrast to the disk-seal tubes commonly used for uhf operation, these tubes are similar in appearance to conventional types and enable receivers to be built which are simple in con- values measured at low frequency. See also 1309 above.

621.385.3.029.64 1315

Passive Four-Pole Admittances of Microwave Triodes—S. D. Robertson. (*Bell Sys. Tech. Jour.*, vol. 28, pp. 647–655; October, 1949.) Measurements at 4,060 Mc for a wide range of cathode/grid and grid/anode spacings are described. Two grids were used: (a) a parallel-wire grid of 0.3 mil tungsten wire wound at 1,000 turns per inch; and (b) a criss-cross grid of the same wire wound at 550 turns per inch. The microwave transadmittances were found to be much higher than the values measured at low frequency. See also 1309 above.

MISCELLANEOUS

43–3=2 1316

Technisches Worterbuch (Deutsch-Englisch) [Book Review]—R. Ernst. Publishers: Tauchnitz Edition, Hamburg. vol. 1, 612 pp., 16.50 DM. (*Elektrotechnik* (Berlin), vol. 3, p. iv; May, 1949.) "The problem of including as many technical words as possible in a handy volume has been solved surprisingly well. Space economies in the case of related words and those common to the two languages might have been extended to include other less familiar equivalents, such as geschwindigkeitsmodulierte Röhre = klystron. More verbs are included than is usual in most technical dictionaries, and also the principal expressions for mathematical operations."

45–3=2:621.396 1317

Dizionario Tecnico della Radio. Italiano Inglese. Inglese Italiano [Book Review]—L. Bassetti. Publishers: Il Rostro, Milan, 275 pp., 900 lire. (*Radio* (Turin), no. 8, pp. 5–6; November, 1949.) Radio terms and terms in physics and electronics connected with radio are explained. A list of abbreviations and symbols for circuit diagrams is included.

621.396 1318

Radio Engineering: Vol. 2 [Book Review]—E. K. Sandeman. Publishers: Chapman and Hall, Ltd., London, 579 pp., 40s. (*Wireless Eng.*, vol. 26, pp. 412–413; December, 1949; *Wireless World*, vol. 56, p. 10; January, 1950.) The subject matter of this second half of the work includes interference and noise, receivers, measuring equipment, equalizer design, feedback, network theory, and filters. Appendices contain a large collection of formulas and information on a variety of subjects. Vol. 1 was noted in 600 of 1949.

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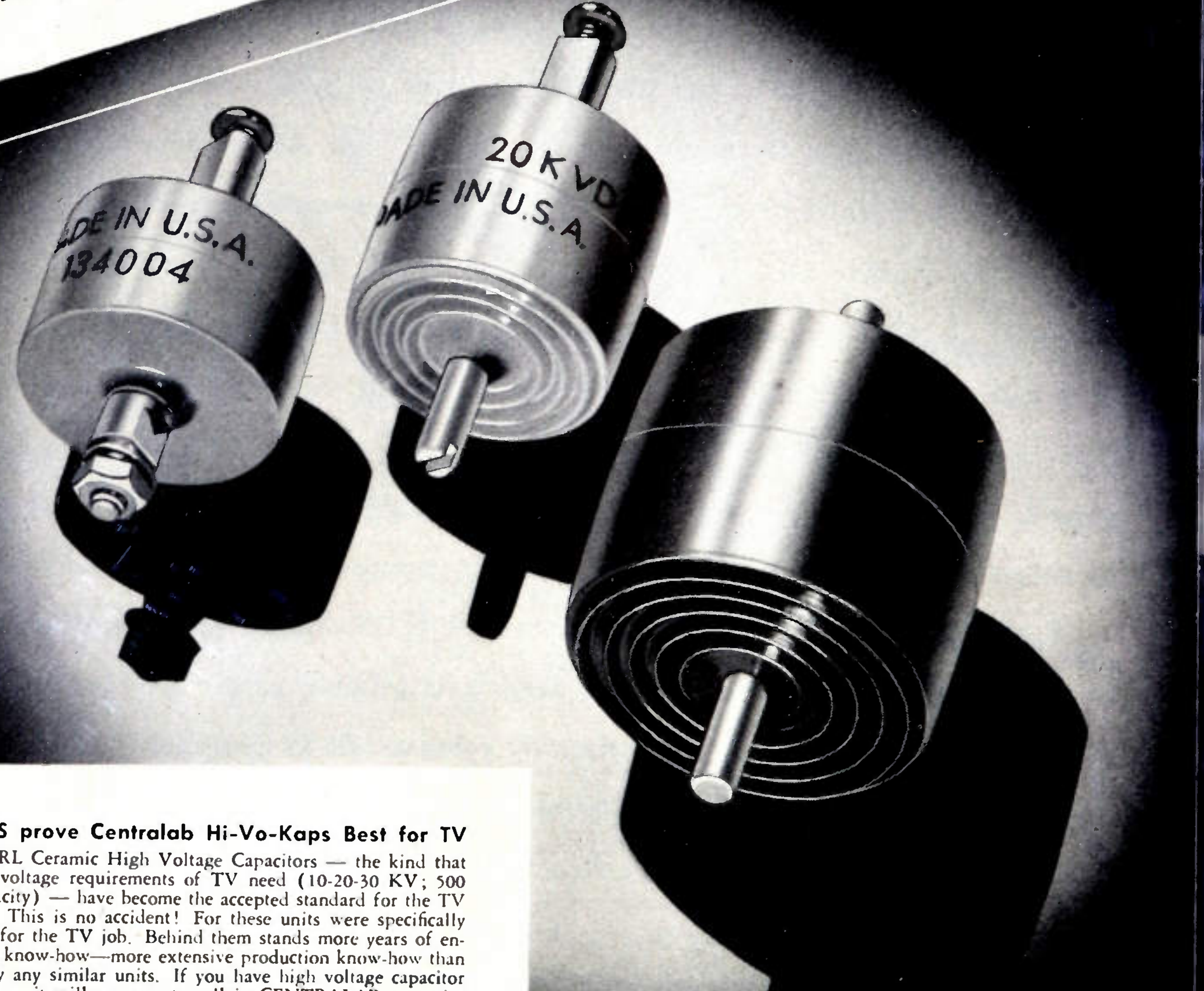

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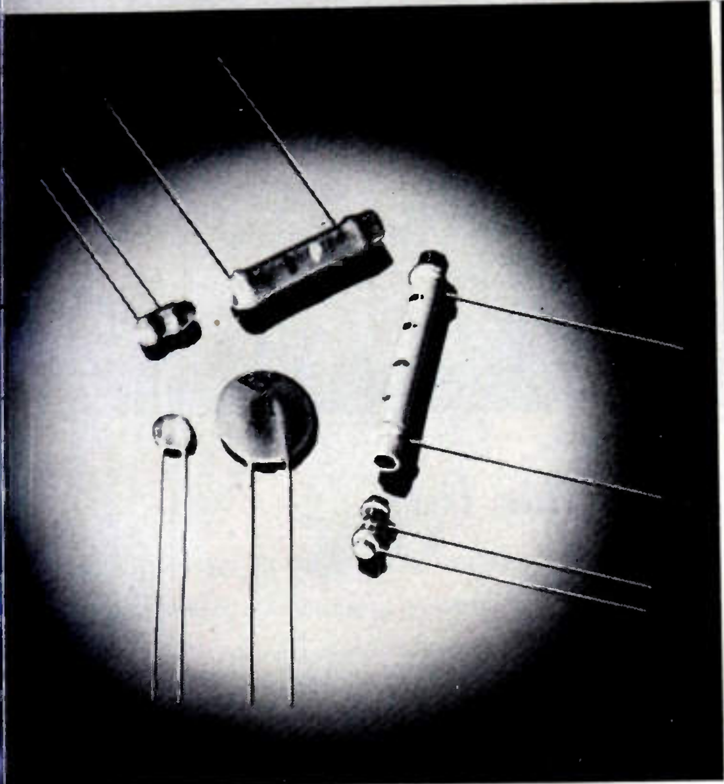

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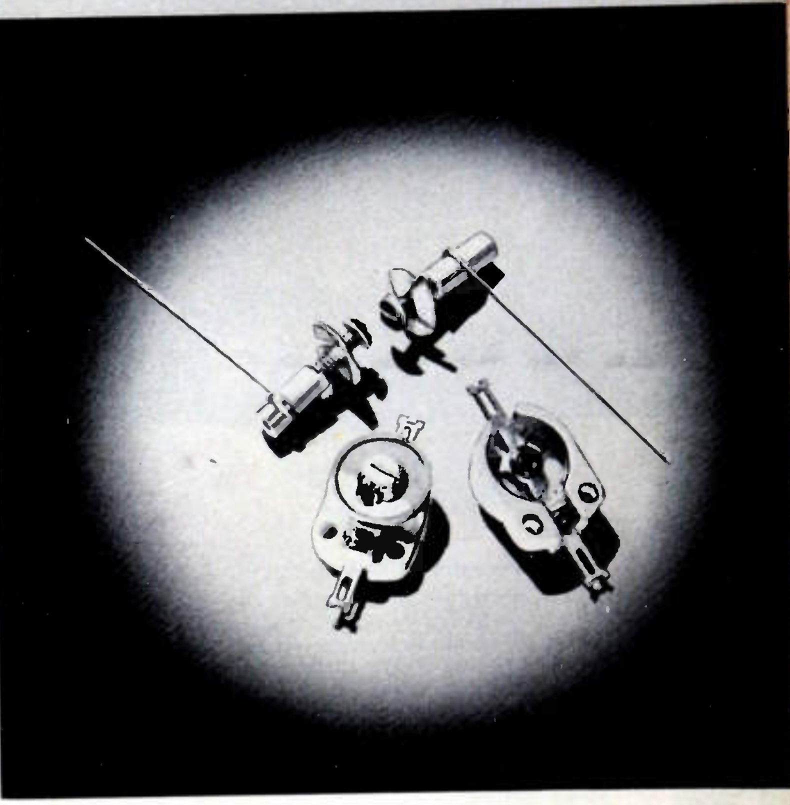

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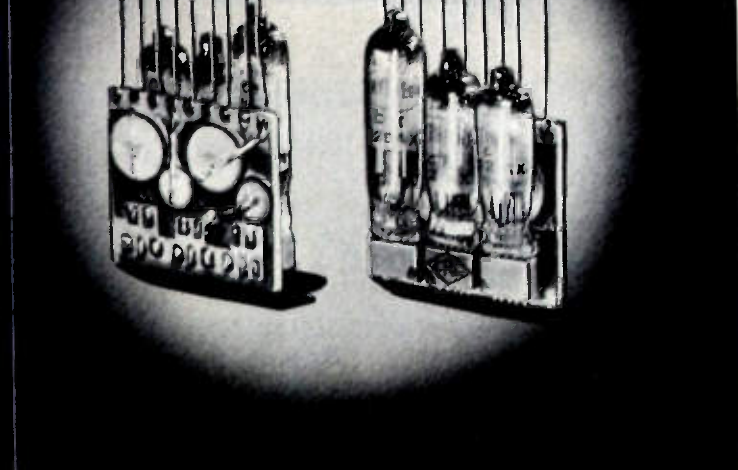

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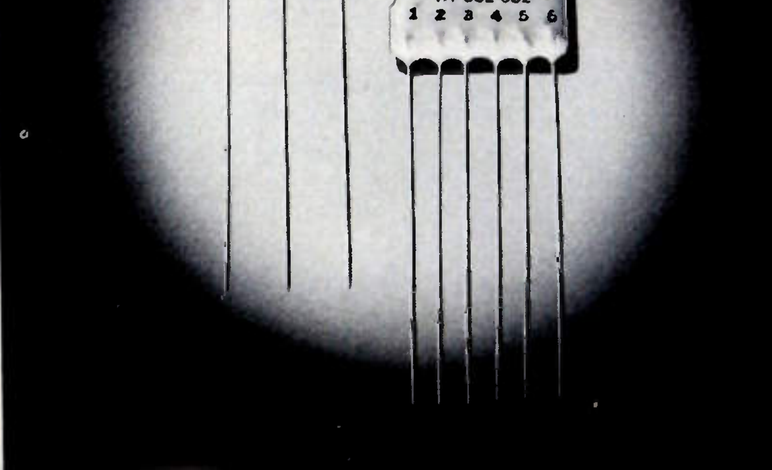

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GL-8008	5 v	7.5 amp	10,000 v	5 amp	1.25 amp
GL-673	5 v	10 amp	15,000 v	6 amp	1.5 amp
GL-869-B	5 v	19 amp	20,000 v	10 amp	2.5 amp
GL-857-B	5 v	30 amp	22,000 v	40 amp	10 amp

SECTION MEETINGS

Akron

"Today's Television Receivers," by Karl Wendt, Colonial Radio; October 18, 1949.

"Color Television," by T. T. Goldsmith, Allen B. DuMont Laboratories, Inc.; November 17, 1949.

"Semiconductor Amplifiers and Applications of Semiconductor Crystals," by L. P. Hunter, Westinghouse Research Laboratories; December 13, 1949.

"Distributed Amplifiers," by William Hewlett Hewlett-Packard Company; March 14, 1950.

Atlanta

"Velocity-Modulated Television," by M. A. Honnell, Faculty, Georgia Institute of Technology; March 31, 1950.

Baltimore

"Radar Target Simulating Equipment," by T. G. Arnold, Jr., Faculty, The Johns Hopkins University; April 18, 1950.

Beaumont-Port Arthur

"Protective Relays and Associated Apparatus on a Power System," by R. O. Hopkins, Gulf States Utilities Company; March 31, 1950.

Cedar Rapids

"Measurement of Nonlinearity in Audio Systems," by W. R. Hewlett, Hewlett-Packard Company; March 16, 1950.

Chicago

"Intercity Networks for Television," by M. E. Strieby, American Telephone and Telegraph Company; Symposium: "New Electronic Instruments and Their Measuring"; March 17, 1950.

Cincinnati

"Computers for Aeronautical Navigation," by O. H. Schuck, Minneapolis-Honeywell Company; February 21, 1950.

"Problems Encountered in the Analytical Counting of Radioactivity," by P. E. Ohmart, Mound Laboratory; Films: "Atomic Explosions at Hiroshima," and "Nagasaki"; March 21, 1950.

"The Bell System Coaxial Cable Television Network," by M. E. Strieby, American Telephone and Telegraph Company; March 30, 1950.

Connecticut Valley

"The Navy Prepares for Undersea Warfare," by J. M. Ide, United States Navy Underwater Sound Laboratory; March 23, 1950.

"Microwave and Acoustic Lenses," by W. E. Kock, Bell Telephone Laboratories; April 13, 1950.

Business Meeting; April 15, 1950.

Dallas-Fort Worth

"The Electrophrenic Respirator," by F. W. Geisert, "Industrial X-Ray," by F. H. Harrison, "Guided Missiles," by B. C. Scammel, and "Cathodic Protection for Pipe Lines," by B. J. Whitley, Students, Southern Methodist University; March 20, 1950.

"Bandwidth Measurements at Locations Remote from Transmitters Operating in Normal Service," by N. A. Hallenstein, Federal Communications Commission; April 5, 1950.

Dayton

"Intercity TV Networks," by M. E. Strieby, American Telephone and Telegraph Company; April 3, 1950.

Denver

Symposium: "Long-Distance Circuits by Microwaves," by L. N. Shindel, R. G. Hall, and L. T. Hearson, American Telephone and Telegraph Company; March 7, 1950.

(Continued on page 40A)

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SECTION MEETINGS

(Continued from page 38A)

"Highlights of the 1950 Institute of Radio Engineers National Convention," by P. M. Lahue, Heiland Research Corporation, G. T. Morrissey, Radio Station KFEL, and E. W. Post, United Air Lines; April 14, 1950.

Des Moines-Ames

"TV Receiver Design," by E. A. Shore, Genera Electric Company; Panel Discussion; February 27, 1950.

"TV Transmission Problems," by L. L. Lewis, Radio Station WOI-TV; Tour of Radio Station WOI-TV; March 27, 1950.

Detroit

"Cosmic Rays," by W. A. Nierenberg, Faculty, University of Michigan; March 17, 1950.

Emporium

"Magnetic Amplifiers," by Frank Shepard, Shepard Laboratories; March 21, 1950.

Evansville-Owensboro

Election of Officers; March 21, 1950.

"Design Trends of Home Television Receivers," by J. D. Reid, Crosley Division of Avco Manufacturing Corporation; April 5, 1950.

Fort Wayne

"Theory and Application of Germanium Semiconductors," by Ralph Bray, Faculty, Purdue University; March 22, 1950.

Hawaii

"VHF Omnidirectional Range," by R. F. Bowker, Civil Aeronautics Administration; March 15 1950.

General Discussion by Lee de Forest, American Television, Inc.; April 12, 1950.

Indianapolis

"Bell System—Microwave and Coaxial Cable Development—Television Network," by H. S. Osborne, American Telephone and Telegraph Company; March 14, 1950.

"Some Selected Problems of Television Receiver Design," by Kurt Schlesinger, Motorola, Inc; March 31, 1950.

London

Film: "Atomic Physics," presented by Flight-Lieutenant Fitzgerald, Royal Canadian Air Force Radio School, and discussed by R. C. Dearle, Faculty, University of Western Ontario; "Program Management," by M. T. Brown, Radio Station CFPL; March 13, 1950.

Los Angeles

"Applications of Analogue Computers to Industrial Research Problems," by G. D. McCann, Faculty, California Institute of Technology; "Radio Interference and its Elimination," by Fred Nichols, Airesearch Manufacturing Company; and "Sales Engineering," by G. F. Rucker; April 4, 1950.

Louisville

"Universal Phonograph Styli," by J. D. Reid, The Crosley Division of Avco Manufacturing Company; February 10, 1950.

"Atomic Energy," by D. M. Bennett, Faculty, University of Louisville; March 10, 1950.

Milwaukee

"Geiger Counters," by D. E. Collins, Victoreen Instrument Company; January 11, 1950.

"Tone Control Circuits," by D. E. Mereen, Western Sound and Electric Company; January 31, 1950.

"Characteristics and Application of Magnetic Amplifiers," by W. J. Dornhoefer, Vickers Electric Division; February 22, 1950.

(Continued on page 42A)

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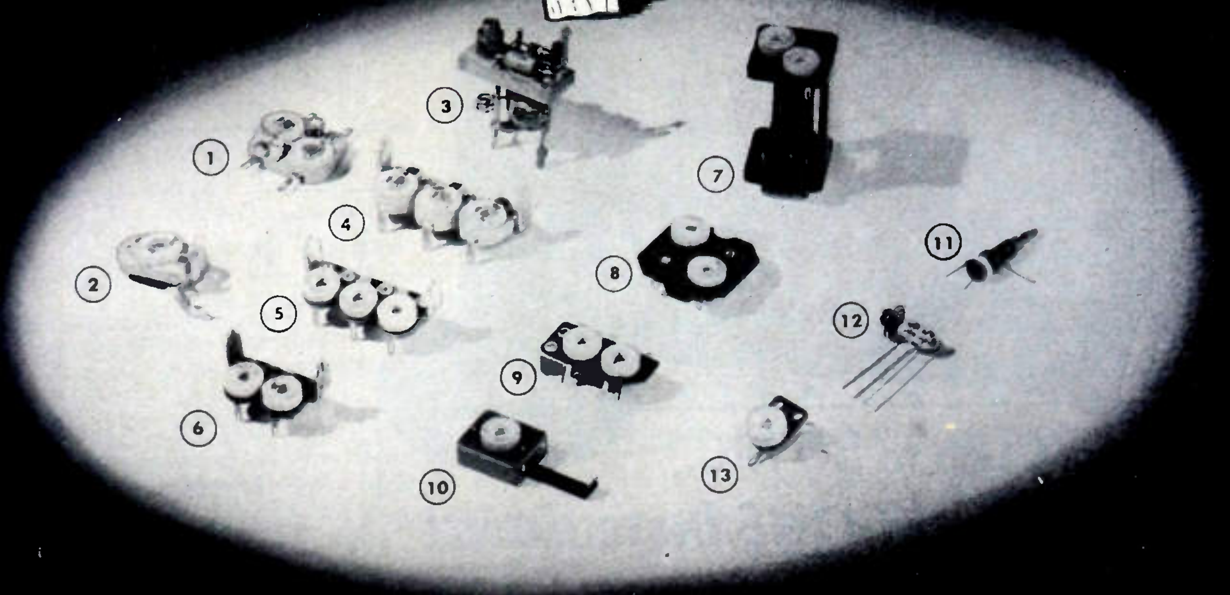

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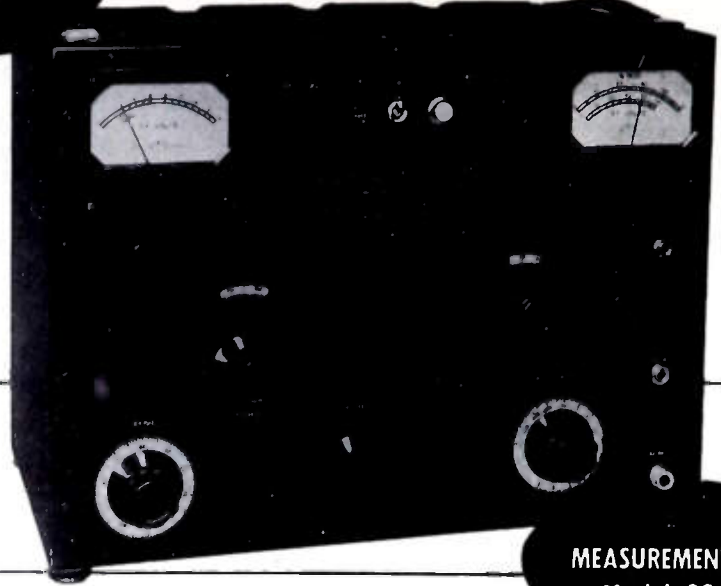

SECTION MEETINGS

(Continued from page 40A)

"Something New in Audio Amplifiers," by F. L. McIntosh, McIntosh Engineering Laboratories; March 15, 1950.

MONTREAL

"Electronic Digital Computers," by C. C. Gotlieb, Faculty, Computation Centre; February 15, 1950.

"Interlocked Radio Intercommunication," by G. S. Jewell, Queen's University; "Nouvel Indicateur Electronique de Pressions Dynamiques," by Rene Fortier, Ecole Polytechnique; and "Design of a Reflexed Speaker Cabinet with Sloping Front," by V. E. Van Zant, Carleton College; March 1, 1950.

"The Image Orthicon Television Camera," by N. S. Bean, Radio Corporation of America; March 15, 1950.

"Radar Echoes from Snow, Rain, and Clouds," by J. S. Marshall, Faculty, McGill University; March 29, 1950.

NEW MEXICO

"Radio Astronomy," by C. R. Burrows, Faculty, Cornell University; March 13, 1950.

"The Application of Magnetic Amplifiers, Hydraulic Transmissions, and Magnetic Fluid Clutches in Servomechanism Problems," by P. D. Tilton, Vickers Electric Company; March 24, 1950.

NEW YORK

"Television Audio and Intercommunication Techniques and Systems," by O. J. Sather, Columbia Broadcasting System, and J. H. Copp, American Broadcasting Company; March 1, 1950.

"New Developments in Transmitting Tubes," by C. E. Rich and T. Sege, Sperry Gyroscope Company, D. A. S. Hale and R. J. Kircher, Bell Telephone Laboratories, and R. H. Rheaume, Machlett Laboratories; April 5, 1950.

OMAHA-LINCOLN

"Production and Detection of Waves Having More Than 100 Per cent Modulation," by A. J. L. Robertson, Faculty, University of Nebraska, and James Weblemoe, Faculty, Midland College; March 20, 1950.

OTTAWA

"Royal Canadian Navy Communications System," by D. C. Rutherford, Royal Canadian Navy; March 16, 1950.

"FM Discriminators," by Lieutenant Rious, CSRDE, Army; March 30, 1950.

PORTLAND

"Complex Frequency and Potential Analogy," by J. M. Pettit, Faculty, Stanford University; January 20, 1950.

"Tuning of Directional Broadcast Antennas," by Clifford Moulton, Graduate Assistant, Oregon State College; February 16, 1950.

"Application of Scale Models to Antenna Measurements," by G. L. Hollingsworth, Boeing Aircraft Company; March 20, 1950.

PRINCETON

"The Application of Statistical Measurement Techniques to Communication Problems," by J. B. Wiesner, Faculty, Massachusetts Institute of Technology; April 13, 1950.

ROCHESTER

"Serendipity, Cybernetics, and Electronics," by W. C. White, General Electric Research Laboratory; March 16, 1950.

SACRAMENTO

"New Developments in Audio Amplifiers," by Rudy Poucher, Normand B. Neeley Enterprises; December 13, 1949.

"Important Aspects of Tube Developments and Uses," by Clayton Murdock, Eitel-McCullough, Inc; January 17, 1950.

(Continued on page 44A)

REVERE FREE-CUTTING COPPER ROD

...INCREASES ELECTRONIC PRODUCTION

SINCE its introduction, Revere Free-Cutting Copper has decisively proved its great value for the precision manufacture of copper parts. Uses include certain tube elements requiring both great dimensional precision, and exceptional finish. It is also being used for switch gear, high-capacity plug connectors and in similar applications requiring copper to be machined with great accuracy and smoothness. This copper may also be cold-upset to a considerable deformation, and may be hot forged.

Revere Free-Cutting Copper is oxygen-free, high conductivity, and contains a small amount of tellurium, which, plus special processing in the Revere mills, greatly increases machining speeds, makes possible closer tolerances and much smoother finish. Thus production is increased, costs are cut, rejects lessened. The material's one important limitation is that it does not make a vacuum-tight seal with glass. In all other electronic applications this special-quality material offers great advantages. Write Revere for details.

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Executive Offices: 230 Park Avenue
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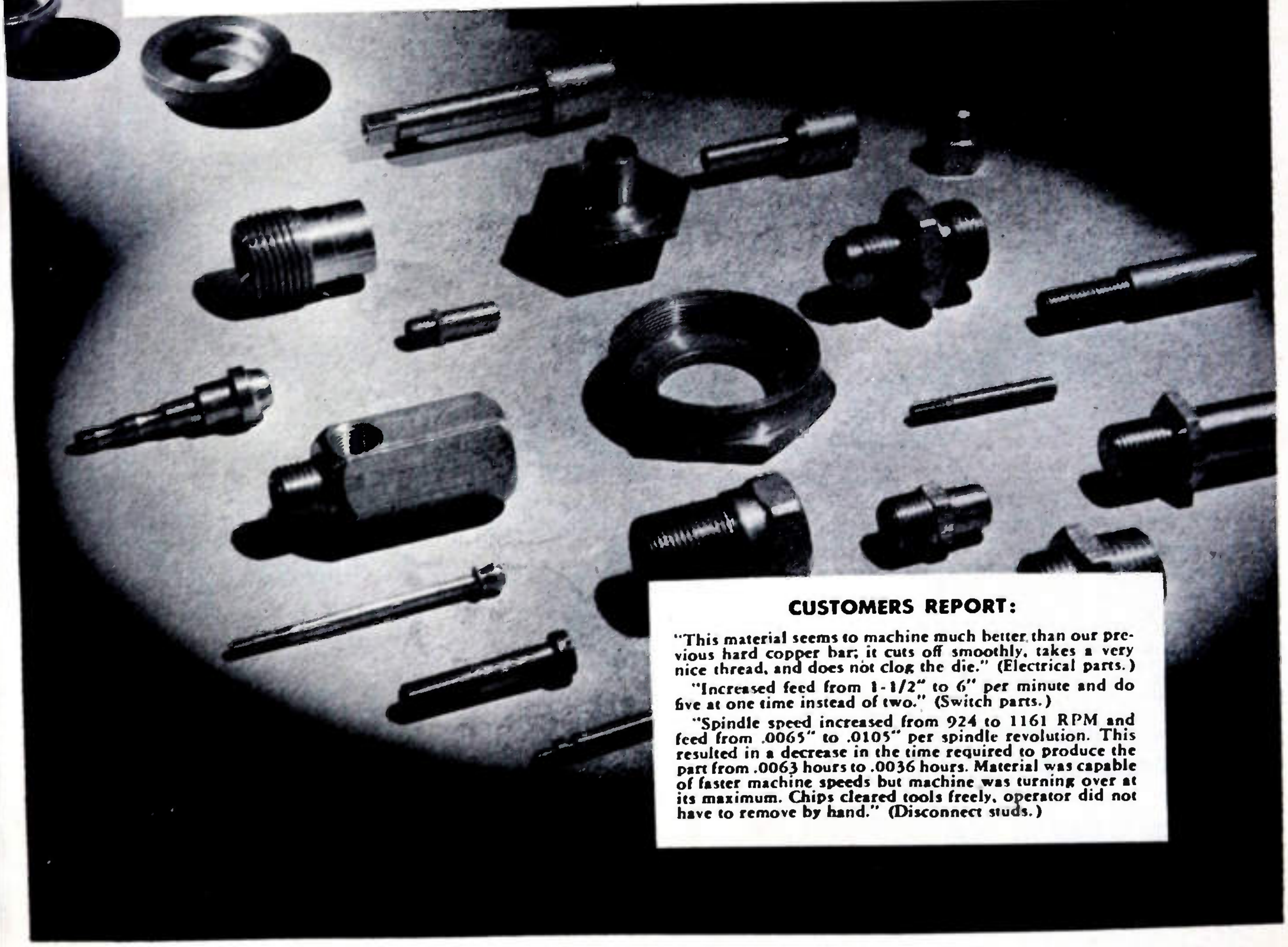

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"This material seems to machine much better than our previous hard copper bar; it cuts off smoothly, takes a very nice thread, and does not clog the die." (Electrical parts.)

"Increased feed from 1-1/2" to 6" per minute and do five at one time instead of two." (Switch parts.)

"Spindle speed increased from 924 to 1161 RPM and feed from .0065" to .0105" per spindle revolution. This resulted in a decrease in the time required to produce the part from .0063 hours to .0036 hours. Material was capable of faster machine speeds but machine was turning over at its maximum. Chips cleared tools freely, operator did not have to remove by hand." (Disconnect studs.)

SECTION MEETINGS

(Continued from page 42A)

"Electronics in Astronomy," by Gerald Kron. Faculty, University of California; February 14, 1950.

"Design and Manufacture of Television Receivers," by Boyd Farr and Allan Hyne, RCA Service, Inc; April 11, 1950.

SAN ANTONIO

"New Uses for Electronic Wizards," by C. N. Hoyler, RCA Laboratories; March 23, 1950.

SAN DIEGO

"Use and Applications of a 20-MEV Betatron," by W. K. Lyons, United States Navy Electronics Laboratory; March 21, 1950.

"The Role of Professional Groups in The Institute of Radio Engineers," by L. L. Van Atta, United States Naval Research Laboratory; April 4, 1950.

SCHENECTADY

"Electronics Park—Present Activities and Operations," by W. R. G. Baker, General Electric Company; February 27, 1950.

"The Color Television Controversy," by Walter Hausz, General Electric Company; April 10, 1950.

SEATTLE

"Microwave Installation of the Bonneville Power Administration," by R. F. Stevens and T. W. Stringfield, Bonneville Power Administration; March 21, 1950.

ST. LOUIS

"Recent Developments in Electronic Computers," by K. V. Newton, Emerson Electric Company; Color Film: "Gatti-Hallicrafter African Expedition"; March 23, 1950.

SYRACUSE

"Basic Principles of Computing Machines," by Burton Lester, General Electric Company; February 2, 1950.

"Engineering and Education," by K. G. Bartlett, Faculty, Syracuse University; March 16, 1950.

WASHINGTON

"Distortion and Noise in Communications Systems," by M. J. DiToro, Federal Telecommunications Laboratories; March 13, 1950.

"An Exploration of the Effects of Strong Radio-Frequency Fields on Micro-Organisms in Aqueous Solutions," by G. H. Brown, RCA Laboratories; April 10, 1950.

WILLIAMSPORT

"Design and Application of Electrolytic Capacitors," by Joseph Collins, Aerovox Corporation; Appointment of Nominating Committee; April 12, 1950.

SUBSECTIONS

AMARILLO-LUBBOCK

"Silicones," by M. H. Leavenworth, Dow-Corning Corporation; February 1, 1950.

"Selecting the Right Distribution System," by Bob McClure, Southwestern Public Service Company; March 16, 1950.

CENTRE COUNTY

"Semiconductor Devices," by R. W. Haegele, Sylvania Electric Products Inc; March 21, 1950.

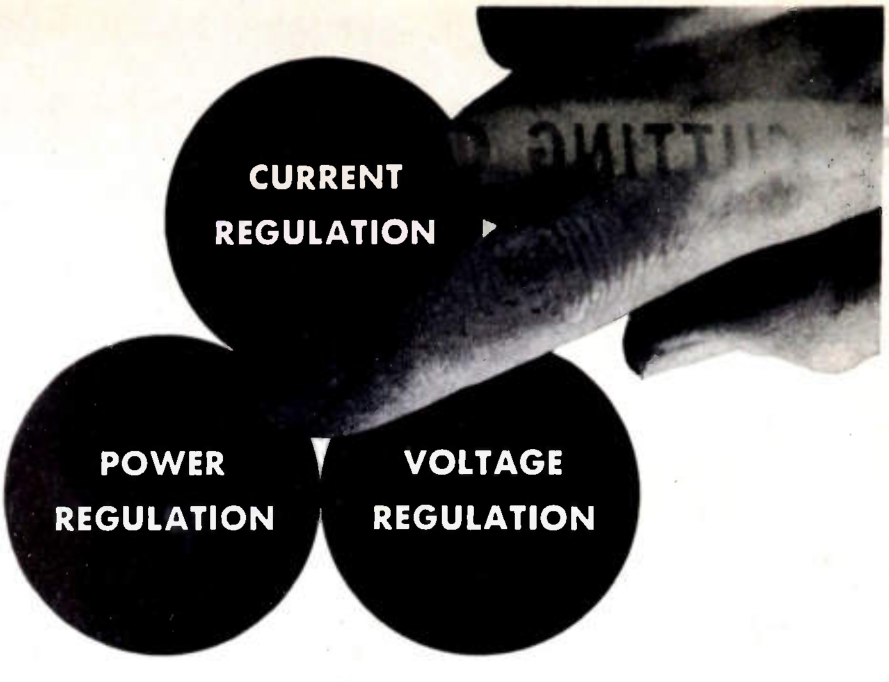
LONG ISLAND

"Magnetic Amplifiers," by F. G. Willey and F. S. Macklem, Servo Corporation of America; April 11, 1950.

NORTHERN NEW JERSEY

"Topics in Communication Theory," by C. E. Shannon, Bell Telephone Laboratories; and "Computers," by B. McMillan, Bell Telephone Laboratories; March 15, 1950.

"Some Aspects of 'G' Line Transmission," by H. Englemann, J. Kostriza, and D. D. Grief, Federal Telecommunications Laboratories; April 12, 1950.



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Type 310-A Z-Angle Meter — 30 to 20,000 c.p.s.

Measures impedance directly in polar co-ordinates as an *impedance* magnitude in ohms and *phase angle* in degrees $Z\angle\pm\Theta$. Measures, with equal ease, pure resistance, inductance, capacitance or complex impedances comprised of most any RLC combinations. Range: Impedance (Z), 0.5 to 100,000 ohms; Phase Angle (Θ), +90° (X_L) through 0° (R) to —90° (X_C). Accuracy: Within ± 1% for impedance and ± 2° for phase angle. Price: $425.00.

Type 311-A R-F Z-Angle Meter for radio frequencies — 100 kc to 2 mc.

Simplifies laboratory and field *impedance* and *phase angle* measurements. Ideal for checking impedance of coils, transformers, coupling networks, lines, filters, antennas, etc. Direct-reading Impedance Range: 10 to 5,000 ohms up to 200 kc, and 10 to 1,000 ohms at 1 mc. Phase Angle: +90° (X_L) through 0° (R) to —90° (X_C). Accuracy: Impedance to within ± 3%, and phase angle ± 4°. Price: $350.00.

Type 410-A R-F Oscillator — 100 kc to 10 mc. (Special models 46.5 kc to 4.65 mc available.)

Power oscillator for use as bridge driver and general laboratory measurements. Features: High stability, high output (approximate 30 volts), 50-60 Ω output impedance, expanded frequency scale, direct reading output voltmeter, compact design. Price: $350.00.

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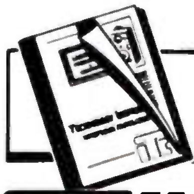
The *first* commercially available all-electronic instrument that directly measures the phase angle between two voltages *in a simple* operation. Ideally suited to applications in such fields as audio facilities, ultrasonics, servomechanisms, geophysics, vibrations, acoustics and many others.

Phase angle readings made directly without balancing . . . stable at frequencies as low as 2 to 3 cycles. Voltage range: 1 to 170 peak volts. Terminals for recorder . . . choice of relay-rack or cabinet mounting. Price $475.00. Cabinet $20.00.

Type 110 Slide-Wire Resistance Box

Convenient combination consisting of precision decade resistor and continuously adjustable slide-wire which provides smooth, continuous variation of resistance between decade steps (permits adjustment of resistance to one part in 10,000). For most applications, eliminates need for more elaborate multi-dial decade boxes. Ideal for student and general laboratory use. Decade resistance cards adjusted to within ± 0.1% of nominal values, and slide-wire resistors direct-reading to within 1% of their maximum values. Cast aluminum cabinet. All resistance elements completely enclosed. Suitable for use at audio and ultrasonic frequencies. Type 110-A, range 0-11,000 ohms: $42.50. Type 110-B, range 0-110,000 ohms: $45.00.



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STUDENT BRANCH MEETINGS

ALABAMA POLYTECHNIC INSTITUTE, IRE BRANCH

"Your Telephone Voice of the Future," by F. A. Woods, Southern Bell Telephone Company; March 28, 1950.

Election of Officers; April 11, 1950.

UNIVERSITY OF ARKANSAS, IRE BRANCH

Business Meeting; April 19, 1950.

BUCKNELL UNIVERSITY, IRE-AIEE BRANCH

"Fusetrons," by W. J. Ulmer, Jr., Bussman Manufacturing Company; March 15, 1950.

CLARKSON COLLEGE OF TECHNOLOGY, IRE BRANCH

"Stepping Along with Television," by Mr. Davenport, New York Telephone Company; March 16, 1950.

UNIVERSITY OF COLORADO, IRE BRANCH

"A New Long-Range Navigational Aid," by W. R. Luebke, Graduate Student, University of Colorado; Film: "Television"; April 5, 1950.

COLUMBIA UNIVERSITY, IRE-AIEE BRANCH

"Engineer-Employer Relationships with Respect to Inventions," by G. S. Rich; March 24, 1950.

UNIVERSITY OF DAYTON, IRE BRANCH

"Causes and Reduction of Hum," by Albert Chong, Student, University of Dayton; December 6, 1949.

"The Radio Compass," by Richard Thome, Student, University of Dayton; December 13, 1949.

"Radio Aids to Navigation," by Paul Northrop, Student, University of Dayton; December 20, 1949.

"FM Radio Altimeter," by Robert Cooper; January 10, 1950.

"FM Radio Altimeter," by Robert Cooper; January 31, 1950.

UNIVERSITY OF DELAWARE, IRE-AIEE BRANCH

Student Paper Presentation; Election of Officers; March 30, 1950.

FENN COLLEGE, IRE BRANCH

"Use of the Square Wave in Amplifier Testing," by F. A. Schwaller, Student, Fenn College; March 30, 1950.

UNIVERSITY OF FLORIDA, IRE-AIEE BRANCH

Field Trip to University of Florida Experimental Station and Electronics Laboratory; April 4, 1950.

Business Meeting; Election of Officers; April 18, 1950.

ILLINOIS INSTITUTE OF TECHNOLOGY, IRE BRANCH

"Shipboard Use of Electronic Equipment," by A. L. Gallin, Faculty, Illinois Institute of Technology; March 21, 1950.

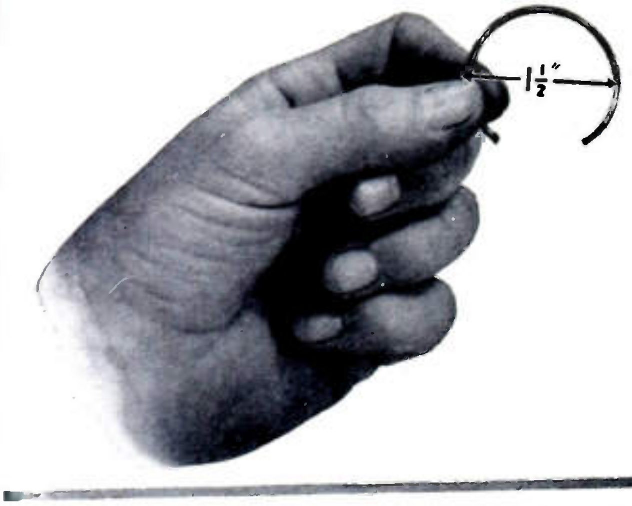
STATE UNIVERSITY OF IOWA, IRE BRANCH

Business Meeting; March 22, 1950.

Paper Submitted on Disc: "Universal Phonograph Styli," by J. D. Reid, Crosley Division of Avco Manufacturing Corporation; March 29, 1950.

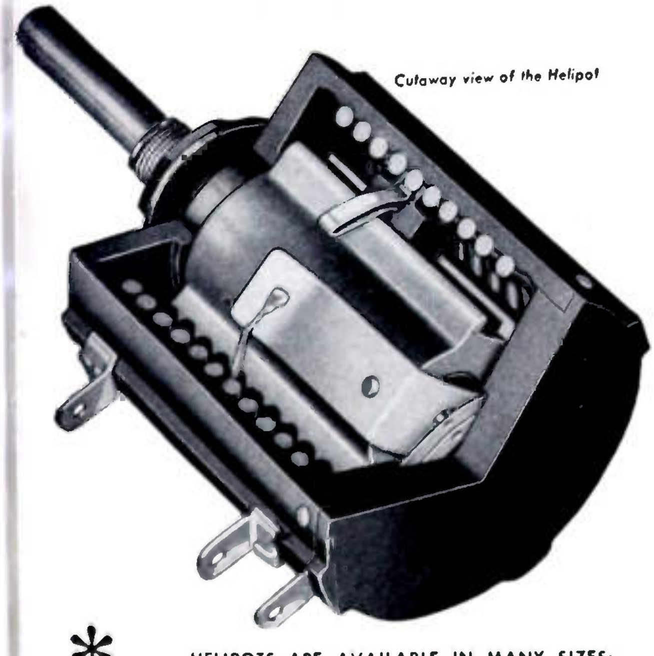
(Continued on page 48A)

Compare! Here's the Helipot Principle that is Revolutionizing Potentiometer Control in Today's Electronic Circuits



CONVENTIONAL POTENTIOMETERS have a coil diameter of approximately 1½" and provide only 4" (about 300°) of potentiometer slide wire control.

THE BECKMAN HELIPOT has the same coil diameter, yet gives up to 46" (3600°)* of potentiometer slide wire control—nearly TWELVE times as much!



Cutaway view of the Helipot

Some of the multiple Helipot advantages

EXTENSIVELY used on precision electronic equipment during the war, the Helipot is now being widely adopted by manufacturers of quality electronic equipment to increase the accuracy, convenience and utility of their instruments. The Helipot permits much finer adjustment of circuits and greater accuracy in resistance control. It permits simplifying controls and eliminating extra knobs. Its low-torque characteristics (only one inch-ounce starting torque*, running torque even less) make the Helipot ideal for power-driven operations, Servo mechanisms, etc.

And one of the most important Helipot advantages is its unusually accurate linearity. The Helipot tolerance for deviations from true linearity is normally held to within ± 0.5%, while precision units are available with tolerances held to 0.1%, .05%, and even less—an accuracy heretofore obtainable only in costly and delicate laboratory apparatus.

The Helipot is available in a wide range of types and resistances to meet the requirements of many applications, and its versatile design permits ready adaptation of a variety of special features, as may be called for in meeting new problems of resistance control. Let us study your potentiometer-rheostat problem and make recommendations on the application of Helipot advantages to *your* equipment. No obligation of course. Write today.

* HELIPOTS ARE AVAILABLE IN MANY SIZES:

MODEL A—5 watts, incorporating 10 helical turns and a slide wire length of 46 inches, case diameter 1¾", is available with resistance values from 10 ohms to 300,000 ohms.

MODEL B—10 watts with 15 helical turns and 140" slide wire, case diameter 3¼", is available with resistance values from 50 ohms to 500,000 ohms.

MODEL C—2 watts, with 3 helical turns and 13½" slide wire, case diameter 1¾", available in resistance from 5 ohms to 50,000 ohms.

MODEL D—15 watts, with 25 helical turns and 234" slide wire, case diameter 3¼", available in resistances from 100 ohms to 750,000 ohms.

MODEL E—20 watts, with 40 helical turns and 373" slide wire, case diameter 3¼", is available with resistance values from 200 ohms to 1,000,000 ohms.

Other types and designs of Potentiometers are also available.

• *Data is for Model A unit*

Send for the New Helipot Booklet!

THE Helipot CORPORATION, SOUTH PASADENA 6, CALIFORNIA

STUDENT BRANCH MEETINGS

(Continued from page 46A)

"The Drive-in Theatre Electrical Equipment," by W. J. Carr, "The English System of Radio Tube Markings," by Paul Meltzer, and "Fundamental Concepts of the Atomic Bomb," by John Stafford, Students, State University of Iowa; April 12, 1950.

Colored slides shown and explained by R. L. Riddle, Chester Lodge and James Fankhauser, Graduate Students, State University of Iowa; April 19, 1950.

IOWA STATE COLLEGE, IRE-AIEE BRANCH

"Electronic Instrumentation of Meteorological Quantities," by Bob Stewart, Graduate Student, Iowa State College; April 5, 1950.

UNIVERSITY OF KENTUCKY, IRE BRANCH

Movie: "Curiosity Shop," by Aluminum Company of America; March 14, 1950.

"Television—Its Mechanism and Promise," by W. L. Lawrence, RCA Victor; March 28, 1950.

LAFAYETTE COLLEGE, IRE-AIEE BRANCH

"Electricity in Lightning," by William Rohland, Student, Lafayette College; "Carrier Current," by S. S. Lesh, Pennsylvania Power and Light Company; March 30, 1950.

MARQUETTE UNIVERSITY, IRE-AIEE BRANCH

Election of Officers; March 30, 1950.

"Silicone Insulating Matrials," by Henry Worthman, Dow Corning Corporation; April 5, 1950.

Social Meeting; April 21, 1950.

UNIVERSITY OF MIAMI, IRE BRANCH

Election of Officers; March 29, 1950.

MICHIGAN STATE COLLEGE, IRE-AIEE BRANCH

"Modern Trends in Power Transmission," by Frank Sanford, Commonwealth Services; February 22, 1950.

"Operation of an Electronic Analogue Computer," by A. F. Martz, Jr., Holley Carburetor Company; March 28, 1950.

MISSOURI SCHOOL OF MINES AND TECHNOLOGY, IRE-AIEE BRANCH

"Trans-Oceanic Telegraphy," by R. H. Duncan, Graduate Student, Missouri School of Mines and Technology; Election of Officers; March 29, 1950.

UNIVERSITY OF NEBRASKA, IRE-AIEE BRANCH

"Ignitron Rectifier Application," by A. R. Edison, Student, University of Nebraska; April 5, 1950.

NEW YORK UNIVERSITY, IRE-AIEE BRANCH (EVENING DIVISION)

"Servomechanisms," by Charles Rehberg, Facelty, New York University; March 13, 1950.

"Planning of a Large Metropolitan Electric System," by J. Steinberg, Consolidated Edison Company; April 3, 1950.

UNIVERSITY OF NOTRE DAME, IRE-AIEE BRANCH

Student Paper Competition Finals; March 21, 1950.

"Television Antennas and Transmission Lines," by Ringland Krueger, American Phenolic Corporation; Election of Officers; April 13, 1950.

OHIO STATE UNIVERSITY, IRE-AIEE BRANCH

"The Importance of Geology to the Electrical Engineer," by Richard Anderson, Faculty, Battelle Memorial Institute; April 6, 1950.

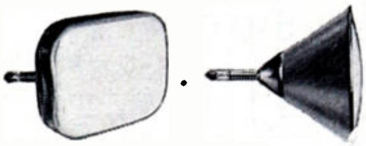
OREGON STATE COLLEGE, IRE BRANCH

"Safety in Engineering Practice," by Scott Hazen, Bonneville Power Administration; March 29, 1950.

(Continued on page 58A)

THIS IS THE FACTORY

THAT CUSTOM-MAKES TELEVISION PICTURE TUBES



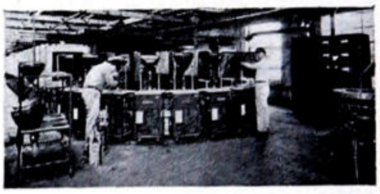
THAT CARRY THE NAME OF **Sheldon**. THIS IS HOW WE WASH

OUR GLASS BLANKS. THIS IS WHERE WE PHOSPHOR-COAT 'EM

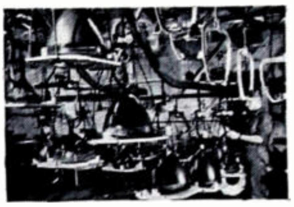
NOW WE TAKE 'EM

OVERHEAD TO PLACES WHERE WE DAG, BAKE &

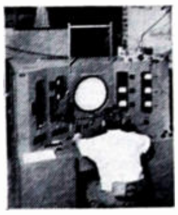
SEAL 'EM. AGAIN WE TAKE 'EM OVERHEAD TO THE EXHAUST



& THEN



WE BRING 'EM OVERHEAD TO AGE, BASE & FINAL TEST 'EM



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ELECTRICAL ENGINEER

As INSTRUCTOR in Electronics for Engineering College in the Chicago area. Should have teaching and research experience with a wide background in electronics field. Administrative experience very desirable.

Permanent position having good environment and excellent facilities. Desirable living quarters in new conventional housing available.

Outline in complete detail experience, research activities, accomplishments, and educational degrees. Doctorate desirable. Write fully regarding family status and salary requirements. Salary is flexible to meet ability and record of applicant selected. Congenial faculty knows of this opening. Replies will receive full confidence. Reply Box #613.

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Radio Engineers**

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ELECTRONIC ENGINEERS

Excellent opportunities are offered by one of the leading concerns in the electronic computer field to engineers with development or design experience in video and pulse circuitry or test and maintenance experience in the radar, television, or computer fields.

Send complete resumes and salary requirements to:

Personnel Department
**ECKERT-MAUCHLY COMPUTER
CORPORATION**
3747 Ridge Avenue
Subsidiary of Remington Rand Inc.
Philadelphia 32, Pa.

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.

MANUFACTURER-ENGINEER

Wanted for ceramic capacitor manufacturer-engineer familiar with manufacturing techniques, design and development of equipment for use in high speed production of ceramic capacitors. Should be capable of assuming complete charge of manufacturing program. Send resume of education, experience, salary desired to F-58, P.O. Box 3414, Phila, 22, Pa.

ELECTRICAL ENGINEER

Electrical engineer to design sound equipment, audio amplifiers and electric carillons. Requirements: B.S. degree, 3 years experience in audio, electronic and acoustical systems. Location: upstate New York. Box 602.

SALES MANAGER

Sales manager to head up sales force selling public address and intercommunicating systems for old line company. Technical knowledge as well as sales ability required. Location: upstate New York. Box 603.

ELECTRONICS ENGINEER

Opportunities for several experienced electronics engineers in communications, television receivers, television transmitters, classified military equipment, computers, microwave equipment and advanced development of all kinds. Desires graduate engineers with 5 years or more experience. Salary commensurate with ability and experience. We are looking for engineers seeking permanent connections with unusual opportunities for advancement. New and modern facilities and working conditions unequalled anywhere in the world. Reply, Personnel Div. Electronics Dept., General Electronic Co., Syracuse, N.Y.

PHYSICISTS-SENIOR ELECTRONIC ENGINEERS

Familiar with ultra high frequency and microwave techniques. Experience with electronic digital and/or analog, computer research and development program. Salaries commensurate with experience and ability. Excellent opportunities for qualified personnel. Contact C. C. Jones, Personnel Dept. Goodyear Aircraft Corp. Akron 15, Ohio.

ASSISTANT TO CHIEF ENGINEER

Large speciality transformer manufacturer wants man experienced in small transformer work. Excellent opportunity for qualified man. Please state education and experience, also salary on last or

(Continued on page 51A)

Senior Electronic Circuit Physicists

for advanced Research and Development

Minimum Requirements:

1. M.S. or Ph.D. in Physics or E.E.
2. Not less than five years experience in advanced electronic circuit development with a record of accomplishment giving evidence of an unusual degree of ingenuity and ability in the field.
3. Minimum age 28 years.

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Hughes Aircraft Company
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PHYSICISTS AND ENGINEERS

This established but expanding scientist-operated organization offers excellent opportunities for a future in completely new fields to alert experienced engineers and physicists who are weary of making minor improvements in conventional devices and techniques. Men with sound backgrounds and experience in the design of advanced electronic circuits, computers, or precision mechanical instruments, or with experience in gaseous discharges or applied physics are offered the opportunity to qualify for key positions in expanding and completely new fields. This company specializes in research and development work; its well-equipped laboratories are located in the suburbs of Washington, D.C.

**JACOBS
INSTRUMENT CO.**
4718 Bethesda Ave.
Bethesda 14, Maryland

POSITIONS OPEN

(Continued from page 50A)

present position. Southern Ohio location. All replies held strictly confidential. Suitable arrangements will be made to interview qualified applicants. Box 606.

TELEVISION ENGINEERS

Several engineers experienced in either technical or commercial phases of television are required in the formation of a new department. Send resume of qualifications to Personnel Dept., General Precision Laboratory, Inc., 63 Bedford Road, Pleasantville, New York.

PROFESSOR

Ph.D. or D.Sc. required. Age 40-45 with good teaching experience. Some industrial experience helpful. Large midwestern school, undergraduate and graduate programs. State salary expected and qualifications. Box 609.

CALIFORNIA ENGINEERS

Opportunity with California Communications Commission, $415-505. Requirements: California residence; 2 years communications experience; 1 year design, operation, maintenance radio communications systems and 1 year wire communications systems plus college graduate with major in E.E. or physics. Write before June 15th to Recruitment Representative, State Personnel Board, 1015 L St., Sacramento, California.

ENGINEERS

Engineers and assistants needed at new Motorola laboratory in Phoenix, Arizona. Engineers are required to be graduates of accredited engineering school, specialists in VHF and UHF receiver design, microwave communication pulse circuits, VHF, UHF and Microwave antenna design etc. Assistants must be engineering graduate with electronic experience. Replies should be sent to Daniel E. Noble, 4545 Augusta Boulevard, Chicago 51, Ill., stating education, experience and past salary schedules.

ELECTRONIC ENGINEERS

Excellent opportunities are offered by one of the leading concerns in the electronic computer field to engineers with development or design experience in video and pulse circuitry or test and maintenance experience in the radar, television or computer fields. Send complete resumes and salary requirements to: Personnel Dept., Eckert-Mauchly Computer Corp., 3747 Ridge Ave., Phila. 32, Pa.

ELECTRONICS ENGINEER

Electronics engineer about 35 years old who preferably has had some graduate training and who is experienced in electronic circuit and apparatus design and development work. Wanted by a small but expanding and well known company specializing in precision electronic instruments. Located in New Jersey about 30 miles from New York City. Salary up to $7,000 plus bonus. Our employees know of this ad. Box 611.

RCA VICTOR
Camden, N. J.
Requires Experienced Electronics Engineers

RCA's steady growth in the field of electronics results in attractive opportunities for electrical and mechanical engineers and physicists. Experienced engineers are finding the "right position" in the wide scope of RCA's activities. Equipment is being developed for the following applications: communications and navigational equipment for the aviation industry, mobile transmitters, microwave relay links, radar systems and components, and ultra high frequency test equipment.

These requirements represent permanent expansion in RCA Victor's Engineering Division at Camden, which will provide excellent opportunities for men of high caliber with appropriate training and experience.

If you meet these specifications, and if you are looking for a career which will open wide the door to the complete expression of your talents in the fields of electronics, write, giving full details to:

National Recruiting Division
Box 650, RCA Victor Division
Radio Corporation of America
Camden, New Jersey

PROJECT ENGINEERS

Real opportunities exist for Graduate Engineers with design and development experience in any of the following: Servomechanisms, radar, microwave techniques, microwave antenna design, communications equipment, electron optics, pulse transformers, fractional h.p. motors.

SEND COMPLETE RESUME TO EMPLOYMENT OFFICE.

SPERRY
GYROSCOPE CO.
DIVISION OF
THE SPERRY CORP.
GREAT NECK, LONG ISLAND

★ ★ ★ ★

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.

COMMUNICATIONS ENGINEER

B.S.E.E. West Virginia University, August 1949. Eta Kappa Nu, Sigma Pi Sigma. Age 24, married. 2 years AAF Radio Maintenance. Desires communications or electronic work anywhere in U.S. Box 395 W.

ENGINEER

Graduate June 1950 with B.S. in E.E. University of Virginia. Age 24. Tau Beta Pi. FCC radiotelephone 1st class license. Amateur radio operator. 2 years in U. S. Navy as radio-radar technician. 2 years design and development of electronic uniformity analyzer for textile materials. Interested in sales engineering, design and development. Box 405 W.

ASSISTANT PROFESSOR

B.S., M.S. in E.E. Illinois Institute of Technology and University of Illinois respectively. 1 year teaching experience. 2½ years experience in design of computer servos and missile systems. Desires to teach electrical engineering courses, preferably at a college located in rural or suburban area. Available June 1950. Box 406 W.

ELECTRICAL ENGINEER

Electrical engineer, graduate B.E.E., C.C.N.Y. Age 25, married. Some informal experience with amplifiers and magnetic recorders. Desires position, preferably in audio. Salary secondary. Box 407 W.

ENGINEER

Recent graduate of American Television Institute of Technology with B.S.T.E. Single. ART 1/c in Navy. 1st class radiotelephone FCC license. Desires position with a future in electronics, television or airlines as Junior Engineer or technician. Domestic or foreign assignment. Box 408 W.

ELECTRONIC ENGINEER

B.S. in physics (biology minor). Ex-Navy engineering officer. Graduate C.R.E.I. Married. Experience: electronics instruction, consultation, vacuum tube ruggedization. Desires position in medical electronics, biophysics or vacuum tube development in east. Box 409 W.

(Continued on page 54A)

ENGINEERS
Design, Development and Factory

Senior Electronics Eng., to take charge of the design and construction of a complete radar system. At least a B.S. degree and 5 yrs. experience in electronic circuit work, of which 3 yrs. must have been on radar circuits. Will be required to direct the activities of a group of engineers and technicians.

Factory Engineer, Magnetron Engineer with 2 or 3 yrs. Magnetron product engineering experience. Should have at least a B.S. degree and a knowledge of metals, fabrication and brazing of machined parts, glass to metal seals, exhaust procedure and microwave testing techniques.

Plant located in Boston offering excellent educational, recreational, and cultural facilities. Excellent program of employee benefits available.

Address replies to Personnel Dept., Sylvania Electric Products Inc., Electronics Division, 70 Forsyth St., Boston 15, Mass.

ATOMIC POWER

Position open for engineer, having at least a B.S. degree in electrical engineering and in addition having several years of practical experience in the development and design of instrumentation and servomechanisms. A thorough knowledge of electronic regulating systems, amplifiers, and simulators is also a requirement. The work is related to the development of regulating devices and instrumentation for control of a nuclear power plant. Location, suburb of Pittsburgh, Pennsylvania. For application write

Manager, Technical Employment
Westinghouse Electric Corporation
306 Fourth Avenue,
Pittsburgh 30, Pa.

STANDARD RI-FI* METERS

14kc to 1000mc!

DEVELOPED BY **STODDART**
FOR THE ARMED FORCES.

AVAILABLE COMMERCIALLY.

VHF!

15 MC
to
400 MC

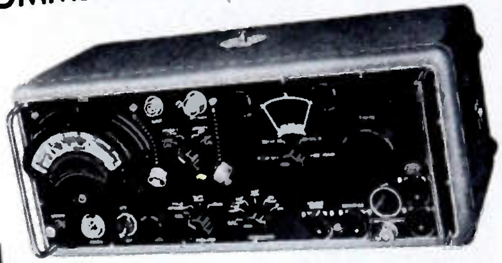
NMA - 5

Commercial equivalent of TS-587/U.
Sensitivity as two-terminal voltmeter, (95 ohms balanced) 2 microvolts 15-125 MC; 5 microvolts 88-400 MC. Field intensity measurements using calibrated dipole. Frequency range includes FM and TV Bands.

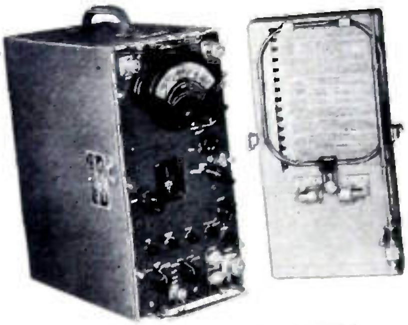
VLF!

14 KC
to
250 KC

NM - 10A



Commercial equivalent of AN/URM-6.
A new achievement in sensitivity! Field intensity measurements, 1 microvolt-per-meter using rod; 10 microvolts-per-meter using shielded directive loop. As two-terminal voltmeter, 1 microvolt.



HF!

150 KC
to
25 MC

NM - 20A

Commercial equivalent of AN/PRM-1.
Self-contained batteries. A.C. supply optional. Sensitivity as two-terminal voltmeter, 1 microvolt. Field intensity with ½ meter rod antenna, 2 microvolts-per-meter; rotatable loop supplied. Includes standard broadcast band, radio range, WWV, and communications frequencies.

UHF!

375 MC
to
1000 MC

NM - 50A

Commercial equivalent of AN/URM-17.
Sensitivity as two-terminal voltmeter, (50-ohm coaxial input) 10 microvolts. Field intensity measurements using calibrated dipole. Frequency range includes Citizens Band and UHF color TV Band.

Since 1944 **Stoddart** RI-FI* instruments have established the standard for superior quality and unexcelled performance. These instruments fully comply with test equipment requirements of such radio interference specifications as JAN-I-225, ASA C63.2, 16E4(SHIPS), AN-I-24a, AN-I-42, AN-I-27a, AN-I-40 and others. Many of these specifications were written or revised to the standards of performance demonstrated in Stoddart equipment.

The **rugged** and reliable instruments illustrated above serve equally well in field or laboratory. Individually calibrated for consistent results using internal standard of reference. Meter scales marked in microvolts and DB above one microvolt. Function selector enables measurement of sinusoidal or complex waveforms, giving average, peak or quasi-peak values. Accessories provide means for measuring either conducted or radiated r.f. voltages. Graphic recorder available.

*Radio Interference and Field Intensity.

Precision Attenuation for UHF!

Less than 1.2 VSWR to 3000 MC.
Turret Attenuator:
0, 10, 20, 30, 40, 50 DB.
Accuracy ± .5 DB.
Patents applied for.

STODDART AIRCRAFT RADIO CO.

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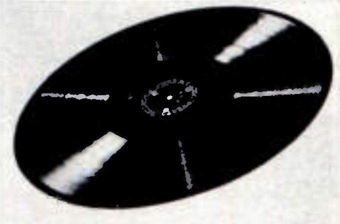
Hillside 9294



Positions Wanted

(Continued from page 52A)

ELECTRONIC ENGINEER

B.S.E.E. 1949 Vanderbilt University, working on M.S.E.E. at present at Syracuse University. Interested in television, electronics, radio communication. 4 years as Navy electronic technician. Tau Beta Pi. Age 26, married. Prefers mid-west, south. Available June 1950. Box 410 W.

AUDIO TECHNICIAN

Experienced in complex audio circuits. Graduate RCA Institutes. 3 years technical training in Army Air Forces. College background. Age 25. Box 411 W.

ELECTRONIC ENGINEER

Harvard B.S. June 1950. Broad physics and electrical engineering background with some emphasis on pulse and timing circuits. 2 years training in U. S. Navy as electronic technician (10 mos. as instructor). 1st class radiophone license, light experience with TV and audio. Age 22. Desires work in electronics anywhere but prefer eastern U.S. Box 412 W.

ENGINEER

B.S. in E.E., M.S. in communications engineering. Expect Ph.D. from Harvard in June 1950. Tau Beta Pi, Eta Kappa Nu. 4 years experience as Air Force radar officer specializing in tactical suitability tests of airborne equipment. 1 year in geophysical operations. Age 32. Married. Box 413 W.

SALES OR FIELD ENGINEER

B.S.E.E. communications major. 5½ years experience in radar and microwave systems test equipment and components. Additional experience in photocell, UHF transmitters and receivers, and servos; also phases of production, development, and supervisory management. Age 29, married, 1 child. Will travel or relocate. Box 414 W.

ELECTRONIC ENGINEER

M.S.E.E., B.S.E.E., communications, Oklahoma A. and M. College. 2 years Oil Company Research Lab. 2 years military Airborne radio Maintenance. Thesis work parasitic antenna array impedance. 1st radiotelephone, Amateur class A, DXCC. Married, age 24. Desires design, development, research anywhere. Box 429 W.

ELECTRONIC ENGINEER

B.S.E.E. June 1950, University of Missouri. Age 26, top one-eighth of class, Eta Kappa Nu. 3 years research laboratory technician, 2 years broadcast Chief Engineer, active HAM. Interested in development, design or application. Box 430 W.

RADIO ELECTRONICS TECHNICIAN

American, age 27, single. Desires long term position anywhere in Philippines. Amiable disposition. Speaks some Tagalog and Visayan. 10 years military, amateur and commercial radio experience. Box 431 W.

ELECTRONICS ENGINEER

B.E.E. September 1949, Georgia Institute of Technology, communications option. Age 27, single. Former Navy ETM. Limited production experience. Desires position with a future in electronics or communications. Salary secondary. Location immaterial. Box 432 W.

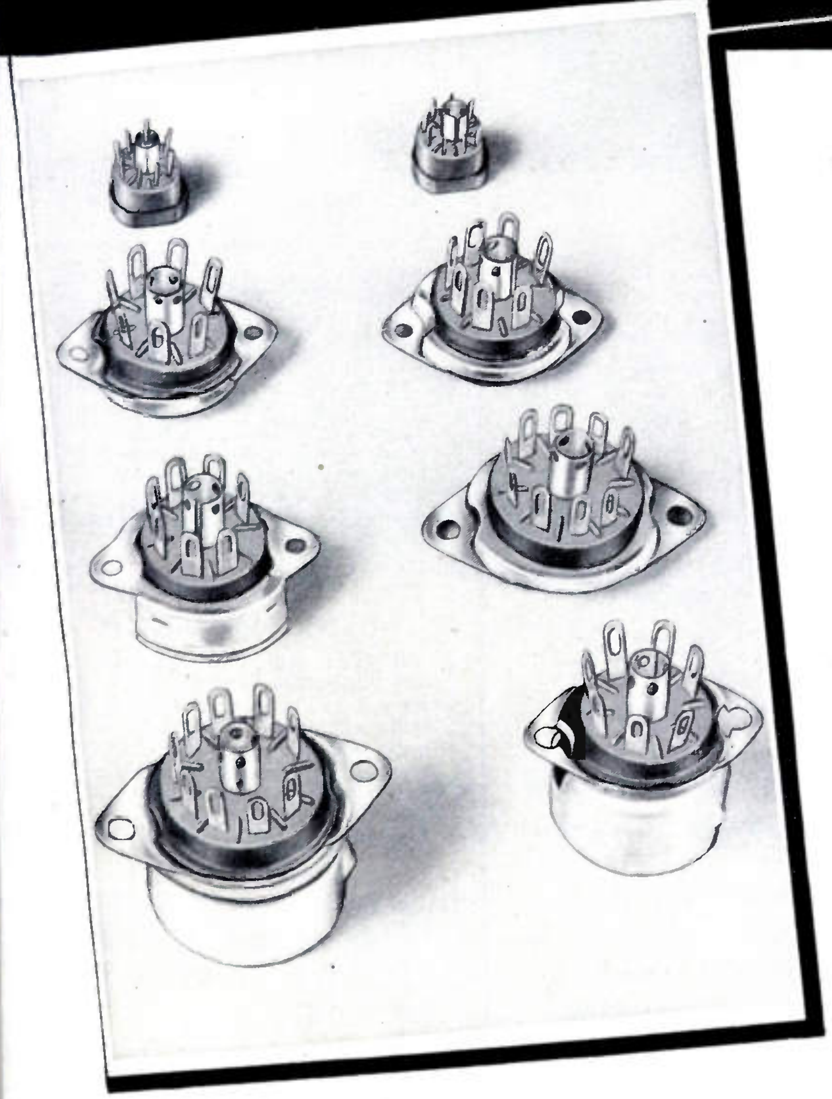
(Continued on page 56A)

MYCALEX

MINIATURE TUBE SOCKETS

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

New Low Prices



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.

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THE INSULATOR
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"Owners of 'MYCALEX' Patents"

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

Positions Wanted

(Continued from page 54A)

PHYSICIST

B.S. physics, University of Washington 1949, age 26. 6 years sub-professional radio and radar experience. FCC licensed radio telephone, 1st class. Primary interest: Electronic instrumentationy anywhere in U. S. Box 433 W.

ENGINEER

B.S.E. June 1949, University of Nebraska. Communications options. Single, age 26. Eta Kappa Nu, Pi Mu Epsilon. Desires position offering prospects of research or of work toward advanced degree. Salary and location secondary. Box 434 W.

ENGINEER

B.E.E. January 1950. Communications option. Rensselaer Polytechnic Institute. First quarter of class, 2 years Army experience as Chief Carrier Repeaterman. Age 31, single. Neat appearance. Salary and location secondary. Interested in communications, electronics and UHF. Box 435 W.

COMMUNICATIONS ENGINEER

7 years training in electronics. B.S.E.E. Graduate Navy electronics school and National Radio Institute. 1st class phone FCC license. Age 23, single. Technician experience. Interested in TV or other communications engineering. Box 436 W.

ENGINEER—EXECUTIVE

B.S. Iowa State. M.S. Purdue. M.B.A. Stanford, Graduate School of Business. E.E. Stanford, Graduate School of Engineering. 3 years Naval engineering officer; Bowdoin M.I.T. radar schools, radar material sea duty. Experience in research, engineering, production, business and industrial economics. Tau Beta Pi, Eta Kappa Nu, Sigma Xi. Interested in operation of a small business manufacturing a technical electronic product, or in a responsible position with a larger company in electronics industry. Box 437 W.

ELECTRONIC ENGINEER

B.S.E.E. and three quarters toward M.S. Specialist in electronics, radio and control. 1½ years as B-29 wing hq. staff radar and radio officer; 2 years as E.E. instructor at prominent university. Age 30, married, 1 child. Conscientious, alert, congenial, excellent references. Primary interest: electronic or control development. Box 438 W.

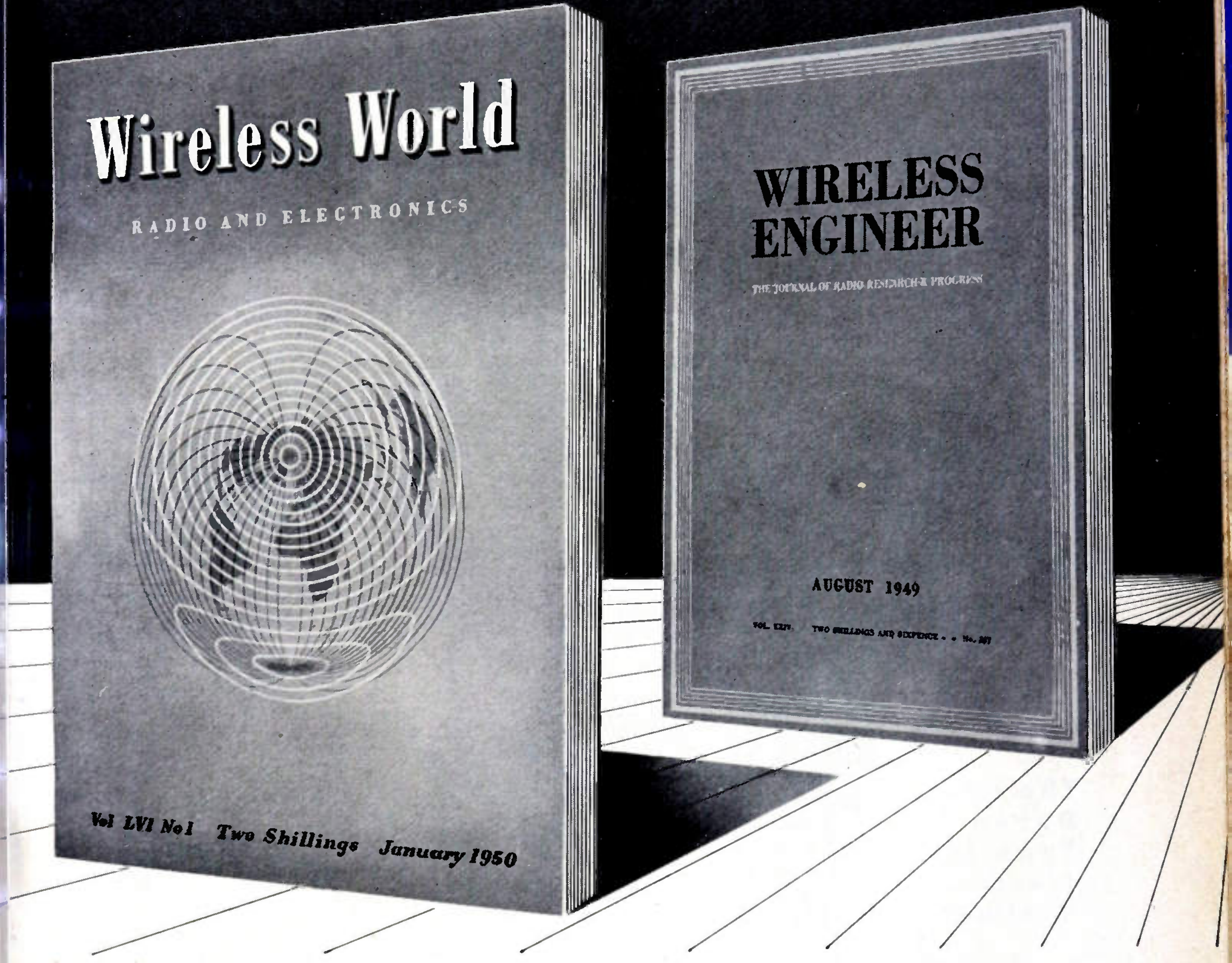
ENGINEER

Engineering Aide. Graduate RCA Institutes, 2 years advanced technology course, 2½ years experience as a laboratory electronic mechanic. Age 29, married. College background. Prefer work in eastern U. S. Box 439 W.

ENGINEER-PHYSICIST

B.E.E. January 1949; M.A. in physics, May 1950, Columbia Univ.; Experience in teaching, design and construction of highly specialized electronic instrumentation involving research and technical paper publication. HAM experience 10 years on UHF to SHF. Member Tau Beta Pi, Eta Kappa Nu. N.Y. State P.E. in training. Age 27, single. Prefer New York City vicinity but not exclusively. Box 440 W.

(Continued on page 58A)


Wireless World
RADIO AND ELECTRONICS
Vol LVI No 1 Two Shillings January 1950
WIRELESS ENGINEER
THE JOURNAL OF RADIO RESEARCH & PROGRESS
AUGUST 1949
TWO SHILLINGS AND SIXPENCE

Positions Wanted

(Continued from page 56A)

ELECTRONIC ENGINEER

Electronic engineer. Age 41. 7 years college; 15 years responsible practical experience in electronic operation, design and research. Now employed as consultant on job which will be completed this fall. Desires responsible position, credentials forwarded upon request. Salary open. Address: American, 72 Mococa, Sao Paulo, Brazil.

SALES, FIELD, OR APPLICATION ENGINEER

B.S.E.E. June 1950, Newark College of Engineering. Age 24. Single. Interested in position which combines engineering and "people." Navy 3 years, Electronic Technician 1/c. Some teaching, writing and industrial experience. Member Omicron Delta Kappa. Resume on request. Box 441 W.

STUDENT BRANCH MEETINGS

(Continued from page 48A)

"A Direct Reading RF Wattmeter," by W. S. Burdic, Faculty, Oregon State College; "Plate Load Characteristics of a Push-Pull Audio Power Amplifier," by Lloyd Craine, Faculty, Oregon State College; "Distortion in Directional Broadcast Antennae," by Clifford Moulton, Faculty, Oregon State College; April 1, 1950.

Pennsylvania State College, IRE-AIEE Branch

"Automatic Train Control," by P. X. Rice, Faculty, Pennsylvania State College; Nomination of Officers; April 13, 1950.

Pratt Institute, IRE Branch

Business Meeting; March 15, 1950.

"Local Broadcasting Stations," by George Travis, Student, Pratt Institute; March 29, 1950.

"Servomechanisms," by Charles Rehberg, Faculty, New York University; April 13, 1950.

Rhode Island State College, IRE-AIEE Branch

Business Meeting; January 16, 1950.

Business Meeting; Election of Officers; April 21, 1950.

Rutgers University, IRE-AIEE Branch

Business Meeting; Nomination of Officers; April 4, 1950.

"System Protection Relays," by W. K. Sonnemann, Westinghouse Electric Corporation; April 18, 1950.

San Diego State College, IRE Branch

"Radio Astronomy," by R. F. Meyers, Student, San Diego State College; March 7, 1950.

"Television Propagation," by John Day, Naval Electronics Laboratory; March 28, 1950.

University of Southern California, IRE-AIEE Branch

"Color Television," by Willard Geer, Faculty, University of Southern California; March 14, 1950.

Film and Guided Tour; March 16, 1950.

Tour of Radio Station KFI-TV conducted by Bryan Cole; March 23, 1950.

(Continued on page 59A)

Student Branch Meetings

(Continued from page 58A)

Tour of Radio Station KECA-TV conducted by Ernest Thornton; March 25, 1950.

Tour of Pacific Telephone and Telegraph Los Angeles Central Office; April 1, 1950.

Syracuse University, IRE-AIEE Branch

Election of Officers; March 28, 1950.

University of Tennessee, IRE Branch

Election of Officers; April 11, 1950.

Tufts College, IRE-AIEE Branch

"The Design and Erection of a Television Station," by S. V. Stadig, Radio Station WBZ-TV; March 15, 1950.

"Radar Countermeasures in World War II," by F. D. Lewis, General Radio Company; April 12, 1950.

Tulane University, IRE-AIEE Branch

"Arc Welding in Refrigeration," by E. L. Peterson, and "Electronics in Neuropsychiatry," by W. M. Nunn, Students, Tulane University; March 24, 1950.

University of Wyoming, IRE Branch

"Civil Aeronautics Administration Growing Up," by M. R. Neary, Civil Aeronautics Administration; January 26, 1950.

"Telephone Company—Past, Present and Future," by C. R. Lewis, Mountain States Telephone and Telegraph Company; February 16, 1950.

Film: "Jet Propulsion," by General Electric Company; March 9, 1950.

Yale University, IRE-AIEE Branch

"Recent Developments in the Electrical Power Engineering Field," by R. G. Warner, United Illuminating Company; March 30, 1950.

"Microwave and Acoustic Lenses," by W. E. Kock, Bell Telephone Laboratories; April 12, 1950.

The following transfers and admissions were approved and will be effective as of June 1, 1950:

Transfer to Senior Member

Abramovich, M. N., 9903 Lorain Ave., Silver Spring, Md.
Adams, G. J., 299 Atlantic Ave., Boston 10, Mass.
Allured, R. B., 904 Dewey, Ann Arbor, Mich.
Arnett, R. A., Box 7166, Dallas 9, Tex.
Benson, J. E., 4 Beaumont Ave., West Ryde, N.S.W., Australia
Byers, V. J., 341 Sutherland Dr., Toronto 17, Ont., Canada
Caldwell, C. M., 5511 Fairglen Lane, Chevy Chase 15, Md.
Cole, W. A., 67 Elwood Blvd., Toronto 12, Ont., Canada
D'Orio, P. A., 1225 N. Linden Ave., Oak Park, Ill.
Duckett, E. J., 319 Barnes St., Pittsburgh 21, Pa.
Eaton, T. T., RCA Victor Division, Camden, N. J.
Flowers, H. L., 208 Elmira St., S.W., Washington 20, D. C.
Gronner, A. D., 50-17-63 St., Woodside, L. I., N. Y.
Hicken, J., 40 Andrews Ave., Binghamton, N. Y.
Holtz, R. F., Box 177, Great Notch, N. J.
Jenny, H. K., R.D. 6, Lancaster, Pa.
Kees, H., 3312 Lake Dr., R.D. 3, Evansville, Ind.
Lewis, F. D., General Radio Company, 275 Massachusetts Ave., Cambridge 39, Mass.

(Continued on page 60A)

MEMBERSHIP

(Continued from page 59A)

Marchese, T. J., 159 Hillside Ave., Nutley 10, N. J.
McConnell, T. E., 122-33 St. Dr., S.E., Cedar Rapids, Iowa
Michaels, S. E., Bell Telephone Laboratories, Inc., Murray Hill, N. J.
Morrissey, T. G., 4161 E. 16 Ave., Denver 7, Colo.
Rao., M. V. S., Indian Radio & Television Institute, Box 33, Buckinghampet P.O., India
Russell, C. M., Electronics Test, U. S. Naval Air Test Center, Patuxent River, Md.
Sands, L. G., Philco Corporation, Industrial Division, Philadelphia, Pa.
Sieminski, E., 134-14 Franklin Ave., Flushing, N. Y.
Stone, R. P., RCA Laboratories, Princeton, N. J.
Walker, R. M., 75 Brent St., Dorchester 24, Mass.
Warriner, B., IV, Federal Telecommunication Laboratories, 500 Washington Ave., Nutley, N. J

Admission to Senior Member

Astin, A. V., 5008 Battery Lane, Bethesda 14, Md.
Bond, W., 5930-14 St., N., Arlington, Va.
Carne, G. G., 46 Valley View Rd., Rockaway, N. J.
Fielder, D. C., School of Electrical Engineering, Georgia Institute of Technology, Atlanta, Ga.
Harris, L. M., Jr., Stromberg Carlson Company, Rochester 4, N. Y.
Harris, W. A., 101 N. Spring Garden Ave., Nutley 10, N. J.
Howard, J. H., Box 340, Highland, N. Y.
Kleen, W. J., C.S.F.-C.R.T., 23 Rue du Maroc, Paris 19, France
Mika, H. S., 444 S. Highland, Dearborn, Mich.
Oser, E. A., Patent Department, Bldg. 5-2, Radio Corporation of America, Camden, N. J.
Taylor, N. H., 48 Central St., Manchester, Mass.
Walsh, C., 437 Walnut St., Emporium, Pa.

Transfer to Member

Affelder, C., 6 Valley St., Newark 6, N. J.
Anderson, W. H., Ashburn, Ont., Canada
Baldwin, J. H. G., 5062 Granville St., Vancouver, B. C., Canada
Cotts, A. C., Electrical Engineering Department, Kansas State College, Manhattan, Kan.
Dagavarian, H. O., 1809-51 St., Brooklyn 4, N. Y.
Dobbins, B. D., 10108 Georgia Ave., Silver Spring, Md.
Ellis, R. E., 2640 N. Roosevelt St., Arlington, Va.
Fezer, H., Ogden Park, Dobbs Ferry, N. Y.
Fielding, B. L., 11331 Valley Spring Lane, N. Hollywood, Calif.
Jensen, A. M., 189 Riverview Ave., Little Silver, N. J.
Jones, W. B., Jr., Electrical Engineering Dept., Georgia Institute of Technology, Atlanta, Ga.
Koski, J. I., 2228 Yucca, Fort Worth 11, Tex.
Lowman, R. V., 122 Cutter Mill Rd., Great Neck, N. Y.
Massell, E., Box 433, Locust, N. J.
Meinheit, C. E., Box 6, Bayside, L. I., N. Y.
Mobley, M. P., Jr., 4610 Laurel Grove Ave., N. Hollywood, Calif.
Pace, R. T., Box 6, Bayside, L. I., N. Y.
Pickens, D. H., 9807 S. Yates Ave., Chicago 17, Ill.
Robbiano, P. P., c/o Electronics Research, Inc., Box 327, Evansville 4, Ind.
Schaefer, R. G., Electrical Engineering Department, University of Wyoming, Laramie, Wyo.
Schott, F. W., 3605 Richmond St., San Diego 3, Calif.
Sessions, S. H., 1886 Malden St., San Diego 52, Calif.
Shaffer, C. V., 206 Annis Blvd., Rt. 5, Gainesville, Fla.

(Continued on page 61A)

MEMBERSHIP

(Continued from page 60A)

Stantz, L. H., 168 Moeller St., Binghamton, N. Y.
Ungvary, R. L., 2009 Summer St., Stamford, Conn.
Witten, A. L., Jr., 20 Evelyn Rd., Port Washington, N. Y.

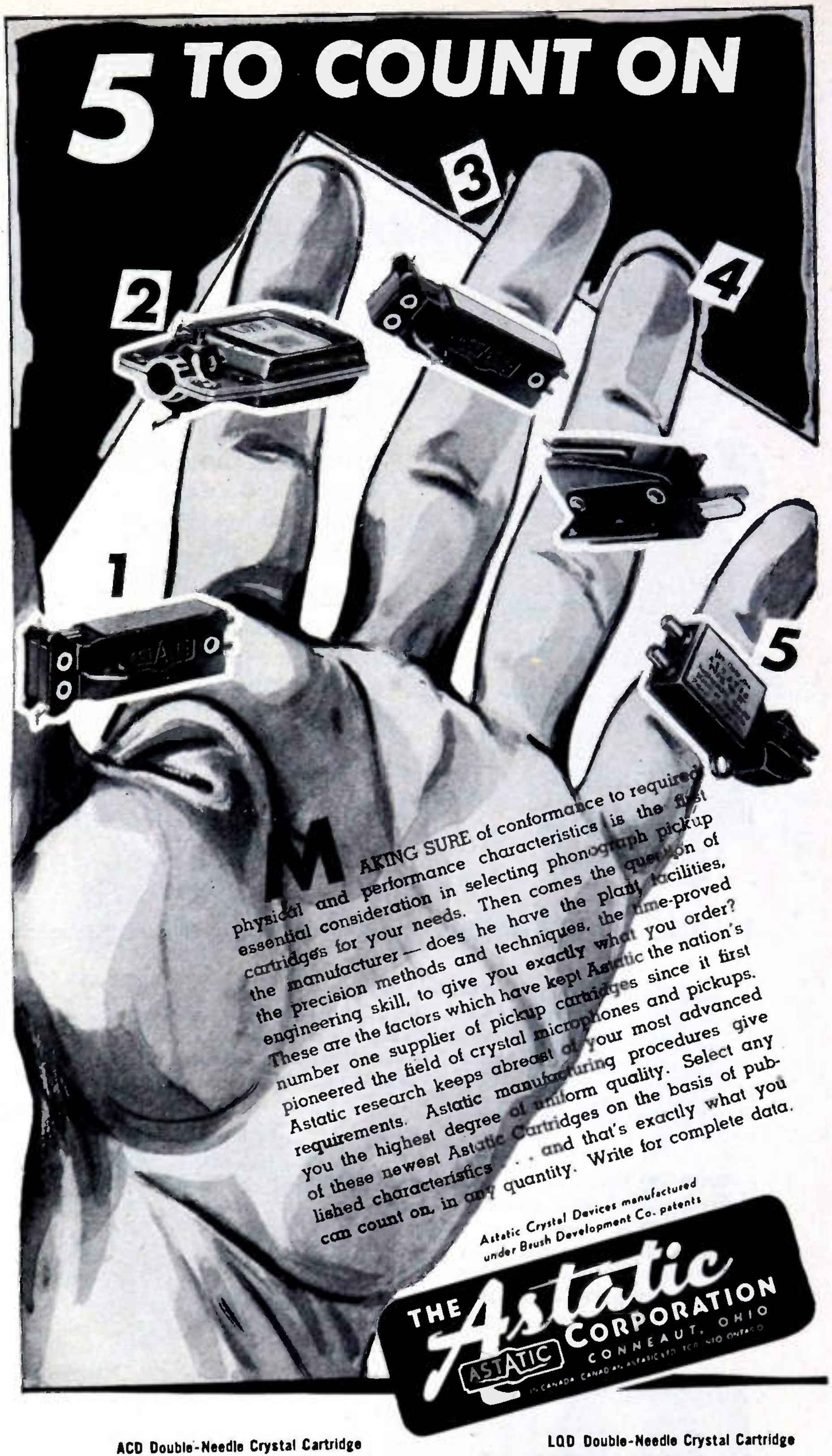
Admission to Member

Arnold, W. F., 1313 Beuna Vista, Ventura, Calif.
Bishop, A. R., 5907 S.E. Hill St., Portland 22, Ore.
Bishop, G. W., 7833 Penn., Kansas City 5, Mo.
Bobenhouse, C. J., 1421 Harrison, Des Moines, Iowa
Bradford, W. O., 155 W. Guava St., Oxnard, Calif.
Chapman, A. K., 136 Radio Security Sq., Brooks A.F. Base, Tex.
Childs, A. F., 605 Adams Ave., W. Hempstead, N. Y.
Connell, J. C., 516 E. Grand Ave., Albuquerque, N. Mex.
Fitzgerald, H. M., 131 Frederick St., Clinton, Ont., Canada
Gordon, L., 157 Broad St., Red Bank, N. J.
Hill, R. R., Mutual Telephone Company, Box 2200, Honolulu, T. H.
Hockman, P. E., R.D. 2, Springfield, Ohio
Huynen, J. R., Hillsdale Apt. 256-D, 37 Ave., San Mateo, Calif.
James, W. E., 68-01A Cloverdale Blvd., Bayside, L. I., N. Y.
Kulmus, R. J., 1665 Hager St., Utica 4, N. Y.
Lee, W., Military Communications Division, Sig. Sect. GHQ FEC, APO 500, c/o PM, San Francisco, Calif.
Lippke, J. A., 35 Church St., Roslyn, N. Y.
Martinez, B., Manrique #6, Apt. 2, Havana, Cuba
Mason, F. L., Electronics Office, Naval Shipyard, Pearl Harbor, T. H.
Meyer, R. G. H., Coles Signal Laboratory, Red Bank, N. J.
Moore, J. D. B., 27 McClary Ave., London, Ont., Canada
Nelson, A. B., 618 Burlington Ave., Bristol, Conn.
Randals, A. W., 1907 Stratford Ave., Neptune, R.D. 1, Asbury Park, N. J.
Redden, M. S., Jr., 4124 N. Third Rd., Arlington, Va.
Rhodes, A. J., Rm. 310, M.C.A., D.H.Q., Adamton House, Monkton, Ayrshire, Scotland
Sengupta, D. P., Assistant Technical Officer, Aeronautical Communication Station, Calcutta Airport, Dum Dum, India
Warnke, G. F., 35 W. 33 St., Chicago 15, Ill.

The following elections to the Associate grade were approved and were effective as of May 1, 1950:

Abbitt, C. W., 1521 Plum St., Springfield, Ohio
Alaimo, J. H., 4652 N. Kenmore Ave., Chicago 40, Ill.
Aseltyne, J. G., 188 King, Box 304, London, Ont., Canada
Ayer, D. R., Box 1, Bell Aircraft Corporation, Buffalo, N. Y.
Balaban, S. A., 1018 Hathaway St., Owensboro, Ky.
Bargfeld, F. G., 3044 N.E. 46 Ave., Portland, Ore.
Barrett, C. S., 2121 Mershon Ave., Dayton, Ohio
Barwick, M. N., 1815 N.W. 48 St., Miami, Fla.
Bary, V., Louisiana Ave., Bridgeport, Conn.
Boehme, C. D., 4 Talbott Bldg., Dayton, Ohio
Bradley, J. R., 7 Eighth St., Hicksville, N. Y.
Brock, K. S., 66 Needham St., Newton Highlands, Mass.
Brooks, L. P., 81 Lincon Ave., Tuckahoe, N. Y.
Brown, W. A., 7120 N. Barnett Lane, Milwaukee, Wis.
Burak, M., 153 Water St., Quincy 69, Mass.
Butler, J. J., 4330 N. Leavitt St., Chicago 18, Ill.

(Continued on page 62A)



5 TO COUNT ON

MAKING SURE of conformance to required physical and performance characteristics is the first essential consideration in selecting phonograph pickup cartridges for your needs. Then comes the question of the manufacturer — does he have the plant facilities, the precision methods and techniques, the time-proved engineering skill, to give you exactly what you order? These are the factors which have kept Astatic the nation's number one supplier of pickup cartridges since it first pioneered the field of crystal microphones and pickups. Astatic research keeps abreast of your most advanced requirements. Astatic manufacturing procedures give you the highest degree of uniform quality. Select any of these newest Astatic Cartridges on the basis of published characteristics . . . and that's exactly what you can count on, in any quantity. Write for complete data.

Astatic Crystal Devices manufactured under Brush Development Co. patents

THE Astatic CORPORATION
CONNEAUT, OHIO
IN CANADA: CANADIAN ASTATIC LTD. TORONTO ONTARIO

ACD Double-Needle Crystal Cartridge

1 Newest Astatic miniature turnover model featuring mechanical drive system with new low in inertia. Result — sensationally smooth response, new tracking excellence, low needle talk. Output 1.0 volt at 1,000 c.p.s. Needle pressure six grams.

AC Crystal Cartridge

3 Tiny, single-needle version of the new ACD, with same unparalleled smooth response. AC-J for slow speed records has five gram needle pressure; AC-AG-J, with special All-Groove needle tip for all record types, has six gram needle pressure; AC-78-J for 78 RPM records has six gram needle pressure. Output of each is 1.0 volt at 1,000 c.p.s.

LQD Double-Needle Crystal Cartridge

2 The PROVED TOP PERFORMER for turnover type pickups today. Outstanding for excellence of frequency response, particularly at low frequencies. Output 1.2 volts on slow speed side, needle pressure six grams; 0.9 volt on 78 RPM side, eight grams. Available with or without needle guards.

GC Ceramic Cartridge

4 The first ceramic cartridge with replaceable needle. Takes the "Type G" needle — with either one, three-mil or special All-Groove tip, precious metal or sapphire — which slips from its rubber chuck with a quarter turn sideways. Output has been increased over that of any other ceramic cartridge available. Light weight and low minimum needle pressure.

CQ Crystal Cartridge

5 Features miniature size and five-gram weight. Models CQ-J and CQ-AG-J fit standard 1/2" mounting and RCA 45 RPM record changers. Model CQ-IJ fits RMA No. 2 Specifications for top mounting .453" mounting centers. Output 0.7 volt at 1,000 c.p.s. Employ one-mil tip radius "Q" Needle, or special All-Groove tip (Model CQ-AG-J).

MEMBERSHIP

(Continued from page 61A)

Buzzard, R., Box 832, Waipahu, Oahu, T. H.
Caggiano, V. D., 8222 Murietta Ave., Van Nuys, Calif.
Capasso, L. V., 2504 Bathgate Ave., New York, N. Y.
Cenci, A. A., 11023 Strathmore Dr., Los Angeles, Calif.
Chalekian, J. S., 239 W. 31 St., Indianapolis, Ind.
Chambers, W. M., Radio Station WLOS, Asheville, N. C.
Clithero, W. H., Jr., 1835 Sul Ross, Houston 6, Tex.
Cochran, K. W., 12531 Barbara Ave., Venice, Calif.
Cohen, A., 109 Jackson Ave., Schenectady 4, N. Y.
Colbert, S. M., Franklin Annex, Cornell University, Ithaca, N. Y.
Conley, W. H., 110 Calyer St., Brooklyn, N. Y.
Curtis, W. T., 924 Lawrence Ave., Chicago, Ill.
DeVincent, J., 227 E. 33 St., New York 16, N. Y.
Diener, B., 3636 Rosewood Ave., Venice, Callf.
Dunn, E. J., 1498 E. 172 St., New York, N. Y.
Egli, E. M., 19 Roetelstrasse, Zurich, Switzerland
Farrington, H. W., 1772-20 Ave., Oakland 6, Calif.
Feinne, L., 8768-254 St., Bellerose, L. I., N. Y.
Fjeldsted, N. B., 1433 S. Rexford Dr., Los Angeles, Calif.
Flynn, L. A., 3105 Meeting St., Naval Base, S. C.
Franks, C. B., 2909 Delmar Ave., Cleveland, Ohio
Ganiere, H. C., 46 Rennell Ave., Lexington Park, Md.
Gault, A. A., 3620 Rutherford St., Harrisburg, Pa.
Gerber, W., Direction Generale Des PTT, Berne, Switzerland
Gilmer, R. L., 2728 Gilbert Ave., Portsmouth, Ohio
Gilmer, W. R., 1608 Harrisville Ave., Portsmouth, Ohio
Golden, H., 1514 Unionport Rd., New York, N. Y.
Gray, G. A., 4117 Illinois Ave., N.W., Washington 11, D. C.
Green, J. S., 116 S. Willow St., Compton, Calif.
Green, L. F., 1425 Chase Ave., Chicago, Ill.
Griffin, D. C., 13 S. Main St., Simsbury, Conn.
Guth, R. J., 215 W. 23 St., New York 11, N. Y.
Haahn, F. J., 3915 Dismount St., Dallas, Tex.
Hailey, W. H., 6340 Walrond, Kansas City, Mo.
Hax, D. H., 3635 Tracy Ave., Kansas City, Mo.
Hellstroem, P. G., Box 279, Götheborg, Sweden
Henderson, J. A., 4125 S. Anthony Blvd., Ft. Wayne, Ind.
Hogan, H. A., 1304 N. Astor St., Milwaukee, Wis.
Hornbostel, W. G., 8042 Altavan Ave., Los Angeles, Calif.
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Jennett, N. E., Jr., Box 342, Camarillo, Calif.
Jobe, R. E., 433 Maluniu Ave., Lanikai, Oahu, T. H.
Johnson, R. L., 23 Linden St., Norwood, Mass.
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Jones, J. T., Jr., 9128 Old Newton Rd., Philadelphia, Pa.
Jordal, V. M., Buffalo Center, Iowa
Kahn, I., 115 Broadway, New York 6, N. Y.
Kandra, J. P., 1255 Flora, N.W., Grand Rapids, Mich.
Kaplan, H., 1227 Grand Concourse, New York 52, N. Y.
Kappes, A. V., R.R. 6, Evansville, Ind.
Kline, C. M., 6915 Horrocks St., Philadelphia, Pa.
Klotzel, E., Caixa Postal 3301, São Paulo, Brazil
Krauer, O., 51 Manhattan Ave., Crestwood, N. Y.
Leichtman, J. A., c/o Brater, 978 Aldus Street, New York, N. Y.
Lewand, J., 1786 Galena, Denver, Colo.
Lampl, M. B., 828 Brooklyn St., San Antonio, Tex.
Lindemann, E. G., 2672 Lowery Ave., Honolulu 54, T. H.
Lyon, R. E., 7342 Lasaine Ave., Van Nuys, Calif.
Machuta, A. J., 38 Bright Rd., Hatboro, Pa.
Madeley, P. E., 145 Madeley St., Conroe, Tex.
Maravich, R., 107 W. 73 St., Chicago 21, Ill.

(Continued on page 63A)

MEMBERSHIP

(Continued from page 62A)

McLaughlin, F. G., 1243 Collingwood Ave., Akron, Ohio
Mehray, P. N., 550 Riverside Dr., New York, N. Y.
Meyer, B., 1435 E. 21 St., Brooklyn 10, N. Y.
Miller, W. E., 4012 Ruskin St., Houston 5, Tex.
Mirakentz, A., 4901 Ninth Ave., Los Angeles, Calif.
Montague, L. E., Box 3045, 760 Laurel, Beaumont, Tex.
Mueller, A. G., 418 F St., Rock Springs, Wyo.
Napoli, C. A., 197-26 Foothill Terr., Holliswood, N. Y.
Needs, W. R., Rt. 1, Box 976, Fort Lauderdale, Fla.
Noskowicz, T. S., 151 Highland Ave., Wooddale, Ill.
O'Connell, T. P., 97 Willow Ave., New York 54, N. Y.
Ohlemacher, R. F., Box 548, State College, N. Mex.
Penmessa, V. R., 5 Varadarakamudali Lane, Madras, S. India
Pheil, C. E., 2401-23 Ave., S., St. Petersburg, Fla.
Plutt, J. A., Jr., 215 W. 23 St., New York, N. Y.
Pressnall, J. R., 4606 Kansas Ave., N.W., Washington, D. C.
Profitt, C. L., Box 722, Augusta, Ga.
Randall, W. T., 4717 Campbell Dr., Culver City, Calif.
Raschke, R. R., 177 Eighth St., Troy, N. Y.
Redfield, S. I., 920 S. Michigan Ave., Chicago, Ill.
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Rivers, M. D., 1327 Grigsby Ave., Dallas 4, Tex.
Robertson, W. J. P., RCAF Sta., Clinton, Ont., Canada
Rockwell, D. A., 4430 Sheridan Rd., Chicago, Ill.
Rockwell, R. G., 5335 W. 99 Pl., Los Angeles, Calif.
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Rudolf, S., 3026 Brighton & 14 St., Brooklyn 35, N. Y.
Rumbold, D. G., 202 Huron St., London, Ont., Canada
Samalonis, A. J., 545 Court St., Elizabeth, N. J.
Schleppegrell, J. R., 88 Drake St., Charleston, S. C.
Shock, W. R., 700 Pine St., Central Falls, R. I.
Sinclair, J. G., Jr., 7400-3-4 Arizona, Los Angeles, Calif.
Singh, K. J., Block E2, 3GTS, Hospital Town, W., Bangalore, S. India
Slocomb, G. M., 734 E. Olive St., Burbank, Calif.
Smith, R. J., 176 Davidson Ave., Buffalo, N. Y.
Smith, R. T., 3377-B Castle Heights Ave., Los Angeles, Calif.
Spellman, F. C., 206 E. Fourth St., Brooklyn 18, N. Y.
Sturrock, J. K., 4412-51 St., Red Deer, Alta., Canada
Subbaraman, V., c/o Radio Sales & Service, Shoranur Rd., Trichur, S. India
Taft, E. A., Jr., Tropical Radio Telegraph Company, Tegucigalpa, Honduras
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Wegner, D. A., 131 Saranton Ave., Lake Bluff, Ill.
Weinberg, D. H., 421 S. Sixth, La Crosse, Wis.
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Wolff, K. A., 1713 N. Orchard St., Chicago 14, Ill.
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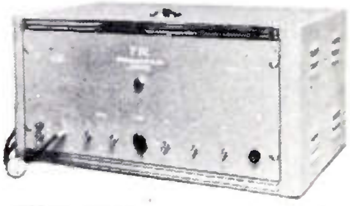
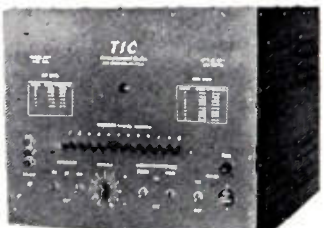


TYPE 2120
R.F. PICTURE SIGNAL GENERATOR

Provides picture and sound carrier. Modulated by standard R.M.A. composite picture signal. Sound carrier stability suitable for testing Inter Carrier type receivers. Internal 400 cycle FM and External audio with 75 microsecond pre-emphasis. Output max. 0.1v p-p across 75 ohm line. Available channels 2-13.



TYPE 1200A
12 CHANNEL
R.F. SWEEP GENERATOR

Intended for precise adjustment of R.F. head oscillator coils and R.F. band pass circuits. Pulse type markers at picture and sound carrier frequencies extend to zero signal reference base line. Accuracy of markers 0.02% of carrier frequency. 12 to 15 MC. sweep on all channels. Max. 1.V peak output across a 75 ohm line. Provisions for balanced input receivers. Instant selection by push button.

TYPE 1900
CRYSTAL CONTROLLED
MULTI-FREQUENCY GENERATOR

A 10 frequency, 400 cps. modulated crystal controlled oscillator, ideal for production line adjustment of stagger tuned I.F. amplifiers. Available with crystals ranging from 4.5 to 40 M.C. Output frequency accurate to 0.02%. Immediate push button selection of frequency. Output attenuator range .5V to 500 microvolts. Self contained regulated power supply.

TYPE 1500 A
I.F. WOBBULATOR

A two band sweeping generator covering the range of 4.5 to 50 M.C. Capable of a band width of approximately ±25% on either band. Five pulse type crystal generated markers to specified frequencies available for each band. Accuracy of markers .05%. Zero signal reference base line, with markers extending to base line. 1.V output max. into 75 ohms. A saw sweep available for "X" axis of scope.

Write for Detailed Engineering Data Sheets.

Tel-Instrument Co. Inc.
54 PATERSON AVENUE • EAST RUTHERFORD, 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 30A)

Miniature Sensitive Electronic Relay

A new electronic relay system to provide super sensitivity in industrial control applications has been developed by **Servo-Tek Products Co.**, 4 Godwin Ave., Paterson 1, N. J.

This economical electronic relay system utilizes recently developed cold-cathode type gas tubes. Simplicity of design and minimum number of components make possible this unit which mounts on a standard 4-inch electrical connection box.

The unit operates from the 115-volt 50-60 cps line and uses no filament to draw standby power with the relay circuit energized. Unit operation is initiated by the contact of a drop wire, contact making instrument, or any conductive medium between the two contacts having a resistance as high as one million ohms. Maximum current flow through the initiating resistance is in the order of microamperes.

The load relay contacts are arranged to permit a choice of either opening or closing a circuit, or simultaneously to open one circuit and close another.

A hold circuit is incorporated for optional use.

Pulse-Rise Time Indicator

A new Model 632-B, pulse-rise time indicator, with a rise time of 0.005 to 0.1 microsecond in 20 steps, is available from the manufacturer **Electronic Systems Co.**, 555 E. Tremont Ave., New York 57, N. Y.

(Continued on page 65A)

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

(Continued from page 64A)

The instrument employs a specially designed delay line of variable length and a vacuum-tube voltmeter. The voltmeter reads the peak of the resultant pulse produced by the pulse under test and its reflection; its magnitude is a function of the rise time and the delay of the line. When the delay is less than the rise time, the resultant pulse amplitude will be smaller than that of the impressed wave; when the delay of the line is equal to the build-up time of the pulse, the voltmeter reading will have reached maximum. Thus the calibrated line is a measure of the rise time and may be used to plot the leading edge. This instrument is also available in other ranges upon request.

Portable Oscilloscope

Waterman Products Co., Inc., 2445 Emerald St., Philadelphia 25, Pa., introduces the new Model S-14-B wide-band pocketscope. This oscilloscope demonstrates that laboratory oscilloscopes can be made increasingly practical through portability and still be characterized by superior electrical performance.

Suitable for random pulse analysis, or transient investigation, the S-14-B has amplifier fidelity constant within 2 db from dc to above 700 kc without peaking. Its linearized triggered or repetitive time base is continuously variable from ½-cps to 50 kc with either + polarity of sync or trigger. Amplifier sensitivity is of the order of 50 mv rms/inch and internal calibration of trace amplitude is provided. Observation of limited wave-form areas is facilitated by a trace expansion better than four times screen face. Input attenuators and gain controls are nonfrequency discriminating.

Decade Inductors

Available from **General Radio Co.,** 275 Massachusetts Ave., Cambridge 39, Mass., in single-decade units for building into other equipment and in three- and four-decade cabinet assemblies for laboratory use, are new decade inductors which provide precise decade steps of inductance from one millihenry to one henry per step.

(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)

Cores are molybdenum permalloy dust toroids with precisely adjusted, banked windings. Temperature coefficient of inductance is—24 parts per million per degree Centigrade over the normal range of room temperatures, and maximum storage factor Q is between 200 and 330. Accuracies range from 2 per cent for the one millihenry steps to 0.25 per cent for the 1-henry steps.

RF Interference Test Equipment

A new radio interference locator, Model 302, for the 550 kc to 30 Mc frequency range has just been made available by the **Sprague Products Co.**, North Adams, Mass.

The instrument utilizes a sensitive 8-tube superheterodyne circuit and operates either from self-contained batteries or 115 volt supply. An auxiliary inverter power supply is available for storage battery operation.

Each locater is supplied with both a loop and a collapsible rod antenna for normal use. An rf search probe, insulated for 35,000 volts ac, is also available for field use as is an audio probe for circulating current faults and cable fault location.

Other features of the Locator include a built-in loudspeaker, built-in dual range output meter and battery test meter, calibrated rf and audio gain controls, a beat frequency oscillator for detecting unmodulated signal sources, etc.

A complete description of the new instrument is given in Bulletin M-446, available upon letterhead request.

(Continued on page 67A)

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)

Low Cost DC Power Supply

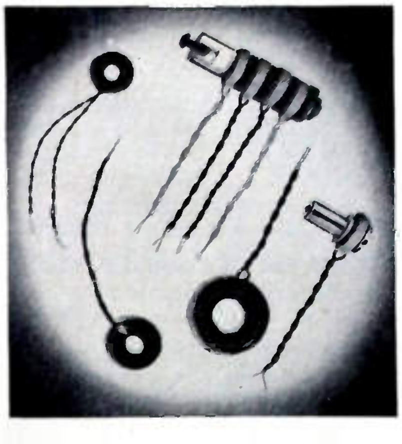
A new Model "BJ" Junior, just announced by **Electro Products Laboratories, Inc.**, 4501 N. Ravenswood Ave., Chicago 40, Ill., offers a low-cost source of filtered power, utilizing the same application of selenium rectifiers used in their Model "B." This application, using conduction cooling, doubles the rectifier power rating, dissipates over three times the heat, and provides lower cost per ampere output.

The new "BJ" Junior supplies 1- to 12.5-amperes, 6 volts, continuous duty, with an intermittent rating up to 25-amperes. Supplies 3 to 9 volts at other ratings, operating from 115 volts, 50/60 cps power source.

Designed for demonstrating and testing low voltage devices, the "BJ" Junior is constructed of heavy duty components which withstand high overloads, with an ac ripple of less than 0.4 volt at 6 volts 8 amperes.

Standoff Mounts for Toroidal Inductors

The design of various types of standoffs for mounting all types of uncased



toroidal inductors is announced by **Torocoil Company**, 5387 Northland Ave., St. Louis 12, Mo.

(Continued on page 68A)

Want the World's Finest Tape Recordings?

THERE'S A Magnecorder FOR EVERY PURPOSE ...EVERY PURSE!

3 Heads!

PT63-A to MONITOR YOUR MAGNECORDINGS

Three separate heads — *erase, record,* and *playback* for monitoring from tape — prevent recording errors. Same high fidelity and flexibility as the Magnecorder PT6-A — the world's most widely used professional tape recorder. New PT63-J Amplifier has separate playback and recording amplifiers to monitor from the tape. Includes 10 watt audio amplifier which also will drive external speaker.

OR CONVERT YOUR PT6-A TO MONITOR KIT 101

Conversion kit includes a three-head unit, monitor amplifier and power supply to modernize your present PT6-A. Head unit plugs into receptacles for present two-head unit.

The New PT-7 Series

3 Heads (erase, record, playback for monitoring from tape) in single housing, yet separately alignable, replaceable. New positive drive. 2-speed hysteresis-synchronous motor. Push-button controls can be remotely operated. Uses 7" or 10½" N.A.B. reels. 3 channel portable amplifier has high-level mixing.

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RACK or
CONSOLE

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World's Largest and Oldest Manufacturers of Professional Magnetic Recorders

MODERN
RADIO TECHNIQUE

This series, edited by J. A. Ratcliffe, deals with advances in radio technique made during and since the war. All the books are written by men responsible for important advances in the subjects they discuss. All are illustrated with plates and diagrams. More titles will follow.

The Principles and Practice of Wave Guides
L. G. H. HUXLEY. An introductory survey, based upon courses on micro-wave technique given during the war, of the phenomenally rapid development since 1940 of radar equipment in Britain and the United States. *328 pp. $4.75*

Radio Aids to Navigation
R. A. SMITH. Concerned mainly with the application of radio technique to navigation of aircraft, this book also discusses briefly the navigation of ships. It shows how many limitations of radio were overcome by new methods, especially in radar. *114 pp. $2.50*

Principles of Radar
DENIS TAYLOR and C. H. WESTCOTT. This exposition of the principles of radar defines the common factors which underlie radar design and the many types of equipment used during the war. It is largely concerned with the performance of radar equipment and its suitability for a specified purpose. "*Will delight any technical student.*"—Proceedings of IRE. *141 pp. $350*

Velocity-Modulated Thermionic Tubes
A. H. W. BECK. An expert on velocity-modulation tubes and their operation presents a general introduction to and the theory of the interchange of energy between field and beam, with the application of this theory to types of V. M. tube. *180 pp. $3.75*

Aerials for Metre and Decimetre Wavelengths
R. A. SMITH. Since a comprehensive description of aerial systems used for short-wave radio would not be possible in one volume, the author illustrates general principles with selected applications, largely from radar where the outstanding developments have taken place. *218 pp. $3.75*

Recent Advances in Radio Receivers
L. A. MOXON. Concentrating on advances made during the war years, this book deals mainly with circuits used with modern valves. The emphasis is on receivers for radar and on the design of intermediate frequency amplifiers to follow the crystal frequency-changer. *183 pp. $4.00*

Aerials for Centimetre-Wave-Lengths
D. W. FRY and F. K. GOWARD. The theory and application of scanning aerials for use on wave-lengths of 10 cm. and less, emphasizing principles of design rather than constructional details, and the advantages of particular types of aerial to meet particular requirements. Non-scanning Directive and Broadcast aerials are discussed. *182 pp. $3.50*

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51 MADISON AVENUE, NEW YORK 10, N.Y.

News—New Products

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(Continued from page 67A)

These types consist of units for individual mounting or units which can be used for stacking as many as four toroids. Both of these types solve the problem of firmly attaching and insulating the unfiased toroids from the chassis. These mounts supplement the complete line of toroids that are available to the industry through this organization.

Battery Powered G-M Scaler

A portable battery-operated scaler Model 80, designed to meet the need for determinations of radioactive levels in locations where conventional power supplies are not available or where line transients make conventional scalers unreliable, has been designed and developed by **Berkeley Scientific Co., 6th. St. & Nevin Ave., Richmond, Calif.**

It is not a survey-type meter in the usual sense, but rather a true-Geiger-Mueller Scaler with which accurate counts can be made.

The instrument consists of a G-M tube and probe, a scale-of-eight electronic counter, a mechanical register, and an adjustable, high-voltage, battery supply. It has a maximum continuous counting rate of 14,400 counts per minute and will resolve individual pulses at 90 microseconds. The summation of the counts is indicated by the mechanical register, plus the reading of the interpolation microammeter.

Wide-Spaced Sections to Cool Selenium Rectifiers

A new line of Selenium rectifiers with wide-spaced sections for rapid cooling has been announced by **Sarkes Tarzian Inc.,** Bloomington, Ind.

(Continued on page 69A)

Dependable
PERFORMANCE
at
LOWER COST

H&P

TOWER LIGHTING EQUIPMENT

H & P lighting equipment, consistently specified by outstanding radio engineers, is furnished as standard equipment by most leading tower manufacturers.

300 MM CODE BEACON
Patented ventilator dome circulates the air, assures cooler operation, longer lamp life. Concave base with drainage port at lowest point. Glass-to-glass color screen supports virtually eliminate color screen breakage. Neoprene gaskets throughout. CAA approved

MERCURY CODE FLASHER
Lifetime-lubricated ball bearings. No contact points to wear out. Highest quality bronze gears. Adjustable, 14 to 52 flashes per minute

SINGLE and DOUBLE OBSTRUCTION LIGHTS
Designed for standard A-21 traffic signal lamps Prismatic globes meet CAA specifications.

"PECA" SERIES PHOTO-ELECTRIC CONTROL
Turns lights on at 35 f.c.; off at 58 f.c., as recommended by CAA. High-wattage industrial type resistors. Low-loss circuit insulation.

ALSO COMPLETE LIGHT KITS FOR A-2, A-3, A-4 and A-5 TOWERS

PROMPT SERVICE and DELIVERY
First-day shipments out of stock. Immediate attention to specifications and unusual requirements.

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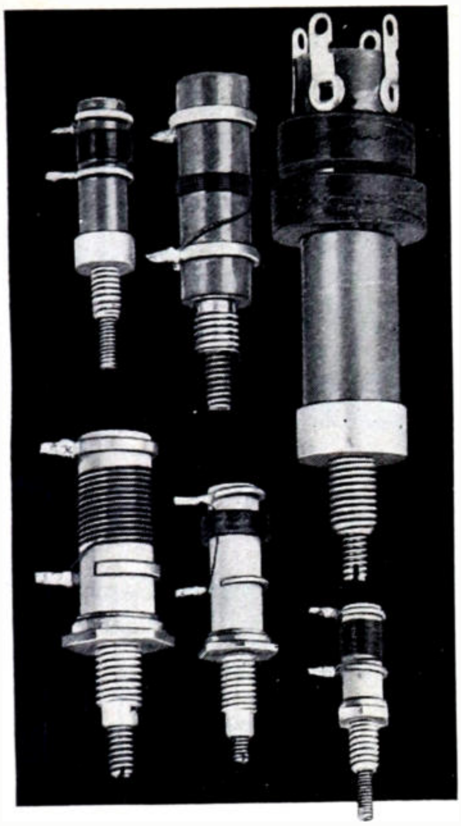


Which Of These Coil Forms Best Fits YOUR Needs?

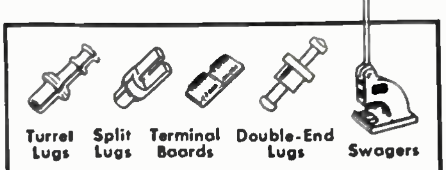
Coil Forms Only, Or Coils Wound To Your Specifications . . . Cambridge Thermionic will furnish slug tuned coil forms alone or wound with either single layer or pie type windings to fit your needs, in high, medium or low frequencies . . . and in small or large production quantities.

See table below for physical specifications of coil forms.

SEND COMPLETE SPECIFICATIONS FOR SPECIALLY WOUND COILS

Coil Form	Material	Mounting Stud Thread Size	Form O.D.	Mounted O.A. Height
LST	L-5 Ceramic	8–32	3/16"	19/32"
LS6	L-5 Ceramic	10–32*	1/4"	27/32"
LS5	L-5 Ceramic	1/4–28*	3/8"	1 1/16"
LSM	Paper Phenolic	8–32	1/4"	27/32"
LS3	Paper Phenolic	1/4–28	5/8"	1 1/8"
LS4†	Paper Phenolic	1/4–28	1/2"	2"

*These types only provided with spring locks for slugs.
†Fixed lugs. All others have adjustable ring terminals.
All ceramic forms are silicone impregnated. Mounting studs of all forms are cadmium plated.



custom or standard the guaranteed components

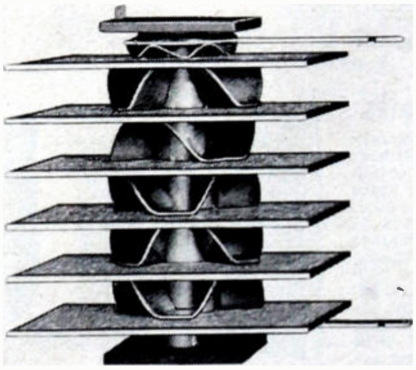
CAMBRIDGE THERMIONIC CORP.
456 Concord Ave., Cambridge 38, Mass.
West Coast Stock Maintained By: E. V. Roberts,
5014 Venice Blvd., Los Angeles, California

News—New Products

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

(Continued from page 68A)

Described as "Centre-Kooled," these rectifiers have been designed for use in radio, TV, or electronic equipment, and feature new developments in design. The center cooling feature provided by a special spacer between the cells insures lower over-all operating temperatures by allowing air to reach the portions of the cells in which the current density is the greatest.

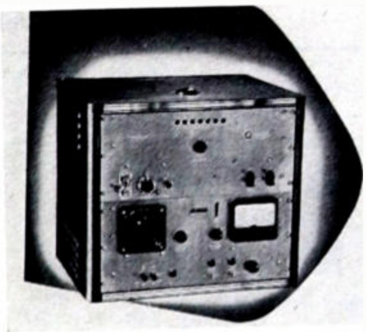


Sixteen models are available in the standard line ranging from units rated at 65 ma at 130 volts to units capable of handling 450 ma at 130 volts. Many special assemblies are possible for special applications.

Complete data and engineering information are available from the company.

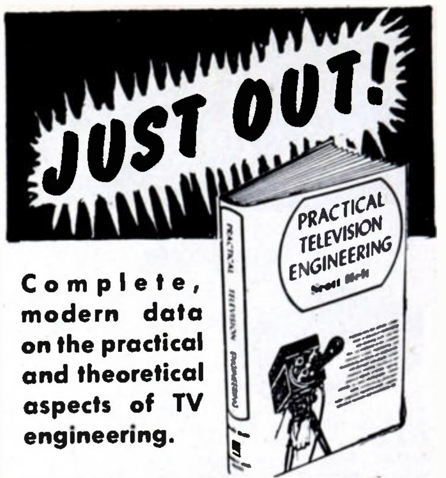
All-Purpose Nuclear Scaler

A new all-purpose scaler, which provides facilities for every type of counting within one instrument, has just been announced by **Nuclear Instrument & Chemical Corp.**, 229 W. Erie St., Chicago 10, Ill.



This new instrument, known as the "Ultra-Scaler" because of its versatility, is the only commercially manufactured instrument of its kind, the manufacturer claims. Geiger, proportional, or scintillation counting can be accomplished with this unit by either automatic or manual methods.

(Continued on page 70A)



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- Electron Tubes for Image Pickup
- Synchronizing Generators—Timing
- Shaping and Deflection Circuits
- The Video Amplifier and Cathode Follower
- Voltage-regulated Power Supplies
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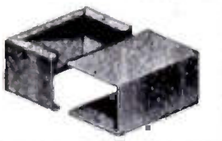
MINIBOXES



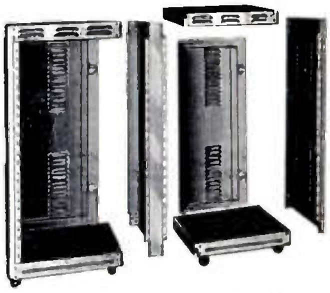
There are thousands of uses in the fields of radio and electronics for these new boxes. They are made from heavy gauge aluminum. The design of the box permits installation of more components than would be possible in the conventionally designed box of the same size. It is of two piece construction, each half forming three sides. The flange type construction assures adequate shielding. Available in etched aluminum finish and gray hammerloid finish.

Catalog Numbers Gray	Catalog Numbers Etched	Length	Width	Height	Dealer Cost
CU-2100	CU-3000	2¾"	2⅛"	1⅝"	$ .50
CU-2101	CU-3001	3¼"	2⅛"	1⅝"	.50
CU-2102	CU-3002	4"	2⅛"	1⅝"	.50
CU-2103	CU-3003	4"	2¼"	2¼"	.70
CU-2104	CU-3004	5"	2¼"	2¼"	.72
CU-2105	CU-3005	5"	4"	3"	.72
CU-2106	CU-3006	5¼"	3"	2⅛"	.72
CU-2107	CU-3007	6"	5"	4"	.81
CU-2108	CU-3008	7"	5"	3"	1.05
CU-2109	CU-3009	8"	6"	3½"	1.68
CU-2110	CU-3010	10"	6"	3½"	1.80
CU-2111	CU-3011	12"	7"	4"	2.34
CU-2112	CU-3012	17"	5"	4"	2.76
CU-2113	CU-3013	10"	2"	1⅝"	.78
CU-2114	CU-3014	12"	2½"	2¼"	.96
CU-2115	CU-3015	4"	2"	2¾"	.60
CU-2116	CU-3016	4¼"	2¼"	1¼"	.66

Prices 10% higher west of the Mississippi River

ADD-A-RACK SERIES

It has always been necessary to buy special racks without louvers on one side to obtain a maximum of panel space with a minimum of floor space. Now, you no longer need to buy a whole new cabinet when you want additional panel space. Through our new and exclusive Add-a-Rack series, BUD not only offers additional racks at a lower cost, but provides you with a sturdier, better looking assembly.



The illustration above at left shows two Add-a-Rack cabinets assembled together. The illustration above at right shows the unique and ingenious method of adding a unit to your present equipment. Instead of buying an entire new outfit, you purchase only four parts: (1) a door, (2) a top, (3) a bottom and (4) an Add-a-Rack coupling unit. The right (or left) hand side of your present relay rack is removed and replaced by the Add-a-Rack coupling unit; next, a top and bottom is fastened into place, and the side taken from the first rack is fastened onto the second rack which has been added. Place the additional door into position and you have two racks properly and efficiently coupled together.

In the same simple way, more racks can be added at any time and every one will be in a **CONTINUOUS ONE-PIECE** assembly.

This series is available in two ways: (1) a double unit consisting of two racks and the Add-a-Rack coupling unit, (2) Add-a-Rack unit, consisting of a door, a top, a bottom and an Add-a-Rack coupling unit. These units are furnished with all necessary assembling and panel mounting hardware.

BUD RC-7756 Casters will fit this unit. Casters are not included in price of cabinet.

Add-a-Rack Unit	To Add-a-Rack to	Overall Height	Panel Space	Dealer Cost
AR-1778	CR-1774	46 1/16"	36¾"	$26.25
AR-1775	CR-1771	47 5/16"	42"	32.50
AR-1776	CR-1772	66 9/16"	61¼"	40.75
AR-1777	CR-1773	82 5/16"	77"	48.00

Complete unit, consisting of the knocked-down parts necessary for two relay racks coupled together.

		Dealer Cost
CR-1779	two coupled relay racks same size as CR-1774	$54.75
CR-1780	two coupled relay racks same size as CR-1771	67.95
CR-1786	two coupled relay racks same size as CR-1772	83.05
CR-1799	two coupled relay racks same size as CR-1773	98.40

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BUD RADIO, INC.

2110 EAST 55TH STREET
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CONTINENTAL NOBLELOY
METAL FILM
NF RESISTOR

- Low Temperature Coefficient
- Range; 1 Ohm to 1 Megohm
- Axial Leads; ½ Watt
- Small Size ½" long x .150" Dia.
- Tolerance; + — 1% & 5%

Continental type NF "Nobleloy" resistors were designed to meet the needs of miniature, stable, precision resistors in critical applications.

Write for further details.

CONTINENTAL CARBON INC. CLEVELAND 11, OHIO

News—New Products

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

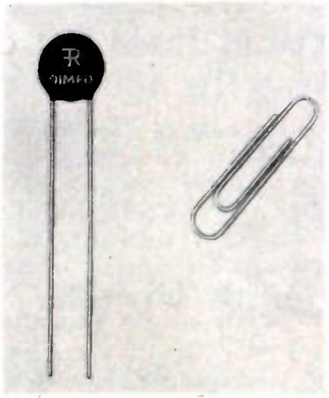
(Continued from page 69A)

A built-in timer is provided which may be set for a predetermined length of time and will then turn off the scaling unit automatically. The instrument may also be set for a predetermined number of counts, and the scaling unit will stop at that number of counts and indicate the time on the timer.

The "Ultra-Scaler" incorporates two inputs, one for Geiger pulses and the other for very small proportional pulses requiring linear amplification. The Higinbotham-type scale of 128 has a resolution time of only two microseconds.

Midget Ceramic Capacitor

A 0.01 μfd disk Ceramicon in a new small size, only 19/32 inch in diameter, is now being manufactured by **Erie Resistor Corp.**, 640 W. 12 St., Erie, Pa. This midget Ceramicon capacitor provides easy application in small spaces and simplifies soldering operations.



Capacity of the new Ceramicon is 0.01 μfd+100 per cent−0 per cent. Voltage rating is 400 volts dc, which is based on a life test of 800 volts dc at 85° C for 1,000 hours. The power factor is 2.5 per cent maximum at 1 kc at not more than 5 volts rms. Insulation resistance is 7,500 megohms minimum.

This disk Ceramicon is insulated with red dipped phenolic.

9 Pin Tube Socket

Mycalex Tube Socket Corp., operating under exclusive license of **Mycalex Corp.** of

America announces the addition of a 9 pin NOVAL) miniature tube socket to their line. Sockets are injection molded MYCALEX.

(Continued on page 71A)

News—New Products

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

(Continued from page 70A)

The sockets are obtainable in Mycalex 410 which was developed for applications requiring close dimensional tolerances not possible in ceramics and at much lower loss factor than mica filled phenolics with advantage in economy, the manufacturer claims, and in Mycalex 410X which has been developed to compare favorably with general purpose bakelite in economy but with a loss factor of only about one-fourth of that material.

These sockets are manufactured to precise specifications and fully meet RMA standards. Further information is obtainable from Mycalex Tube Socket Corp., 30 Rockefeller Plaza, New York 20, N. Y.

HF Inductor Alternator

A new "Inductor Alternator" which provides mobile high frequency ac power from a dc source, was recently announced by the **Carter Motor Co.**, 2644 N. Maplewood Ave., Chicago 47, Ill.

Basic in design, this new rotary power supply is suitable for aircraft, geophysical, government, and laboratory research, and other applications where a small mobile source of 400 to 800 cps of 100 watts or less continuous power is desired.

The ac output is obtained from the rotary inductor principle, eliminating conventional slip rings and brushes. The ac rotor and field are electrically isolated from the dc motor end. No permanent magnets are used for power generation. The ac output is not affected by the influence of strong magnetic fields, vibration, or aging. A separate dc output is also available, in addition to the ac output. While designed primarily for 24- to 29-volt airborne equipment, any input voltage from 5.5 to 230 volts dc is available on special order. For details, request Bulletin #350.

Stand-off Ceramic Capacitor

The **Erie Resistor Corp.**, 640 W. 12 St., Erie, Pa., is offering a new ceramic capacitor which has been designed especially for high frequency decoupling.

(Continued on page 72A)

MR. ENGINEER..

TRIAD TRANSFORMERS

will meet and maintain the most exacting requirements of your equipment

AVAILABLE FROM STOCK

Triad "HS" Series Transformers are engineered to meet precisely and maintain indefinitely the most exacting requirements of any industrial electronic application. Maximum protection against failure of your product is assured through:

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- Conservative Ratings
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- Exceptional Electrical Characteristics
- Long, trouble-free life

Triad "HS" Series Transformers are carried in stock for immediate delivery by Triad distributors or at the factory. **Write for Catalog TR-49A.**

Also readily available in standard designs for most used circuits are Triad Geoformers (Geophysical Transformers). **Write for Catalog GP-49.**

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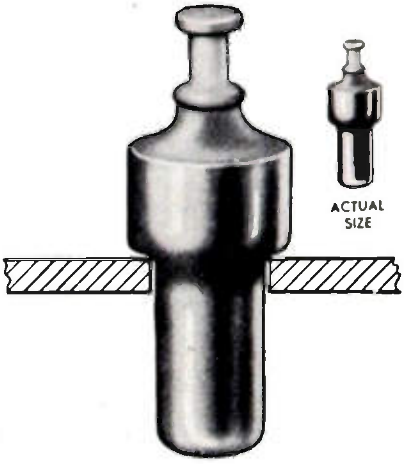
News—New Products

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

(Continued from page 71A)

Its applications will be in the field of television tuners, receivers, and other high-frequency equipment.



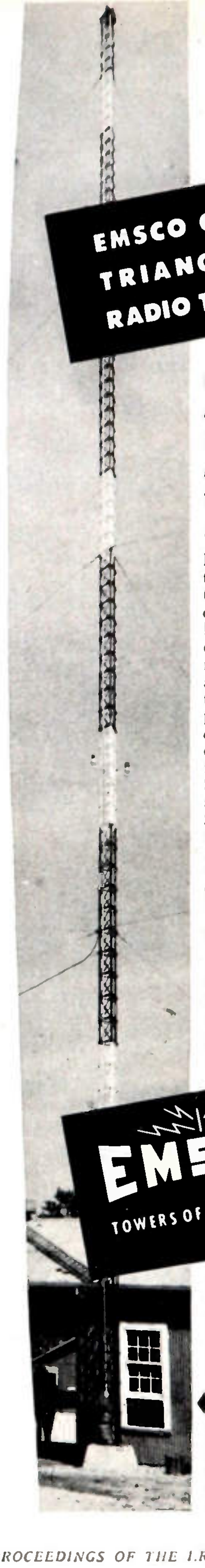
"Type 325 Stand-off Ceramicon" has a hermetically sealed case, and the company claims that it provides a by-pass to ground through the shortest possible path. Full advantage is taken of the concentric cylindrical electrode configuration in maintaining this short path, with resultant extremely low series inductance and effective vhf by-pass. Electrical shielding is provided by means of the grounded case. Standard capacitance 1,500 $\mu\mu$f ± 20 per cent, 500 volts working.

Special Coils and Chokes

A new line of coils and chokes adaptable to "tailor-made" specifications has been introduced by the **Shallcross Manufacturing Co.**, Collingdale, Pa.

Types include high-Q rf chokes, progressively-wound slug-tuned broadcast coils, and oscillator coils, all of them having out-of-the-ordinary characteristics which cannot be matched by standard coil types. Shallcross rf chokes may be made up as two separate coils having a specified coupling coefficient. High-premeability iron cores are sometimes used to provide greater inductance in a small unit.

(Continued on page 73A)



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Emsco guyed triangular radio towers are engineered for high wind load capacity, low maintenance costs and perfect transmission pattern . . . for *all* types of communication . . . in *all* industries. Standard towers available with 20½", 3', 5' or 14' faces, heights to 1000 feet and to withstand 20, 30 or 40 pound wind loads.

FREE BULLETIN

New Emsco Bulletin F-173 gives complete information on Emsco guyed triangular radio towers and Emsco free-standing square and triangular towers. Write for your free copy today!

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TOWERS OF STRENGTH

EMSCO DERRICK & EQUIPMENT COMPANY
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Shown here is an Emsco 20½-inch face, 160-foot Type IRT Emsco radio tower with 30-pound wind load rating.

POWER-PACKED FOR HEAVY DUTY

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5-SECOND HEATING

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LONGER REACH

DUAL HEAT
—single heat
200 watts,
dual heat
200/250 watts;
115 volts,
60 cycles

New WELLER 250-watt Soldering Gun

Heavy jobs and light jobs—the new 250-watt Weller Soldering Gun speeds them all. Chisel-shaped RIGID-TIP provides more soldering area for faster heat transfer. New "over-and-under" terminal design gives bracing action to tip. Your Weller Gun does delicate or heavy soldering with equal efficiency; compact and lightweight, it gets into the tightest spots.

Weller Guns actually pay for themselves in a few months. Fast 5 second heating means no time lost. Trigger-switch control means no current wasted—no need to unplug gun between jobs. Prefocused spotlight and longer length let you *see* the job and *reach* the job with ease. No other soldering tool offers so many time-and-money-saving features. Order your new 250-watt Weller Gun from your distributor today, or write for bulletin direct.

SOLDERING GUIDE

Get your copy of "SOLDERING TIPS"—new fully illustrated 20 page booklet of practical soldering suggestions. Price 10c at your distributor's or order direct.

WELLER MANUFACTURING COMPANY
821 Packer Street, Easton, Pa.

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 72A)

Sensitive DC Amplifier

A new and improved dc amplifier, Model 10, of the General Motors breaker type, which offers many advantages in the measurement of dc microvolt regions, is in production by **Liston-Folb, Div. of Atlas Coil Winders Inc.**, Dept. M, Box 1334, Stamford, Conn.

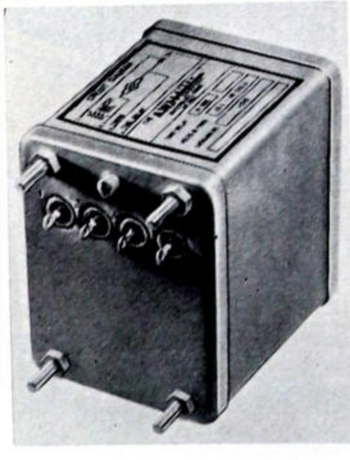
It is useful for the amplification of low-level thermocouple voltages, infrared detectors, photovoltaic cells, and the like. It can be used to replace suspension galvanometer systems.

This new amplifier features high immunity to the effects of ac pickup in the input circuit. The discrimination ratio against 60 cps pickup is over 1,000. It has an improved life breaker.

This instrument has a zero stability of better than 0.005 microvolt per day after warm up. The noise level approaches the limit imposed by the Johnson noise of the external circuit. This amplifier is available for operation with input circuits from 0 to 1 megohm. The dc output of the amplifier is sufficient to operate standard recorders, milliampmeters, and dc relays.

Hermetic Sealing For Timers

Hermetically sealed enclosures are now furnished for timers manufactured by the **A. W. Haydon Co.**, 232 N. Elm St., Waterbury 20, Conn.



(Continued on page 74A)

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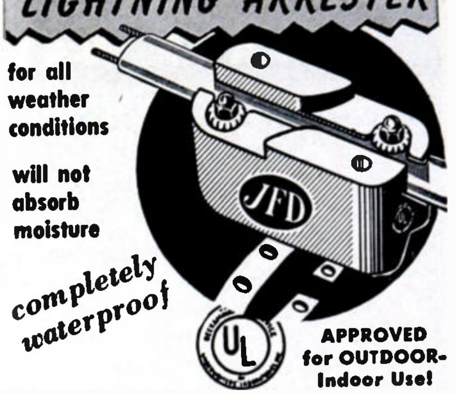
for all weather conditions

will not absorb moisture

completely waterproof



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Protects Television Sets Against Lightning and Static Charges

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Fits Any Type of Twin Lead

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No. AT103 Also for Tubular Twin Lead
BOTH Models Conform With Fire Underwriters and National Electrical Code Requirements for OUTDOOR Installations.

$2.25 EACH

SIMPLE TO INSTALL . . . For maximum efficiency, arrester should be mounted outside window nearest to TV receiver, with ground wire attached to nearest grounded point. No stripping, cutting or spreading of wires necessary. Supplied complete with 4 ft. length of Ductile Aluminum Ground Wire for Wall Mounting, and Strap for Mast or Grounded Pipe installation.

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The World's MOST-LISTENED-TO magnetic tape RECORDER

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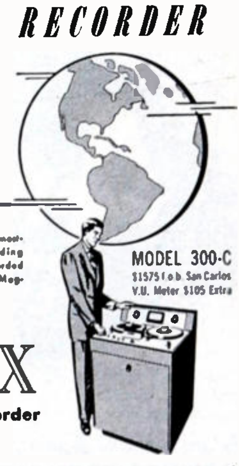
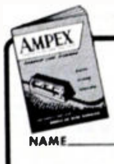
Fully ninety percent of the most-listened-to programs, including "Voice of America", are recorded and played-back on Ampex Magnetic Tape Recorders.



MODEL 300-C
$1575 f.o.b. San Carlos
V.U. Meter $105 Extra

AMPEX
Magnetic Tape Recorder



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AMPEX ELECTRIC CORP., San Carlos, California
Without obligation please send 16-page illustrated booklet containing technical specifications of Ampex Magnetic Tape Recorders.

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SQUARE WAVE GENERATOR

Continuously Variable, 25 CPS-1 MC
Rise Time, .02 Microseconds
Direct Reading Frequency Meter
Versatile Output Circuit

TEKTRONIX Type 105
Square Wave Generator
Price $395.00 f.o.b.
Portland, Oregon

Square wave testing techniques come into wider use as the need for good transient response in wide band amplifiers becomes increasingly important. In order to test the high frequency response it is necessary to have a signal which has a rise and fall at least equal to and preferably faster than the risetime of the amplifier being tested. In addition to a sharp rise and fall, the test signal should be free of over-shoot and other spurious responses. For examination of the low frequency response a square wave signal having flat horizontal portions is needed.

The TEKTRONIX **Type 105** Square Wave Generator provides a suitable signal for both of the above tests. Its frequency range, extending continuously from 25 cycles to 1 mc., combined with its risetime of .02 microseconds, makes it possible to quickly and accurately test amplifiers, filters, etc., having pass bands from a few cycles to 20 mc.

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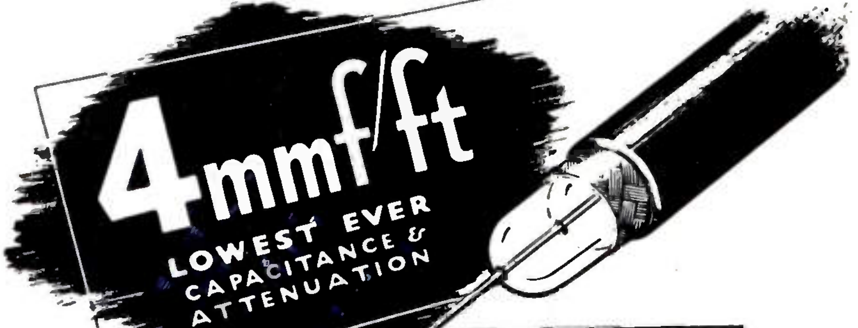

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FOR RADIO FREQUENCIES

LOW ATTEN. TYPES	IMPED. OHMS •	ATTEN. db 100 ft at 100 Mc/s	LOADING K.w at 100 Mc/s	O.D." •
A.1	74	1.7	0.11	0.36
A.2	74	1.3	0.24	0.44
A.34	73	0.6	1.5	0.85

PHOTOCELL CABLE

VERY LOW CAPACITANCE CABLES

FOR VIDEO and SPECIAL APPLICATIONS

LOW CAPAC. TYPES	CAPAC. mmf/ft •	IMPED. OHMS •	ATTEN. db 100 ft 100 Mc/s	O.D." •
C 1	7.3	150	2.5	0.36
P.C 1	10.2	132	3.1	0.36
C.11	6.3	173	3.2	0.36
C.2	6.3	171	2.15	0.44
C.22	5.5	184	2.8	0.44
C.3	5.4	197	1.9	0.64
C.33	4.8	220	2.4	0.64
C.44	4.1	252	2.1	1.03

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News—New Products

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

(Continued from page 73A)

These new timers are suited to any application involving corrosive atmospheres or high humidity as encountered in aircraft, ships, and mines.

All enclosures are evacuated to 100 microns and filled to one atmosphere with dry nitrogen so that full switch ratings can be used even at extremely high altitudes where contact capacity is normally reduced. Deterioration of motor brushes is completely eliminated.

Crystal Calibrator

Measurements Corporation, Boonton, N. J., has announced the production of their new Model 111 Crystal Calibrator. This instrument was designed for the frequency calibration of signal generators, transmitters, receivers, grid-dip meters, and other equipment in the range of 250 kc to 1,000 Mc. The frequency accuracy is ±0.001 per cent.

The Model 111, a dual-purpose calibrator, not only provides a test signal of crystal-controlled frequency, it also has a self-contained receiver with a sensitivity of 2 microwatts.

A new circuit arrangement utilizes the cross-modulation products of three separate oscillators operating at the fundamental frequencies of 0.25, 1.0, and 10 megacycles. This system extends the usable range of harmonic frequencies far beyond that of previously available equipment.

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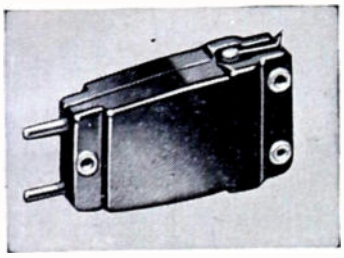
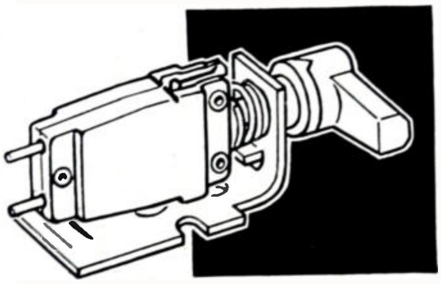
Webster Electric

Model "A" Cartridge



with Twist Mechanism



A complete unit with top performance and absolute minimum of service and installation problems.

The twist mechanism is factory assembled with Model A7 cartridge in place, ready for installation in tone arms without adjustment or modification. This completely assembled unit gives positive tracking at all playing speeds. High vertical and lateral compliance eliminate "skating". The simple, foolproof twist mechanism gives positive indexing, eliminating the possibility of twisting and damaging the leads in the tone arm.

There are no delicate parts to break or get out of order. The Model A7 with twist mechanism reverses through a 180 degree arc for playing either 33⅓ — 45 or 78 R.P.M. records.

Send for a sample assembly today . . . try it . . . then note first hand the advanced improvement.

WEBSTER ELECTRIC

Webster Electric Company, Racine, Wis. • Established 1909

"Where Quality is a Responsibility and Fair Dealing an Obligation"

S·S·White MOLDED RESISTORS

The "All-Weather" Resistors

Of particular interest to all who need resistors with inherent low noise level and good stability in all climates

TYPE 65X

Actual Size

STANDARD RANGE

1000 OHMS TO 9 MEGOHMS

Used extensively in commercial equipment including radio, telephone, telegraph, sound pictures, television, etc. Also in a variety of U. S. Navy equipment.

HIGH VALUE RANGE

10 to 10,000,000 MEGOHMS

This unusual range of high value resistors was developed to meet the needs of scientific and industrial control, measuring and laboratory equipment—and of high voltage applications.

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It gives details of both the Standard and High Value Resistors, including construction, characteristics, dimensions, etc. Copy with Price List mailed on request.

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MOLDED PLASTICS PRODUCTS—MOLDED RESISTORS

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Features

- Low Loss
- High KVA Rating
- Shielded From External Electrostatic Fields
- Low Internal Distributed Inductance
- Complete Dependability

Use of a gas dielectric under pressure permits high voltage ratings and large values of capacity in a small volume of space, yet all the advantages of air dielectric capacitors are retained. Construction prevents erratic performance due to changes of barometric pressure or humidity as well as excluding all foreign matter which could cause flashovers. In contrast to comparable solid dielectric capacitors, permanent damage to JOHNSON pressurized capacitors from flashovers is improbable.

JOHNSON designed and built pressurized capacitors are available in fixed, variable and semi-variable types. Capacity values to 10,000 mmf., voltage ratings to 32,000 volts peak and currents from 40 to 80 amperes are available in standard units. Special units with even higher voltage and current ratings can be supplied.

Plates are polished aluminum with rounded edges. Shells are copper plated steel; insulation steatite. Seals are corprene which is impervious to moisture and oil, is stable and does not deteriorate with age. Dielectric is 200 P.S.I. oil pumped nitrogen.

The reliable performance of JOHNSON pressurized capacitors is due to conservative design and excellent workmanship. Complete dependability is assured.

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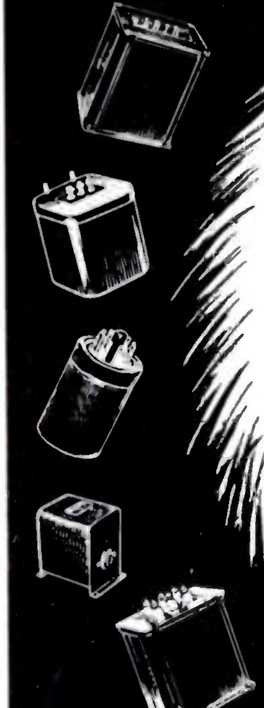
JOHNSON *. . . a famous name in Radio*

E. F. JOHNSON CO. WASECA, MINNESOTA



KENYON *Fits Your Production To A "T"*

KENYON "T's"—high quality, uniform transformers, are your best bet for development, production and experimental work. For over 20 years, the KENYON "K" has been a sign of skillful engineering, progressive design and sound construction.

Now—reduce inventory problems, improve deliveries, maintain your quality—specify KENYON "T's," the finest transformer line for all high quality equipment applications.

New Catalog Edition! Write Today!

KENYON new modified edition tells the complete story about specific ratings on all transformers. Our standard line saves you time and expense. Send for your copy of our latest catalog edition *now!*

KENYON TRANSFORMER CO., Inc.
840 BARRY STREET · NEW YORK 59, N. Y.

CONTINUOUSLY
VARIABLE FILTERS
MODEL 302
VARIABLE ELECTRONIC FILTER

"Another SKL first"

The — SKL — Model 302 includes two independent filter sections, each having a continuously variable cut-off range of 20 cps to 200 KC. Providing a choice of filter types each section has 18 db per octave attenuation. When cascaded 36 db is obtained in the high and low pass setting and 18 db in the band pass position. With low noise level and 0 insertion loss this versatile filter can be used as an analyzer in industry and the research laboratory or to control sound in the communications laboratory, radio broadcasting, recording and moving picture industries.

SPECIFICATIONS

- CUT-OFF RANGE 20 cps to 200 KC
- SECTIONS 2—can be high, low and band pass
- ATTENUATIONS 36 db/octave maximum
- INSERTION LOSS . 0 db
- NOISE LEVEL 70 db below 1 volt
- FREQUENCY RESPONSE 2 cps to 2 MC

SKL SPENCER-KENNEDY LABORATORIES, INC.
181 MASSACHUSETTS AVE., CAMBRIDGE 39, MASS.

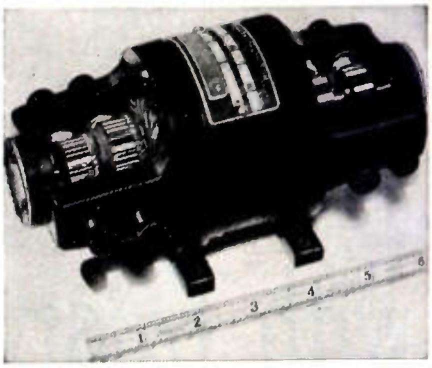
News—New Products

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

(Continued from page 74A)

Dynamotor Power Supply

Carter Motor Co., 2644 N. Maplewood Ave., Chicago 47, Ill., announces the Multi-Magmotor, which was designed to provide reliable rotary power for applications demanding multiple input, multiple output, or both.



Because field flux remains constant regardless of input or output voltage, single, dual, or even triple input voltages are possible without affecting the field circuit. One, two, or three dc output voltages are also available, or many other combinations of input and output arrangements not exceeding the maximum 4 commutator design.

All windings are insulated from each other, and the frame. Open construction permits greater outputs and higher continuous duty ratings.

Heavy-Duty Inverter

American Television & Radio Co., 300 E. Fourth St., St. Paul, Minn., announces a complete line of heavy-duty inverters which will supplement the existing ATR lines of inverters.

The new inverter line is for operation on dc input voltages ranging from 6 volts to 220 volts dc, having output of 110 volts 60 cps at capacities ranging from 175 watts to 1,000 watts.

This line of inverters was designed especially for heavy-duty applications including tape recorders, television sets, portable transmitters, and similar electronic and electrical equipment within the output capacity ranges indicated.

(Continued on page 77A)

Spectrum Analysis

from AF to UHF

***Faster and Simpler with these* Panoramic Instruments**

Whether your problem is investigation of noises, vibrations, harmonics, characteristics of AM, FM or pulsed signals, oscillations, cross modulation, transmission characteristics of lines and filters, telemetering or any phenomena requiring spectrum analysis, these panoramic instruments will help collect information faster, **easier and accurately.** Panoramic instruments automatically visualize spectral content. Indications in the form of vertical pips whose height and horizontal location respectively show signal level and frequency, enable rapid examination of one or more signals at one time.

AP-1

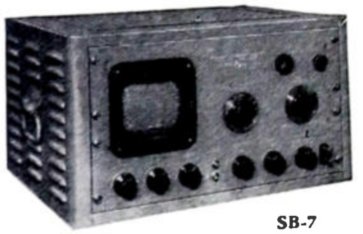
PANORAMIC SONIC ANALYZER AP-1
Complete Audio Waveform Analysis in One Second

Recognized as THE practical answer for analyzing waves of random or static character, the AP-1 automatically separates and measures complex wave components in only one second.

Frequency Range: 40-20,000 c.p.s., log scale.

Input Voltage Range: 500 μV-500V.

Voltage Scale: linear and two decade log, harmonic products suppressed by at least 60 **db. Direct Reading. Simple Operation. Optimum Resolution.**



SB-7

PANORAMIC ULTRASONIC ANALYZER
Entirely New for Ultrasonic Studies

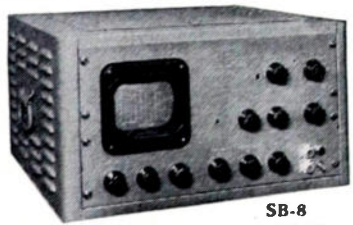
An invaluable new direct reading instrument, the SB-7 enables overall observation of the ultrasonic spectrum or very highly detailed examination of any selected narrow segment of the spectrum.

Frequency Range: 2KC-300KC, linear scale.

Scanning Width: Continuously variable, 200 KC to zero.

Input Voltage Range: 1 mV-50V.

Amplitude Scale: linear and two decade log.



SB-8

PANALYZOR—PANADAPTOR
For RF Spectrum Analysis

Long accepted as the simplest and fastest means of observing segments of the RF Spectrum, Panadaptor units operate with superheterodyne receivers, which tune in the segment to be examined. Panalyzors use an external signal generator for this purpose and have a flat amplitude response for determining relative levels of signals.

Both are available in over a dozen standard models and types differing in . . . **Maximum Scanning Widths** ranging from 50KC to 20MC, continuously variable to zero. **Signal Resolution Capabilities from** 250KC down to 100 CPS.

Write for Complete Technical Data

PANORAMIC RADIO PRODUCTS Inc
10 SOUTH SECOND AVE MOUNT VERNON N.Y.

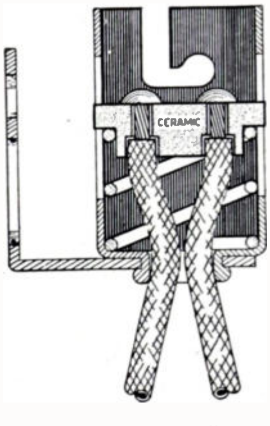
News—New Products

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)

Lamp Sockets with Ceramic Insulation

The **Dial Light Co. of America, Inc.,** 900 Broadway, New York 3, N. Y., announces the availability of sockets of the new NE 12 series intended for mounting lamps with double contact bayonet bases. They are listed by the Underwriter's Laboratories, Inc., and are especially suitable for the new 10C7DC pilot light lamp and bayonet base appliance lamps.



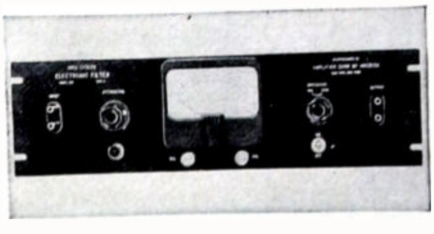
These sockets have flexible wire leads for circuit connections. A great variety of mounting bracket forms permits the design to adapt them to any situation.

The wire leads terminate inside the socket shell in contacts mounted in a ceramic disk. Recesses in the under side permit the wire covering to enter the disk so that no live metal is exposed.

The ceramic disk insures effective insulation of these sockets even when hot lamps are employed in them. When high ambient temperatures are anticipated, asbestos covered wire leads may be specified for maximum resistance to the effects of heat.

Electronic Filters and Tuned Audio Amplifiers

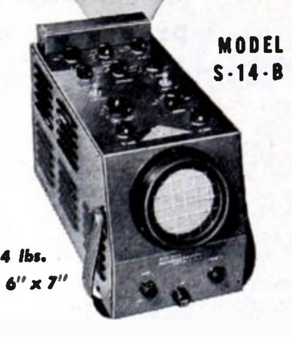
A new series of electronic filters and tuned audio amplifiers specifically designed for all audio and specialized electronic fields, involving single frequency circuits, is now in production at the **Instrument Div., Amplifier Corp. of America,** 398-1 Broadway, New York 13, N. Y.



Six standard instruments are available in each series, providing pass frequencies and tuned frequencies of 400, 500, 1,000,

(Continued on page 79A)

THE WIDE BAND POCKETSCOPE®
BY WATERMAN



Another Waterman POCKETSCOPE confirming the obsolescence of conventional oscilloscopes. Characterized by wide band amplifier fidelity without peaking as well as amazing portability. S-14-B POCKETSCOPE is ideal for laboratory and field investigation of transient signals, aperiodic pulses, or recurrent electrical wave forms.

Vertical channel: 50mv rms/inch, with response within −2DB fram DC to 700KC, and pulse rise of 0.35μs. Horizontal channel: 0.3v rms/inch with respanse within −2DB from DC to 200KC, and pulse rise of 1.8μs. Non-frequency discriminating attenuatars and gain controls, with internal calibration of trace amplitude. Repetitive or trigger time base, with linearization, from ½cps to 50KC, with ± sync. or trigger. Trace expansion. Filter graph screen. Mu metal shield. And a host of other features.

WATERMAN PRODUCTS CO., INC.
PHILADELPHIA 25, PA.
CABLE ADDRESS: POKETSCOPE

WATERMAN PRODUCTS INCLUDE:

S-10-B GENERAL POCKETSCOPE
S-11-A INDUSTRIAL POCKETSCOPE
S-14-A HI-GAIN POCKETSCOPE
S-15-A TWIN TUBE POCKETSCOPE
S-21-A LINEAR TIME BASE

Also RAKSCOPES, LINEAR AMPLIFIERS, RAYONIC® TUBES and other equipment

WATERMAN PRODUCTS

Latest WELLS Tube Price List

Many Types Are Now Scarce At These Low Prices. Check your requirements at once for your own protection. All tubes are standard brand, new in original cartons, and guaranteed by Wells. Order directly from this ad or through your local Parts Jobber.

TYPE.	PRICE EA.
0A4G	.95
01A	.45
EI-C1A	3.95
1A3	.60
1A5GT	.65
C1B/3C31	1.85
1B4P	1.75
1B21A/GL471A	2.55
1B22	3.40
1B23	7.50
1B32/532A	1.85
1B42	5.25
1B48	9.90
EL1C	4.85
1C5GT	.65
1C6	.75
1C7G	.85
1D8GT	.95
1E7GT	1.95
1F4	.90
1G6	.65
1H4G	.80
1H6G	.80
1L4	.50
1LC6	.75
1LN5	.85
1N5GT	.75
1P24	2.50
1Q5GT	.85
1R4	.55
1S5	.70
1T4	.75
2A3	1.05
2A7	.75
2AP1	4.75
2B7	.75
2B22/GL559	3.25
2C22/7193	.35
2C26	.30
2C26A	.40
2C34	.40
2C44	1.25
2J21	10.45
2J21A	11.45
2J22	9.85
2J26	8.45
2J27	12.95
2J31	9.95
2J32	12.85
2J33	18.95
2J34	17.50
2J37	13.85
2J38	6.95
2J48	12.95
2J61	24.50
2J62	14.95
2X2	.55
2Y3G	1.20
3A4	.35
3A4/17	.45
3B7	.45
3B22	2.35
3B24	1.75
3BP1	3.75
EI-3C	3.95
3C21	5.00
3C24/24G	.50
3C31-C1B	4.85
3CP1-S1	1.95
3DP1	3.75
3D6/1299	.45
3FP7	1.85
3FP7A	4.95
3GP1	4.50
3H17	1.00
3HP7	2.95
3Q5	.90
3Q5GT	.90
3S4	.75
GA4	2.00
REL5	14.95
VT5/215A	1.20
5AP1	3.95
EI-C5B	4.25
5BP1	2.75
5BP4	3.95
5CP1	3.75
5D21	24.75
5FP7	2.75
5GP1	2.75
5HP4	4.75
5J23	14.25
5J29	13.45
5V4G	1.07
5W4	.76
5Z3	.80
5Z3G	.75
6-4	.35
6-7	.35
EI-C6A	2.00
6A3	.80
6A6	.65
6AB7	.95
6AC7	.90
6AE6G	.85
6AG5	1.20
6AH6	1.10
6AK5	1.20
6AK6	.80
6AL5	.95
6AU6	.95
6AV6	.81
6B4G	.95
6B7	.75
6B8	.95
6B8G	.95
6BA6	.95
6BE6	.65
6C4	.40
6C8G	1.05
6C21	19.25
6D6	.50
6F5	.85
6F6	.60
6F6G	.60
6F8G	.95
6G6G	.80
6H6	.45
6J6	.90
6J7GT	.70
6J8G	.95
6K6GT	.55
6K7	.80
6K7G	.80
6L6G	1.35
6L7	.75
6N7	.75
6N7/GT	.75
6Q7	.55
6R7G	.75
6SA7	.65
6SC7	.75
6SC7GT	.70
6SF7	.80
6SF5	.65
6SG7	.65
6SH7	.40
6SH7GT	.40
6SJ7	.60
6SJ7GT	.60
6SK7	.60
6SK7GT	.60
6SL7GT	.60
6SN7GT	.85
6SQ7	.60
6SQ7GT	.60
6SR7	.60
6SR7GT	.60
6SS7	.60
6U7G	.85
6V6GT	.75
6W5G	.80
6X5GT	.73
6Y6G	.65
7-7-11	.35
7A4/XXL	.60
7A5	.80
7A6	.75
7A7	.60
7B4	.60
7B6	.60
7B8	.60
7BP7	4.95
7C4/1203A	.35
7C5	.65
7C7	.65
7E5/1201	.60
7E6	.60
7F7	.70
7H7	.70
7L7	.70
7N7	.70
7T7	.90
7Y4	.90
9-3	.45
10	.55
10 ACORN	.65
10 (VT-25A)	.55
10E/116	1.00
10T1	.60
10Y (VT-25)	.45
12A6	.25
12A6GT	.25
12A7	.80
12AH7GT	1.12
12AU7	.96
12AX7	1.20
12BD6	1.20
12C8	.50
12F5GT	.65
12H6	.40
12J5GT	.40
12J7GT	.70
12K8	.65
12Q7GT	.75
12SA7	.73
12SC7	.75
12SF7	.60
12SG7	.65
12SH7	.40
12SJ7	.73
12SK7	.60
12SK7GT	.60
12SL7GT	.60
12SN7GT	1.10
12SQ7GT	.60
12SR7	.60
12X825-2AMP. TUNG.	1.95
12Z3	.90
13-4	.35
14A7	.90
14B6	.75
14F7	.90
14H7	.90
14Q7	.90
14R7	.90
15E	1.50
15R	1.20
16X879-2AMP. TUNG.	1.95
FG17/967	3.25
19	1.20
20-4 BALLAST.	.45
21-2 BALLAST.	.45
REL21	2.75
23D4	.45
RK24	1.75
24A	.75
RK25/802	2.85
VT-25-A/10	.55
25Z5	.73
25Z6GT	.55
25Z6G	.55
26	.65
27	.50
28D7	.40
30/VT-67	.75
30 (NOT VT-67)	.75
33	.75
34	.35
GL34	1.50
RK34/2C34	.45
35/51	.60
35L6GT	.73
35W4	.73
35Z5GT	.62
36	.40
37	.40
38	.40
39/44	.35
41	.55
42	.50
43	.50
45 SPEC.	.50
46	.75
EF50	.45
UH50	1.00
50B5	1.00
50L6GT	.96
56	.65
57	.45
58	.50
RK60/1641	.65
VT62 (BRITISH)	1.10
65	1.25
6684	.90
VT67/30	.90
70L7	1.05
71A	.75
CEQ72	1.50
CRP72	.95
CYN72	1.75
RKR72	.90
RKR73	1.25
75	.89
76	.55
77	.55
78	.60
VR78	.65
80	.45
FG-81-A	3.95
83V	.90
89	.73
89Y	.40
VR90	.65
VT90 (BRITISH)	2.55
VR92	.65
FG95/DG1295	9.95
VT98/REL5	14.95
100R	2.75
100TH	11.50
101/837	1.65
102F	3.55
FG105	9.75
VR105	.85
VU-111-S	.55
114B	1.20
121A	2.65
122A	2.65
VT127 (BRITISH)	.35
VT127A	2.95
VR150	.60
VT158 (HK)	14.95
FG172	19.75
205B	1.45
211 (VT-4-C)	.60
215A (VT5)	1.20
CEP220	2.00
221A	1.75
227A	4.75
231D	1.20
RX233A	1.95
250R	9.00
257A	3.00
268A	2.95
282B	1.25
287A/722A	9.50
304TL	1.75
304TH	5.75
307A	4.25
316A	.55
327A	2.60
350B	2.55
354C	14.95
356B	4.95
368AS/703A	3.95
371A/VT62	.95
371B	.85
388A	3.95
393A	4.65
395A	4.95
MX408U-BALLAST.	.30
417A	14.50
434A	3.40
446A	1.55
446B	1.55
450TH	17.95
GL451	1.90
GL471A	2.55
SS501	3.00
527	9.95
WL530	5.00
WL531	12.95
WL532	1.85
532A/1B32	3.55
GL559	3.75
KU610	7.45
HY615	1.05
WL632A	8.75
700	7.95
700B	7.95
700C	7.95
700D	7.95
701A	3.00
702A	2.95
703A/368AS	3.95
704A	1.75
705A	1.55
706AY	17.50
707A	14.00
707B	15.00
708A	3.75
709A	4.75
710A	2.45
713A	1.50
714AY	3.75
715B	9.75
717A	.85
718BY	15.00
718EY	15.00
721A	3.75
721B	3.95
722A/287A	9.50
723AB	14.95
724A	4.25
725A	9.95
726A	12.50
726B	13.50
730A	9.95
801	.50
801A	.70
803	5.25
804	8.95
805	5.95
807	1.25
808	1.65
809	2.50
812	2.95
813	7.85
814	3.75
815	2.85
826	.75
830B	3.95
832A	7.95
834	5.75
835/38111A	1.10
836	1.35
837	1.65
838	3.25
841	.50
842	2.75
843	.50
851	39.00
852	6.25
861	29.45
864	.45
865	2.55
866-JUNIOR.	.85
866A	1.30
869	19.75
869B	27.25
872A	2.45
874	1.95
876	.50
878	1.95
879/2X2	.55
902	3.50
923 (PHOTO)	1.35
930	1.05
931A	3.95
954	.35
955	.55
956	.55
957	.45
958	.55
958A	.55
959	.55
967/FG17	3.25
991/NE-16	.24
1005	.35
1007	4.50
1148	.35
1201	.45
1203	.55
1203A	.65
1294	.55
DG1295	9.95
1299/3D6	.45
1613-SELECT. 6F6	.55
1616	1.25
1619	.35
1624	1.25
1625	.35
1626	.35
1629	.35
1630	3.95
1638	.90
1641/RK60	.65
1642	.50
1852/6AC7	.90
1853/6AB7	.95
1960	1.35
1961/532A	1.85
2050	.78
2051	.75
UX6653	1.20
7193	.35
8011/VT90. BRITISH	2.55
8012	3.25
8013	1.25
8013A	1.50
8019	1.75
8020	3.25
8025	6.75
9001	.65
9002	.45
9003	.60
9004	.40
9006	.40
38111A/835	1.10

XTAL DIODES

TYPE.	PRICE EA.
1N21	.65
1N21A	.95
1N21B	1.20
1N22	.80
1N23	.80
1N23A	.85
1N27	.85
1N29	.85
1N51 (GE)	.75
1N48 (GE)	.75
1N52 (GE)	1.00

JUST OUT — CATALOG H500

Manufacturers, Distributors and Amateurs: Write for the brand new Wells Electronic Catalog H500. It's full of Tremendous values in highest quality components.

WELLS
SALES, INC.

320 N. LA SALLE ST. DEPT. P, CHICAGO 10, ILL.

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 77A)

2,000, 3,000, and 10,000 cps. Multituned amplifiers and multifilters utilizing switching networks for selecting any one of up to six predetermined frequencies are also available.

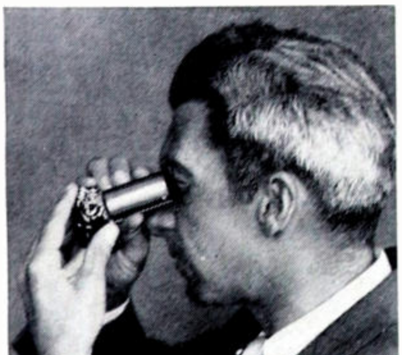
A dual looped network is employed, which consists of a combination of positive and negative feedback, arranged so that positive feedback is predominant at the pass frequency while negative feedback prevails throughout the remainder of the audio spectrum.

The tuned amplifiers include, in addition to the frequency selective network and output stage, a low-noise level input bridging stage which may be coupled to any impedance source under 500,000 ohms.

Where balanced input or output circuits are required, auxiliary input and output transformers may be added. The inclusion of the optional output indicating meter makes the electronic filters or the tuned amplifiers complete indicating instruments.

Alpha Counter

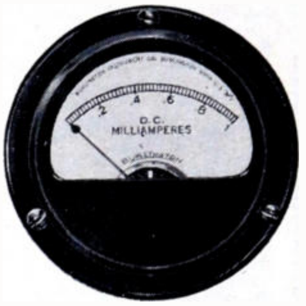
Nuclear Div., Clarkstan Corp., 11925 West Pico Blvd., Los Angeles 64, Calif., announces their Model 501 Alpha-Counter, a nuclear instrument for the detection of alpha particles from any source. All uranium and thorium ores may be qualitatively and quantitatively analyzed, when the radio active content is as low as a fraction of 1 per cent. The device is stable in operation so that it may be calibrated against known sources and samples.

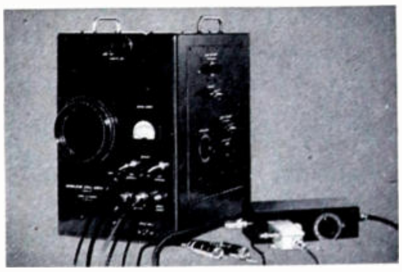


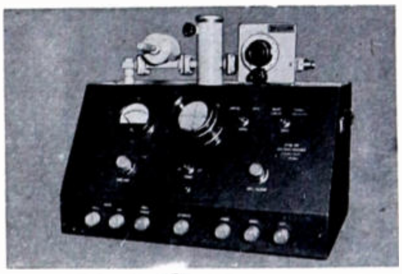
The counter utilizes a highly efficient optical system and a newly developed phosphor screen. From these it provides a statistical count of alpha particle scintillations. Unlike the spintharascope which must be used in the dark, the Model 501 may be used under subdued white light or red light illumination.

The counter comes complete with luminescence-quenching filter, radio-active samples, aluminum carrying case, and complete instructions.

(Continued on page 82A)







COMMUNICATIONS EQUIPMENT COMPANY

RADAR SETS
TEST EQUIPMENT

YD-2 MARKER BEACON EQUIP. Compl. Installation in Trailer w/Gas Generator—WRITE.

X BAND

Cross gd. direction Coupler 20 DB, Mtd on 90° bend	$14.50
90° bend H plane 4" Radius cover to cover	$8.00
Directional coupler, UG-40/U take off, 20 DB	$17.50
Directional coupler, APS-6, Type "N" take off, 20 DB, calibrated	$17.50
Broad Band Directional coupler, type "N" take off, choke to cover, 23 DB, calibrated	$18.50
Directional coupler, APS-31, type "N" take off, 25 DB	$17.50
Bi-directional coupler, type "N" take off	$22.50
Flexible Section 18" long	$12.00
Straight Sections 2½ ft. long choke to cover, silver plated	$6.50
Pressure Test Section with 15 lb. gauge and pressurizing nipple	$10.00
Bulk Head Feed Through, choke to cover	$12.00
Mitered Elbow, choke to cover or choke to choke	$12.00
Right Angle Bend 2½" Radius, choke to cover	$12.00
90° Twist, 6" long	$7.50
45° Twist, 6" long	$7.50
90° Twist, 5" long with pressurizing nipple	$7.50
15° Bend, 10", choke to cover	$4.50
5 ft. Sections UG-39 to UG-40, silver plated	$9.50
180° Bend, 26" choke to cover 2½" radius	$5.00
SWR Measuring Section 4" long, 2 type "N" probes mounted full wave apart 1¼" x ⅝" guide	$8.50
WE attenuator 0 to 20 DB, less cards, bell size guide	$12.50
90° Bend, E Plane 18"	$4.00
Rotary Joint, choke to choke	$10.00
Rotary Joint, choke to choke with deck mounting	$10.00
TR-ATR Duplexer Section for 1B24 and 724B	$12.50
Wavemeter-Therimstor MTG Sect.	$6.00
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
723AB Mixer—Beacon Dual Oscillator Mount with Crystal Holder. Used	$12.00
723AB Mixer—Beacon Dual Oscillator Mount with Matching Slugs and tunable termination, new	$24.50
Bi-Directional Couple, type "N" termination, 26 DB, calibrated, 1¼" x ⅝" guide	$24.50
12" Flexible Section, 1¼" x ⅝" guide	$10.00
Crystal Mount in Waveguide	$17.50
SO-3 Echo Box, Transmission type cavity with bellows	$28.50
180° Bend with pressurizing nipple	$5.00
"S" Curve 18" long	$5.00
"S" Curve 6" long	$3.25
APS-31 Mixer Section for mounting two K25's, Beacon Reference Cavity, 1B24 TR Tube	$42.50
Transition 1 x ½" to 1¼" x ⅝", 14" long	$8.00
Receiver Front End, complete, C/O Dual 723AB Klystron mount, TR-ATR Duplexer Section, 2 Stage 30 MC. Preamplifier, new, with all tubes	$59.50
Random Lengths of Waveguide 6" to 18" long	$1.00 per ft.

RADAR SETS

APS-2, Airborne, 10 CM, Major Units, New
APS-4, Airborne, 3 CM, Compl., Used
APS-15, Airborne, 3 CM, Major Units, New
SD-4, Submarine, 200 MC, Compl., New
SE, Shipboard, 10 CM, compl., New
SF-1, Shipboard, 10 CM, Compl., New
SJ-1, Submarine, 10 CM, Compl., Used
SL-1, Shipboard, 10 CM, Compl., Used
SN, Portable, 10 CM, Compl., Used
SQ, Portable, 10 CM, Comp., Used
SO-1, Shipboard, 10 CM, Compl., Used
SO-8, Shipboard, 10 CM, Compl., Used
Mark 4, Gunlaying, 800 MC, Less Ant., Used
Mark 10, Gunlaying, 10 CM, Compl., New
Less Rack, New. $1500; Less Rack, Used
CPN-3, Beacon, 10 CM, Major Units, Used
CPN-6, Beacon, 3 CM, Complete, New
CPN-8, Beacon, 10 CM, Complete New
Less Ant., New
SCR-533, IFF/AIR, 500 MC, New
Search Tracer Airborne Radar Altimeter, 500 MC Compl., New

WRITE OR PHONE FOR INFO. AND SPECS. MANY TYPES AVAIL.

COUPLINGS—UG-CONNECTORS

UG/15U	$.75	UG86U	$1.40		
UG206U	$.90	UG342U	$3.25		
UG87U	$1.25	UG85U	$1.45		
UG27U	$1.69	UG58U	$.60		
UG21U	$.89	UG9U	$.89		
UG167U	$2.25	UG102U	$.45		
UG29U	$.90	UG103U	$.45		
UG254U	$1.69	UG255U	$1.65		
UG 40/U	$.75	UG 52	$1.35	UG 55/U	$4.00
UG 40A	$1.10	UG 210	$1.85	UG 56/U	$4.75
UG 343	$2.35	UG 212	$2.40	UG 65/U	$6.50
UG 344	$3.00	UG 40U	$.70	UG 149/U	$3.00
UG 425	$2.00	⅞ Coax	$.50	UG 148/U	$4.00
UG 116	$1.95	⅞ Coax	$.95	UG 150/U	$3.00
UG 117	$2.50	UG 53/U	$4.00	UG 39/U	$.60
UG 51	$1.00	UG 54/U	$4.75	UG 40/U	$.80

TEST EQUIPMENT

CG-176/AP Directional coupler X Band, 20 DB nominal, type "N" take off, choke to choke, silver-plated	$17.50
X Band 1⅜" x ⅝" absorption type wavemeter, micrometer head, 6000 to 8500 mc. Demornay-Budd #35B	$185.00
C Band "T" gold-plated at	$97.00
C Band Flap attenuator Demornay-Budd type #339, gold-plated	$100.00
X Band 1⅜" x ⅝" Klystron mount with tunable termination, gold-plated	$75.00
X Band 1⅜" x ⅝" low power load, gold-plated	$45.00
X Band 1⅛" x ½" waveguide to type "N" adaptor, gold-plated	$22.50
X Band 1⅛" x ½" "T" Section, gold-plated	$55.00
DEHYDRATOR UNIT. CPD 10137 Automatic Cycling, Compressor to 50 lbs., Compl. for Radar XSMN. Line. New	$425.00
H.V. PWR. SUPPLY 15000 v 30 ma. DC Bridge Rect. Pwr. Sply. Oper. CM. 115 v 60 cy.	$115.00
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
APS-2 10CM RF HEAD COMPLETE WITH HARD TUBE (715B) Pulser, 714 Magnetron 417A Mixer all ⅞" rigid coax. incl. revr. front end	$210.00
MODEL TS-268/U Test set designed to provide a means of rapid checking of crystal diodes IN21, IN21A, IN21B, IN23, IN23A, IN23B. Operates on 1½ volt dry cell battery. 3 x 6 x 7. New	$35.00

3CM WAVEMETER Coverage 9000-9500 Mcs w/Calib. Chart Absorb. Type w/Circ. Flange or XMSN. Type w/Sq. Flanges. New	$75.00
SL WAVEMETER Type Cw60ABM	$125.00
10CM ECHO BOX CABV 14ABA-1 of OBU-3, 2890 MC to 3170 MCS, direct reading micrometer head. Ring prediction scale plus 9% to minus 9%. Type "N" input. Resonance indicator meter. New and Comp. w/access. Box and 10 CM Directional Coupler	$350.00
10 cm. horn assembly consisting of two 5" dishes with dipoles feeding single type "N" output. Includes UG28/U type "N" "T" junction and type "N" pickup probe. Mfg. cable. New	$15.50
10 cm. cavity type wavemeters 6" deep, 6½" in diameter. Coax output. Silver plated	$64.50 ea.
10 Cm. echo box part of SF radar w/115 volt DC tuning Motor sub sig 114BA	$47.50
THERMISTER BRIDGE: Power meter 1-203-A, 10cm. mfg. W.E. Complete with meter, interpolation chart, portable carrying case	$72.50
W.E. I 138. Signal generator, 2700 to 2900 Mc. range. Lighthouse tube oscillator with attenuator & output meter, 115 VAC input reg. Pwr. supply With circuit diagram	$75.00
3 CM. stabilizer cavity, transmission type	$20.00
3 CM. HORN AT-48/UP model 710. Type "N" input Hvy. silver plated	$6.50
AT-68/UP 3 CM Horn with type "N" fitting	$5.00
TS-89/AP Voltage Divider: Ranges 100; ½ for 200 to 20000V 10; 1 for 200 to 2000v. Input Z 2000 ohm, output Z 4 meg. flat response 150-5 meg. cys.	$42.50
10 CM WAVEMETER W.E. type B-43590 Transmission type. "N" fittings. Veeder root mic. dial gold plated w/calib chart. P/O W.E. Freq. mtr. X66404A. New	$99.50

THERMISTORS

D-167332 (tube)	$.95
D-170396 (bead)	$.95
D-67613 (button)	$.95
D-104690 for MTG in "X" band Guide	$2.50
D-167618 (tube)	$.95

VARISTORS

D-170225	$1.25
D-167176	$.95

D-168087	$.95
D-171812	$.95
D-171528	$.95
D-168549	$.95
D-168442	$3.00
D-163293	$1.25
D-98428	$2.00
D-16187A	$2.85
D-171121	$.95
SA (12-43)	$1.50
D-167620	$3.00
D-105598	$2.25

K BAND

APS-34 Rotating Joint	$49.50
Right Angle Bend E or H Plane; specify combination of coupling desired	$12.00
45° Bend E or H Plane, Choke to cover	$12.00
Directional coupler CU-103/APS 32	$49.50
Mitered Elbow, cover to cover	$4.00
TR-ATR Section, choke to cover	$4.00
Flexible Section 1" choke to choke	$5.00
"S" Curve choke to cover	$4.50
Adaptor, round to square cover	$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.
Circular Flanges	$.50
Flange Coupling Nuts	$.50
Slotted line, Demornay-Budd #397, new	$450.00
90° Twist	$10.00

DE MORNAY BUDD

ALL FORMER STOCK AVAILABLE THROUGH COMMUNICATIONS EQUIPMENT

S BAND

Coax xtal Mount for Type "N" Tunable	$15.00
90° Twist, circular cover to circular cover	$25.00
Magnetron to Waveguide Coupler with 721A Duplexer Cavity, gold-plated	$45.00
Waveguide Switch—Transposes one input to any of three outputs. Standard 1½" x 3" square flanges. Complete with 115V drive motor. Raytheon CRT24AAS, new	$150.00
721A TR Box complete with tube and tuning Plungers	$12.50
McNally Klystron Cavities for 707B or 2K28. Three types available	$4.00
Right Angle Bend 5½ ft. over-all with 8" slotted section	$21.00
Pick-up Dipole in Lucite Ball with Sperry Fitting	$4.50
F-29/SPR-2 Filters, Type "N", input and output	$12.50
726 Klystron Mount, Tunable output, to type "N", complete, with socket and mounting bracket	$12.50
WAVEGUIDE to ⅞" Rigid Coax. "Doorknob" Adapter, Choke Flange, Silver Plated, Broad Band	each, $37.50
WAVEGUIDE Directional Coupler, 27 db. Navy type CABY-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
Crystal Mixer with tunable output TR pick up loop, Type "N" connectors. Type 62ABH	$14.50
Slotted line probe. Probe depth adjustable. Sperry connector, type CPR-14AAO	$9.50
Coaxial slotted section. ⅝" rigid coax with carriage and probe	$25.00
Right Angle Bend 6" radius E or H plain	$15.00
Right Angle Bend 3" radius E or H plain—Circular flanges	$15.00
AN/APR5A 10 cm antenna equipment consisting of two 10 CM waveguide sections, each polarized, 45 degrees	$75.00 per set
PICKUP LOOP, Type "N" Output	$2.75
TR BOX Pick-up Loop	$1.25
POWER SPLITTER: 726 Klystron input dual "N" output	$5.00
"S" BAND Mixer Assembly, with crystal mount pick-up loop, tunable output	$3.00
721-A TR CAVITY WITH TUBE. Complete with tuning plungers	$12.50
10CM OSC. PICKUP LOOP, with male Homedell output	$2.00
MAGNETRON To W.G. Coup'g for 1¾" Mag. Outp't Fit'g.	$65.00
10 CM FEEDBACK D'POLE ANTENNA, in lucite ball, for use with parabola ⅞" Rigid Coax. Input	$8.00
PHASE SHIFTER. 10 CM Waveguide. WE type ES-683816. E Plane to H Plane Matching Slugs. Mark 4	$95.00
721A TR cavities. Heavy silver plated	$2.00 ea.
16 cm. horn and rotating joint assembly, gold plated	$65.00 ea.
AS14A/AP 10 CM dipole pickup ant. w/10 ft. cable type N fittings	$3.25
10 CM Mixer	$3.00

⅞" RIGID COAX

Directional coupler, Type "N" take off	$22.50
Magnetron Coupling with TR Loop, gold-plated	$7.50
⅞ Rigid Coax Coupler	$17.50

MICROWAVE COMPONENTS

All merch. guar. Mail orders promptly filled. All prices, F.O.B., N.Y.C. Send Money Order or Check. Only shipping charges sent C.O.D. Rated Concems send P.O.

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PULSE TRANSFORMERS

G.E.K.-23745 .... $39.50

G.E.K.-23744-A. 11.5 KV High Voltage, 3.23 KV Low Voltage @ 200 KW oper. (270 KW max.) 1 microsec. or ¼ microsec. @ 600 PPS $39.50

W.E. #D166173 Hi-Volt input transformer, W.E. Impedance ratio 50 ohms to 900 ohms. Freq. range: 10 kc. to 2 mc. 2 sections parallel connected, potted in oil $36.00

W.E. KS 9800 Input transformer. Winding ratio between terminals 3-5 and 1-2 is 1.1:1, and between terminals 6-7 and 1.2 is 2:1 Frequency range: 380-520 c.p.s. Permalloy core $6.00

G.E. #K2731 Repetition Rate: 635 PPS, Pri. Imp: 50 Ohms. Sec. Imp: 450 Ohms Pulse Width: 1 Microsec. Pri. Input: 9.5 KV PK Sec. Output: 28 KV PK. Peak Output: 800 KW Riflar 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

G.E. #K2748A. Pulse Input line to magnetron $36.00

#9262 Utah Pulse or Blocking Oscillator XFMR Freq. limits 790-810 cy-3 windings turns ratio 1:1:1 Dimensions 1 13/16 x 1⅛" 19/32 .... $1.50

Pulse 131-AWP L-421435 .... $6.00

Pulse 134-BW-2F L-440895 .... $2.25

DELAY LINES

D-163169 Delay Line Small quantity available $50.00

D-168184: .5 microsec. up to 2000 PPS 1800 ohm term $4.00

D-170499: .25/.50/.75 microsec 8 KV 50 ohms imp. $16.50

D-165997: 1¼ microsec. $7.50

MAGNETRONS

Tube	Frq. Range	Pk. Pwr. Output	Price
2J27	2965-2992 mc.	275 KW	$ 8.50
2J31	2820-2860 mc.	265 KW	$25.00
2J21-A	9345-9405 mc.	50 KW	$25.00
2J22	3267-3333 mc.	265 KW	$25.00
2J26	2992-3019 mc.	275 KW	$25.00
2J32	2780-2820 mc.	285 KW	$25.00
2J37			$45.00
2J38 Pkg.	3249-3263 mc.	5 KW	$35.00
2J39 Pkg.	3267-3333 mc.	87 KW	$35.00
2J40	9305-9325 mc.	10 KW	$65.00
2J49	9000-9160 mc.	58 KW	$85.00
2J34			$55.00
2J61	3000-3100 mc.	35 KW	$65.00
2J62	2914-3010 mc.	35 KW	$65.00
3J31	24,000 mc.	50 KW	$55.00
5J30			$39.50
714AY			$25.00
718DY	2720-2890 mc.	250 KW	$25.00
7208Y	2800 mc.	1000 KW	$50.00
720CY	2860 mc.	1000 KW	$50.00
725-A	9345-9405 mc.	50 KW	$25.00
730-A	9345-9405 mc.	50 KW	$25.00
728	AY, BY, CY, DY, EY, FY, GY		$50.00
700	A, B, C, D		$50.00
706	AY, BY, DY, EY, FY, GY		$50.00
Klystrons	723A/B $12.50; 707B W/Cavity		$20.00
	417A $25.00	2K41	$65.00

MAGNETRON MAGNETS

Gauss	Pole Diam.	Spacing	Price
4850	¾ in.	⅝ in.	$ 8.90
5200	21/32 in.	¾ 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 1½" (2200 Gauss) Pole Dia. 1⅝". New Part of SCR 584 .... $34.50

CW" MAGNETRONS

QK 62	3150-3375 mc.
QK 59	2675-2900 mc.
QK 61	2975-3200 mc.
QK 60	2800-3025 mc.

New, Guaranteed .. Each $65.00

QK 915 .... $150.00

FILAMENT TRANSFORMER

for above 115V/60 cy Pri: four 6.3V/4A Sec. 5000VT $27.50

Magnetron Kit of four QK's 2675-3375 inc. w/trans special .. $250.00

MAGNETRONS

PULSE EQUIPMENT

PULSE NETWORKS

Ray-WX4298F $39.50

GE-K6824730 .... $50.00

GE-K9216945 .... $50.00

15A—1-400-50: 15 KV, "A" CKT, 1 microsec 400 PPS 50 ohms imp. $42.50

G.E. #6E3-5-2000-50P2T, 6KV "E" circuit, 3 sections, .5 microsecond, 2000 PPS, 50 ohms impedance $6.50

G.E. #3E (3-84-810; 8-2.24-405) 50P4T; 3KV, "E" CKT Dual Unit: Unit 1, 3 Sections, .84 Microsec. 810 PPS, 50 ohms imp.: Unit 2, 8 Sections, 2.24 microsec. 405 PPS, 50 ohms imp. $6.50

7.5E3-1-200-67P. 7.5 KV, "E" Circuit, 1 microsec 200 PPS, 67 ohms impedance, 3 sections $7.50

7.5E4-16-60.67P. 7.5 KV, "E" circuit 4 sections 16 microsec. 60 PPS, 67 ohms impedance $15.00

7.5E3-3-200-6PT. 7.5 KV, "E" Circuit, 3 microsec 200' PPS 67 ohms imp. 3 sections $12.50

PULSE EQUIPMENT

MIT. MOD. 3 HARD TUBE PULSER: Output Pulse Power 144 KW (12 KV at 12 Amp). Duty Ratio: 001 max. Pulse duration: 5, 1.0, 2.0 microsec. Input voltage: 115 v. 400 to 2400 ops. Uses: 1-715B, 4-829-B, 3-'72's, 1-'73. New $110.00

APQ-13 PULSE MODULATOR. Pulse Width .5 to 1.1 Micro Sec. Rep. rate 624 to 1348 Pps. Pk pwr. out 35 KW Emergy 0.018 Joules $49.00

TPS-3 PULSE MODULATOR. Pk. power 50 amp. 24 KW (1200 KW pk); pulse rate 200 PPS. 1.5 microsec. pulse line impedance 50 ohms. Circuit—series charging version of DC Resonance type. Uses two 705-A's as rectifiers. 115 v. 400 cycle input. New with all tubes $49.50

APS-10 MODULATOR DECK. Complete, less tubes $75.00

BC 1203B Loran pulse modulator $125.00

BC 758A Pulse modulator $395.00

APS-10 Low voltage power supply less tubes $18.50

725A magnetron pulse transformers .... $18.50 ea.

DIRECTION FINDERS

DA8 3 & 4. 2 18 Mc mfg. like new $850.00

DAK Direction Finder Automatic bearing indicators $185.00

complete receiver $225.00

loops $125.00

RG 23U Twin conductor rf cable 250 ft. reel $50.00

DP12 Direct. Finder 100-1500 kc. $250.00

DF Rec. only Bludworth Standard Arrow $150.00

I.F.F. 1 KW Pulsed Output Pkg. Tunable 154-186 mc. adj. modulating pulses 4-10 micro sec. comp. 115v 60 cy ac pwr. supply.

Vidio output receiver. New w/tubes $350.00

Wavemeter for above $75.00

Dipole Array for above $85.00

INDICATORS—SCOPES

BC 931B 4-20-50-100 mile range 5" scope w/mtg. rack, indicator amplifier, BC 932B, visor, ne. w/tubes $24.50

BC 704A 9-36-90 mile range 5" scope. Write

BC 937A & BC 938A 12" PPI & "A" scope. Complete desk Rack assy w/osc. control unit. rec., pwr. supls. in unused cond. but shelf worn $300.00

Radar Indicator RW #81 mfg. by Research Enterprise Ltd. 5" scope $30.00

12" PPI Radar indic. console P/O SK-1M Radar

ASD Indicator
ID30 APS2 Indicator
929 Indicator
} Write or phone for information and price.

Many others in stock.

PRECISION CAPACITORS

D-163707: 0.4 mfd @ 1500 vdc. —50 to plus 85 deg C $4.50

D-163035: 0.1 mfd @ 600 vdc, 0 to plus 65 deg C. $2.00

D-170908: 0.152 mfd. 300 v. 400 cy. —50 to plus 85 deg C $2.50

D-164960: 2.04 mfd @ 200 vdc. 0 to plus 55 deg C $2.50

D-168344: 2.16 mfd @ 200 vdc. 0 to plus 55 deg C $3.00

D-161555: 5 mfd @ 400 vdc. —50 to plus 85 deg C $3.00

D-161270: 1 mfd @ 200 vdc. temp comp —40 to plus 65 deg C $12.50

30' U. S. ARMY SIGNAL CORPS RADIO MASTS

Complete set for the erection of a full flat top antenna. Of rugged plymold construction telescoping into 3 ten-foot sections for easy stowage and transportation. Supplied complete: 2 complete masts, hardware, shipping crate. Shipping wt. approx. 300 lbs. Sig Corps #2A289-223-A. New .... $35.00 per set of 2

MICROWAVE ANT.

RF EQUIPMENT

MICROWAVE ANTENNAS

AN-122 Dipole Assy. $22.50

LP-21-A ADF Loop W/Selsyn and Housins. New $8.00

DAK Bellino Tossi DF Loops ... $125.00

Adcock DF Arrays, Complete $65.00

SA Radar, 200 MC Bed Springs, Complete with Pedastec, Less Drive $600.00

AN MGP-1 Antenna. Rotary feed type high speed scanner antenna assembly, including horn parabolic reflector. Less internal mechanisms. 10 deg. sector scan. Approx. 12'L x 4'W x 3'H. Unused. (Gov't Cost—$4500.00) $250.00

APS-4 3 cm. antenna. Complete. 14½ dish. Cutler feed dipole directional coupler, all standard 1" x ½" waveguide. Drive motor and gear mechanisms for horizontal and vertical scan. New, complete $65.00

AN/TPS3. Parabolic dish type reflector approx. 10' diam. Extremely lightweight construction. New in 3 carrying cases $89.50

RELAY SYSTEM PARABOLIC REFLECTORS: approx. range: 2000 to 6000 mc. Dimensions: 4' x 3' rectangle, now $35.00

TDY "JAM" RADAR ROTATING ANTENNA. 10 cm. 30 deg. beam. 115 v.a.c. drive. New $100.00

DBM ANTENNA. Dual, back-to-back parabolas with dipoles. Freq. coverage 1,000-4,500 mc. No drive mechanism $65.00

AS125/APR Cone type receiving antenna, 1080 to 3208 megacycles. New $4.50

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

AS17/APS 10 CM Antenna, APS-2 30 Inch Dish with ⅞ Coax Dipole and fittings, New and Compl. with 24 V DC Drive motor, selsyn. 360 Deg. Rotation and Vertical Tilt $94.50

RC-224 Antenna, 10 CM, 30" Dish P/O. SCR-717 Radar, New and Complete $94.50

R. F. EQUIPMENT

LHTR. LIGHTHOUSE ASSEMBLY, Part of RT-30/APG-5 & APG 15. Receiver and Transmitter Lighthouse Cavities with assoc. Tr. Cavity and Type N CPLG. To Revr. Uses 2C40, 2C43, 1B27, Tunable APX 2400-2700 MCS. Silver plated. $49.50

APS-2 10CM RF HEAD COMPLETE WITH HARD TUBE (715B) Pulser. 714 Magnetron 417A Mixer all ⅞" rigid coax. incl. revr. front end $210.00

Beacon lighthouse cavity 10 cm with miniature 28 volt DC FM motor. Mfg. Bernard Rice . $47.50 ea.

T-128-/APN-19 10 cm. radar Beacon transmitter package, used, less tubes .... $59.50 ea.

SO-3 "X" band 3 cm RF package, new complete, including receiver unit as illustrated on Page 337, Volume 23 RAD LAB Series .... $375.00 ea.

Pre-Amplifier cavities type "M" 7410590GL, to use 446A lighthouse tube. Completely tuneable. Heavy silver plated construction .... $37.50 ea.

RT32/APS 6A RF HEAD. Compl. with 725A Magnetron magnet pulse xfmr. TR-ATR 723A/B local osc. and beacon mount, pre amplifier. Used but exc. cond. $97.50

AN/APS-15A "X" Band compl. RF head and modulator, incl. 725-A magnetron and magnet, two 723A/B klystrons (local osc. & beacon) 1B24, TR, RCVR. ampl. duplexer, HV supply blower, pulse xfmr. Peak Pwr Out: 45 KW apx. Input: 115, 400 cy. Modulator pulse duration .5-2 microsc., apx. 13KV, PK, Pulse, with all tubes incl. 715B, 829B, RKR 73, two 72's. Complete pkg. $210.00

S BAND AN/APS2. Complete RF head and modulator, including magnetron and magnet, 417A mixer, TR receiver duplexer, blower, etc., and complete pulser. With tubes, used, fair condition $75.00

10 CM RF Package. Consists of: SO Xmtr. receiver using 2J27 magnetron oscillator, 250 KW peak input. 707-B receiver-mixer $150.00

ASB-500 Megacycles Radar Receiver with two GI 446 lighthouse cavities, new less tubes $37.50

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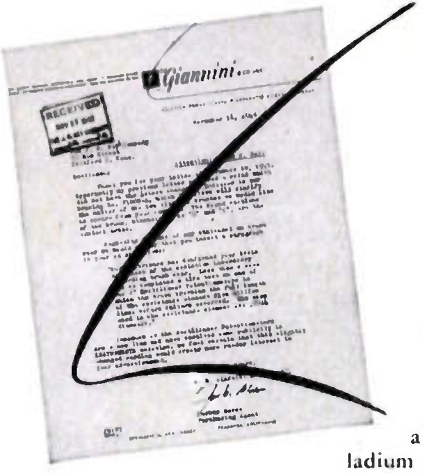
COMMUNICATIONS EQUIPMENT COMPANY

131 LIBERTY STREET, NEW YORK, N.Y. DEPT. 16 P. J. PLISHNER PHONE DIGBY—9-4124

Manufacturer's tests confirm superiority of **PALINEY* #7** for brushes on new Rectilinear Potentiometer . . .



"Our experience has confirmed your tests and those of the Radiation Laboratory regarding brush wear. Less than a week ago we completed a life test on one of our 2" Rectilinear Potentiometers in which the brush traveled the full length of the resistance element five million times before failure occurred. The wire used in the resistance element was .0014 diameter."

PALINEY* #7 . . . a precious metal alloy containing gold, platinum and palladium . . . is giving outstanding service as the sliding contact in many types of potentiometers where long life, low noise and maintained linearity are essential. This and other Tested NEY Precious Metal Alloys are also being used successfully in numerous precision contact and slip ring applications requiring controlled wear resistance, high conductivity and freedom from tarnish and corrosion. Write or call our Research Department for additional technical data, outlining your problem if possible.

THE J. M. NEY COMPANY 171 ELM ST., HARTFORD, CONN.
SPECIALISTS IN PRECIOUS METAL METALLURGY SINCE 1812

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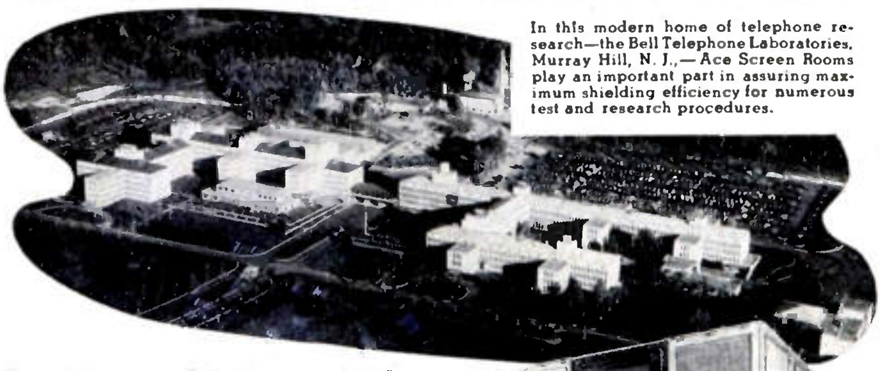
ON THE JOB AT THE FAMOUS BELL "LABS"



In this modern home of telephone research—the Bell Telephone Laboratories, Murray Hill, N. J.,—Ace Screen Rooms play an important part in assuring maximum shielding efficiency for numerous test and research procedures.

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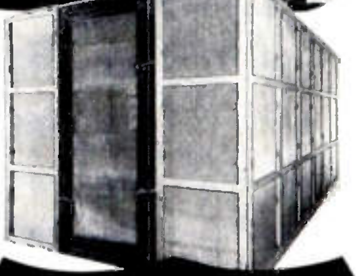
... AT NO GREATER COST!



Designed to exacting wartime laboratory standards, supplied in ready-built "knockdown" form for installation in a few hours, Ace Screen Rooms provide a minimum of 100 db. attenuation from 0.15 to 1000 mc. Total cost is no greater than that of "homemade" screen rooms of far lower efficiency. Numerous sizes are available and rooms can readily be moved or enlarged as required. Write, wire, or 'phone for details.

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3642 N. Lawrence St.
Philadelphia 40, Pa. REgent 9-1019

ACE
READY-BUILT "UNIT CELL"
SCREEN ROOMS

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 79A)

New Resistance Boxes

New resistance boxes, just announced by **Leeds & Northrup Company**, 4934 Stention Ave., Philadelphia 44, Pa., are designed for use as moderate-precision, adjustable standards for dc and ac resistance measurements up to medium frequencies.

Principal feature of the boxes is an entirely new type of rotary switch having exceptionally low and stable contact resistance, obtained through the use of solid silver switch contacts and silver alloy multiple-leaf brushes. Zero or contact resistance is less than 0.002 ohm per decade, and changes less than 0.0005 ohm per decade on accelerated life tests.

Limit of error is ±0.05 per cent 0.005 ohm. The 4-dial model has a range of 11,110 ohms in 1-ohm steps. The 5-dial models cover 11,111 ohms in 0.1-ohm steps or 111,110 ohms in 1-ohm steps. Time constant at 50 kc for usual range of dial settings is less than 0.05 microsecond.

Microwave Signal Source

Polarad Electronics Corp., 100 Metropolitan Ave., Brooklyn 11, N. Y., has developed a series of microwave signal sources, covering the range of 634 to 8,340 Mc. in four units.

These reflex klystron signal sources are controlled by one dial only and frequency is read directly from a linear indicator to accuracies of $\frac{1}{2}$ per cent. The reflector voltage is automatically tracked with the cavity tuner. There are no klystron modes to set, no voltage settings to be made, and frequency can be read directly without resorting to charts or graphs.

The Model SSR microwave signal source covers the range from 634 to 1,174 Mc; Model SSM—4,290 to 8,340 Mc. The signal sources are supplied complete with tube.

(Continued on page 83A)

News—New Products

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

(Continued from page 82A)

New Model Power Supply

Kepco Model 315 power supply, which features one regulated B supply, one regulated C supply, and one unregulated filament supply is now being marketed by the manufacturer, **Kepco Laboratories, Inc.,** 149–14 41 Ave., Flushing, L. I., N. Y.

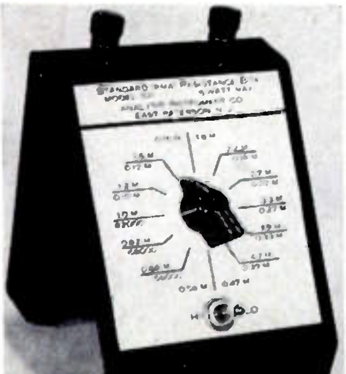
The B supply is continuously variable from 0 to 300 volts and delivers from 0 to 150 ma. In the range 20 to 300 volts the output voltage variation is less than ½ per cent for both line fluctuations from 105 to 125 volts and load variations from minimum to maximum current. Ripple is less than 5 millivolts.

The C supply is continuously variable from 0 to 150 volts and delivers 5 ma. For all output voltages, the output voltage variation is less than 10 millivolts for line fluctuations of 105 to 125 volts. At 150 volts, the regulation is less than ½ per cent between 0 and 5 ma. At other settings below 150 volts, the internal resistance of the supply will increase to a maximum of 25,000 ohms. Ripple is less than 5 millivolts.

The ac output is 6.3 volts, 5 amperes, center-tapped, unregulated.

Standard Resistance Boxes

A new resistor decade for electronic production and laboratory use is being offered by **Analysis Instrument Co.,** P.O. Box 231 East Paterson, N. J.



Model 101 contains all RMA 10 per cent resistance values from 47,000 ohms

(Continued on page 85A)

AMPERITE

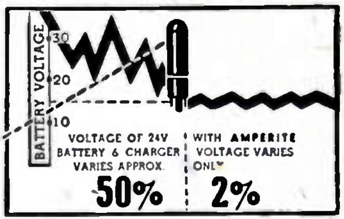
THERMOSTATIC METAL TYPE
Delay Relays

PROVIDE DELAYS RANGING FROM 1 TO 120 SECONDS

FEATURES: — Compensated for ambient temperature changes from —40° to 110° F . . . Hermetically sealed; not affected by altitude, moisture or other climate changes . . . Explosion-proof . . . Octal radio base . . . Compact, light, rugged, inexpensive . . . Circuits available: SPST Normally Open; SPST Normally Closed.

PROBLEM? Send for *"Special Problem Sheet"*

Regulators



Amperite REGULATORS are the simplest, lightest, cheapest, and most compact method of obtaining current or voltage regulation . . . For currents of .060 to 6 Amps. . . . Hermetically sealed; not affected by altitude, ambient temperature, humidity.

Write for 4-page Illustrated Bulletin RE

AMPERITE CO., Inc., 561 Broadway, New York 12, N. Y.
In Canada: Atlas Radio Corp., Ltd., 560 King St., W. Toronto

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10 MICROVOLTS TO 10,000 VOLTS

This enormous range of AC voltages is easily covered by the Ballantine Model 300 Voltmeter, Model 220 Decade Amplifier and Model 402 Multipliers. The accuracy is 2% at any point on the meter scale, over a frequency range of 10 cycles to 150 kilocycles. The Model 300 Voltmeter (AC operated) reads from .001 volt to 100 volts, the Model 220 Amplifier (battery operated) supplies accurately standardized gains of 10x and 100x and the Model 402 Multipliers extend the range of the voltmeter to 1,000 and 10,000 volts full scale.

Write for descriptive Bulletin No. 12-A

MODEL 300

Featuring a Logarithmic Voltage Scale and Uniform Decibel Scale

BALLANTINE LABORATORIES, INC.
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UG-27A/U
UG-22B/U
UG-21B/U
UG-306/U
UG-274/U
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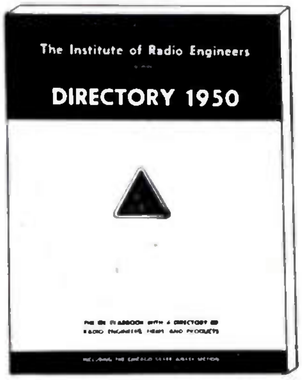
UG-58/U
PL-259A
PL-259
SO-239
UG-36/U



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 83A)

to 2.7 megohms, ½-watt dissipation and 10 per cent accuracy.

Model 102 has resistance values from 680 to 39,000 ohms with ½-watt dissipation and 10 per cent accuracy.

These decades make it possible to place RMA resistor values in a circuit without the necessity of drawing a number from stock for trial. It also eliminates the potentiometer-ohmmeter method of first approximation of circuit values, or the use of expensive precision resistance boxes for this purpose.

Models 101A, and 102A, identical with the 101, and 102 except for 5 per cent resistor accuracy, are also available.

Utility Amplifiers

A new series "E," amplifiers, which includes 10-, 17-, 25-, and 50-watt models, phone-taps, mobile and portable assemblies is in production at **Newcomb Audio Products Inc.**, 6824 Lexington Ave., Hollywood 38, Calif.

Model E25 (illustrated), is a 25- to 30-watt amplifier. It has inputs for two high-impedance microphones and a phonograph, with knock-out holes in chassis to provide for easy conversion of mike inputs to low impedance if needed. Individual bass and treble tone controls offer wider range of adjustment and feature bass emphasis for phonograph without emphasizing voice bass. Molded type coupling condensers give added protection from heat and moisture.

Other specifications: Power output, 25 watts at less than 5 per cent distortion; frequency response, ±2 db, 40 to 15,000 cps; output impedances 4, 8, 16, 500 ohms.

NOTICE

"Information for our News and New Products section is warmly welcomed. News releases should be addressed to Industry Research Division, Proceedings of The I.R.E., Room 707, 303 West 42nd Street, New York 18, New York. Photographs and electrotypes, if not over two inches wide, are helpful. Stories should pertain to products of interest specifically to Radio Engineers."

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LABORATORY POWER SUPPLIES

- STABLE
- DEPENDABLE
- MODERATELY PRICED

MODEL 28
STANDARD RACK MOUNTING
PANEL SIZE 3½" x 19"
WEIGHT 16 LBS.

- **INPUT:** 105 to 125 VAC, 50-60 cy
- **OUTPUT #1:** 200 to 325 Volts DC at 100 ma regulated
- **OUTPUT #2:** 6.3 Volts AC CT at 3A unregulated
- **RIPPLE OUTPUT:** Less than 10 millivolts rms

For complete information write for Bulletin G8

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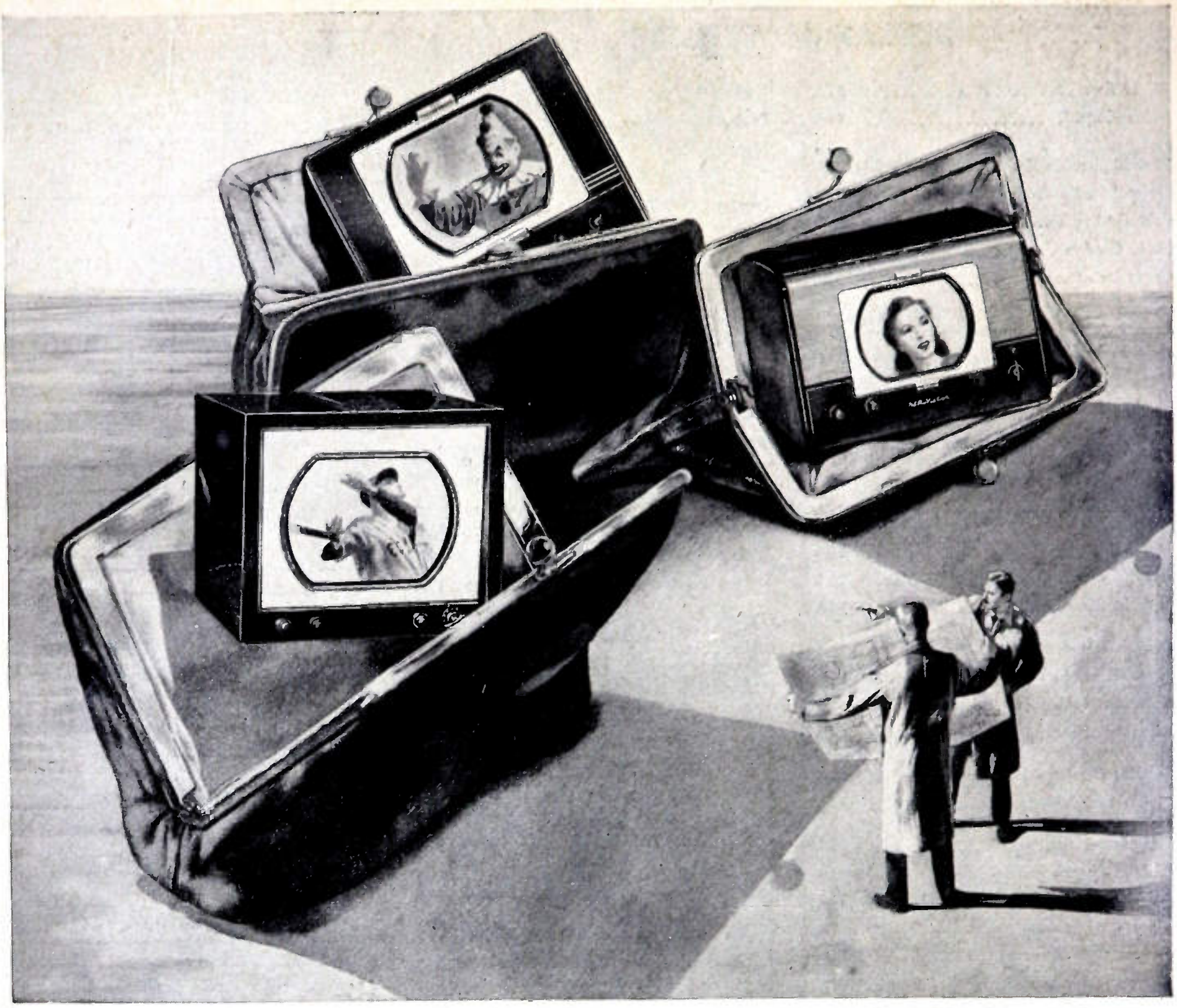

Developments by RCA scientists have made Television part of family life in homes of all incomes.

How research fits television into **more purses**

Remember when television was "just around the corner," and guesses at receiver cost ranged to thousands of dollars? Came *reality*, and pessimists were wrong. Home television sets were reasonable, grew more so year by year.

One factor has been research at RCA Laboratories. For example: In 1949, RCA scientists perfected the glass-and-metal picture tube—so adaptable to mass production that savings of 30% in tube cost were made. Again, these scientists and development engineers learned how to replace complex parts with less costly, and more efficient materials. A third contribution was the use of versatile *multiple-purpose* tubes—so that one could do the work of several!

Most important, the savings effected by RCA scientists have been quickly passed on to you, the consumer. RCA Laboratories is known as a great center of radio, television, and electronic research. It is indeed an institution which fits RCA products into more purses!

See the latest in radio, television, and electronics at RCA Exhibition Hall, 36 W. 49th St., N. Y. Admission is free. Radio Corporation of America, RCA Building, Radio City, N. Y.

New RCA Victor 16-inch television receiver, a leader in the 1950 line.

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Now you can get the same dependability that has made C-D's famous in engineering circles the world over, on a twenty-four-hour delivery basis. The demand for "Silver Mike"* Micas through the years has been so great that every expansion program has proved insufficient. But with the completion of our last expansion program we are making available the entire Providence plant exclusively to the production of the famous, reliable C-D line of "Silver Mike"* Micas. Of course, in addition to the new, improved delivery standards, you still get these other features that have made C-D's "Silver Mike"* Micas so popular:

Extra heavy silver coating thoroughly bonded to mica — results in a uniform and low capacity-temperature coefficient (+.002% per degree C.); excellent retrace characteristics; practically no capacity drift with time.

Molded in low-loss red compound — results in an exceptionally high Q (3,000 to 5,000); fixed electrical characteristics.

Wax impregnated — results in a humidity-proof capacitor.

"Silver Mike"* Mica Capacitors are available in 300 and 500 V.D.C., and in capacities from .000001 to .005 mfd. at standard tolerance of ±5%. "Silver Mike"* Micas can also be supplied, on special order, to tolerances of ±3%, ±2% and ±1%.

Your inquiries are invited. CORNELL-DUBILIER ELECTRIC CORPORATION, Dept. M60 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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RADIART SUBSIDIARY

Announcing A New GR Standard Signal Generator

50 to 920 Mc

THE new General Radio Type 1021-A Standard-Signal Generator operates at frequencies between 50 and 920 Mc with the same convenience and reliability found in other G-R generators in the broadcast frequencies.

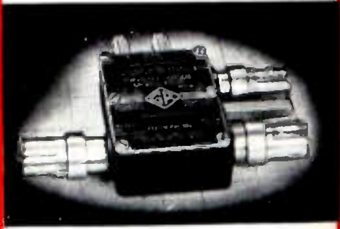
Its main use is the determination of radio receiver and circuit characteristics. With an inexpensive diode modulator, television picture modulation can be produced for overall testing of television receivers.

It is a convenient and well-shielded source of power for measurements with bridges, impedance comparators, and slotted lines. For these uses internal modulation is provided.

With the new G-R Type 874 line of Coaxial Elements, this generator provides a very complete and flexible system for measurements of voltage, power and standing-wave ratio from 50 to 920 Mc.

Type 1021-P2 Oscillator Unit (250-920 Mc) Two separate oscillators are available. They are mechanically and electrically interchangeable, and are sold as separate units to convert the range of one standard-signal generator to that of the other.

TYPE 1021-P2 U-H-F Oscillator Unit only (250-920 Mc)....	**$420.00**
TYPE 1021-P3 V-H-F Oscillator Unit only (50-250 Mc).....	**$400.00**



Type 1000-P6 Crystal Diode Modulator An inexpensive, wide-band modulator for amplitude modulation of carrier frequencies between 20 and 1000 Mc. Modulation-frequency range is 0 to 5 Mc. **$35.00**

FEATURES

- **SIMPLICITY, RELIABILITY, CONVENIENCE** of a standard broadcast generator. **ACCURATE · COMPACT · LIGHTWEIGHT**
- **MODERATELY PRICED**
- **BUTTERFLY TUNING CIRCUIT** . . . no sliding contacts . . . no noise . . . perfectly smooth tuning . . . rugged design with good stability and very low drift
- **REGULATED POWER SUPPLY** assures good heterodyne beat note
- **OUTPUT FROM 0.5 MICROVOLT TO ONE VOLT** with overall accuracy better than ±20%
- **INTERNAL OUTPUT IMPEDANCE** 50 ohms
- **LEAKAGE AND RESIDUAL OUTPUT VOLTAGE** below sensitivity of most receivers
- **INTERNAL 1000-CYCLE AND EXTERNAL AMPLITUDE MODULATION** over audio range, adjustable from 0 to 50% . . . incidental fm under 100 parts per million over most of the ranges
- **T-V PICTURE MODULATION ON ALL CHANNELS** from 50 to 920 Mc with NO INCIDENTAL FM, when Type 1000-P6 Crystal-Diode Modulator and source of video signals are used. The power requirements for modulation are so low, video output from a standard T-V receiver can be used

TYPE 1021-AV V-H-F Standard-Signal Generator (50-250 Mc)..............	$595.00
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GENERAL RADIO COMPANY Cambridge 39, Massachusetts

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