Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

(Including the WAVES AND ELECTRONS Section)

FREDERICK B. LLEWELLYN
PRESIDENT, 1946

WALTER R. G. BAKER
PRESIDENT, 1947

January, 1947
Volume 35 Number 1

PROCEEDINGS OF THE I.R.E.

Waves and Electrons Section

1947 I.R.E. NATIONAL CONVENTION AT NEW YORK — MARCH 3, 4, 5, 6

The Institute of Radio Engineers

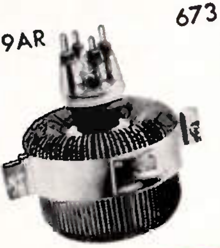
AMPEREX

ONE SOURCE • ALL TYPES

892R

COMMUNICATION

889AR



673

INDUSTRIAL

866A

857B

RECTIFICATION

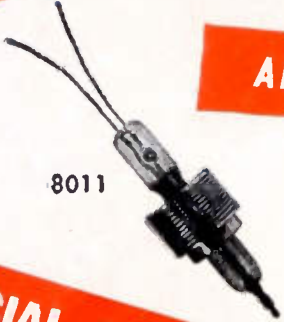
VC50 / VC25

HF200

ELECTRO-MEDICAL

AND ALL OTHER IMPORTANT TYPES

8011



SPECIAL PURPOSE

Is it a special development for new equipment? Or a "standard" tube for replacements? A quarter century of creative research, precision manufacture and helpful service has given Amperex a unique position in the power tube field. This means a backlog of experience and forward-looking viewpoint which naturally translate themselves into tube performance, reliability and economy. Consult us — no obligation.

Write for: Catalog; Technical Rating and Data Sheets

AMPEREX ELECTRONIC CORPORATION

25 WASHINGTON STREET, BROOKLYN 1, N. Y., CABLES: "ARLAB"

In Canada and Newfoundland: Rogers Majestic Limited, 622 Fleet Street West, Toronto 28, Canada

1947 I.R.E. NATIONAL CONVENTION and Radio Engineering Show

March 3 to 6, 1947

Hotel Commodore • **and** • **Grand Central Palace**

42nd Street and Lexington Avenue

46th Street and Lexington Avenue

NEW YORK CITY

More than 7000 members and visitors attended sessions and exhibits at the last National Convention of The Institute of Radio Engineers in January 1946. This year more favorable facilities and better dates have been obtained to improve session halls and exhibits. Plan now for these four days devoted to interesting and instructive technical papers—plus entertainment and The Radio Engineering Show. Free to I.R.E. members.

TECHNICAL PAPERS

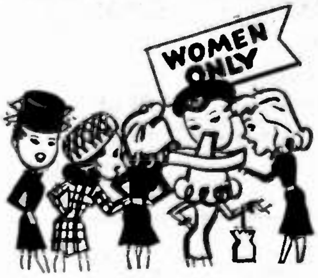
Over 100 papers to be presented during four days, covering twenty-four major topics, will reveal the up-to-the-minute pattern of the radio-electronic engineering interests of I.R.E. members. Topical outline by days will be:

• Monday: Particle Accelerators for Nuclear Studies, Electronic Measuring Equipment, Radar & Communication Systems, FM Reception. • Tuesday: Aids to Air Navigation, Neucleonics Instrumentation, Microwave Repeaters, Television, Electronic Digital Computors, Power Output Vacuum Tubes, Linear Circuit Theory. • Wednesday: Electronic Controls, Aids to Navigation, Microwave Techniques, Broadcasting & Recording, Professional Status of the Engineer. • Thursday: Oscillator Circuit Theory, Basic Electronics Research, Antennas, Relay and Pulse Time Systems of Communication, Repeater Circuits, Vacuum Tubes and Gas Rectifiers, Wave Propagation and Antennas, Wave Guide Techniques.

BANQUET

The annual banquet, on Wednesday, March 5, is the social highlight of the I.R.E. year. 1,600 feasters will hear a nationally prominent speaker, see two major awards made, be entertained royally.

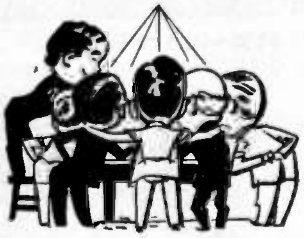
WOMEN'S PROGRAM



Four full days of fun!

"Sightseeing," "Fred Waring Show," "Macy's", Empire State Building, A Tea at the I.R.E. Building, "Cloisters" or U.N. if in session, Choice of matinees, Shopping, Fashion Show, all planned for ladies!

EXHIBITS



150 Exhibitors will present all that is new in radio and electronic equipment, instruments, components and parts on the first two floors of Grand Central Palace in our greatest "Radio Engineering Show." (Part of the Technical Sessions will be held on the third floor.) Exhibitors have taken double the space of our last Show. The exhibits are an engineering "must see"!

PRESIDENT'S LUNCHEON

The 1947 president will be honored on Tuesday, March 4, at a get-together which has come to be a feature of these annual meetings.

Proceedings of the I.R.E. and Waves and Electrons, January, 1947, Vol. 35, No. 1. Published monthly in two sections by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price $1.50 per copy. Subscriptions: United States and Canada, $12.00 a year; foreign countries $13.00 a year. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

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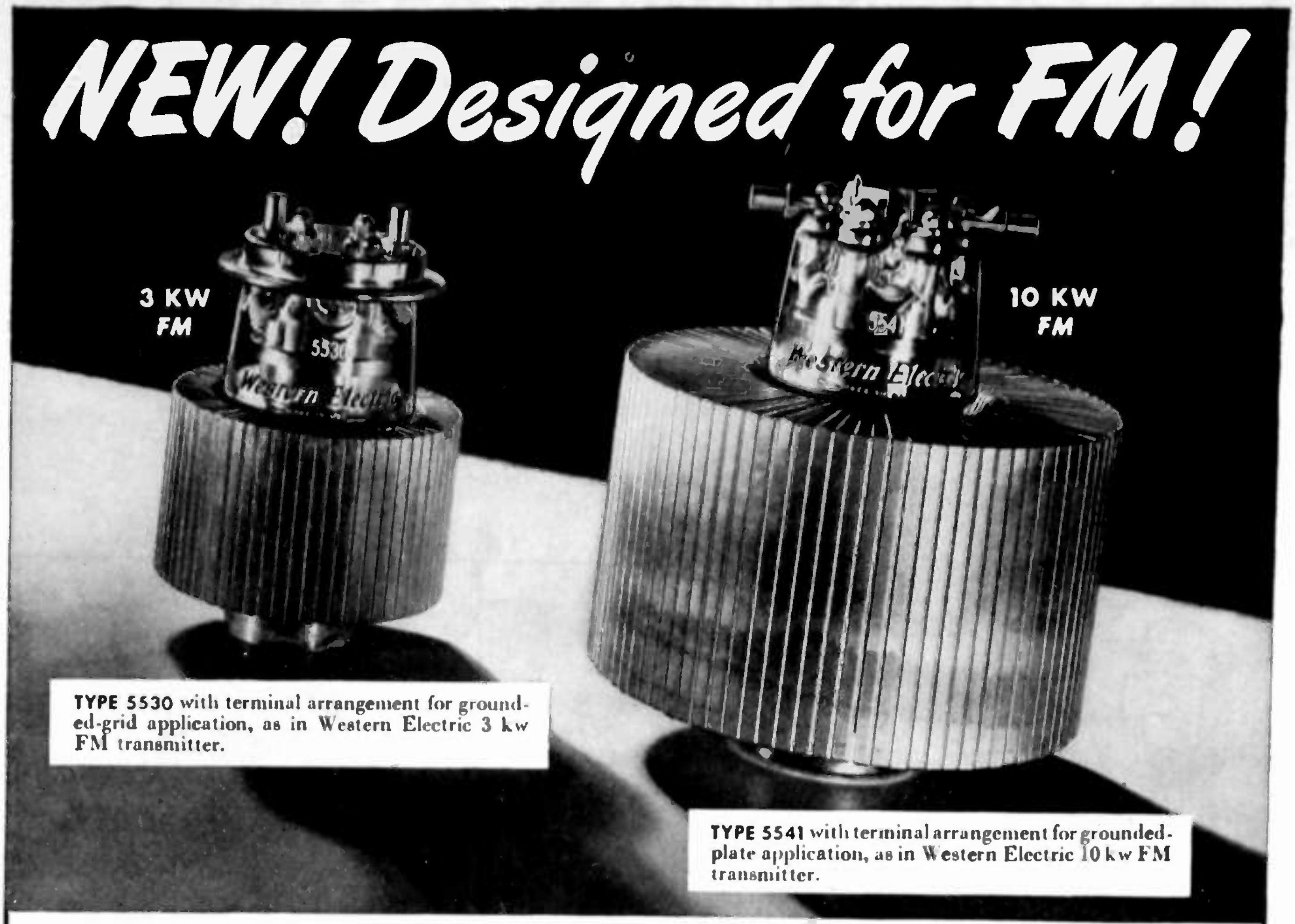

NEW! Designed for FM!
3 KW FM
10 KW FM
5530
Western Electric
5541
Western Electric

New BRUSH MAGNETIC RECORDING TAPE AND WIRE OFFER
LOWER COST... UNIFORMITY... EXCELLENT FIDELITY

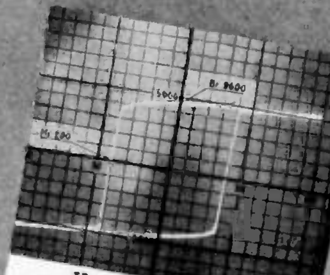
Outstanding Developments Produced by Pioneer and Leader in this Field

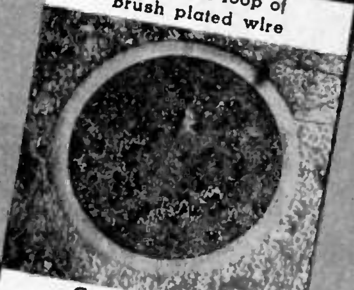
BRUSH PAPER TAPE

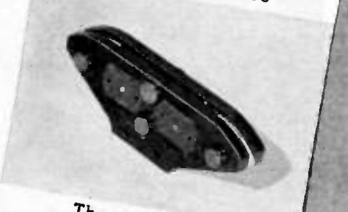
- Easy to Handle
- Extremely low-cost
- Can be edited . . . spliced
- Greater dynamic range
- Minimum wear on heads
- Excellent high frequency reproduction at slow speed
- Permanent—excellent reproduction for several thousand play-backs

BRUSH PLATED WIRE

- Constant plating thickness assures uniform signal
- Correct balance of magnetic properties assures good frequency response and high level
- Excellent surface finish assures low noise and minimum wear
- Corrosion resistant
- Easy to handle—ductile—can be knotted


Hysferysis loop of Brush plated wire


Cross section of Brush plated wire


The new Brush wire recording head


The new Brush tape recording head

Vastly Improved Tape and Wire Recording Heads and Cartridges

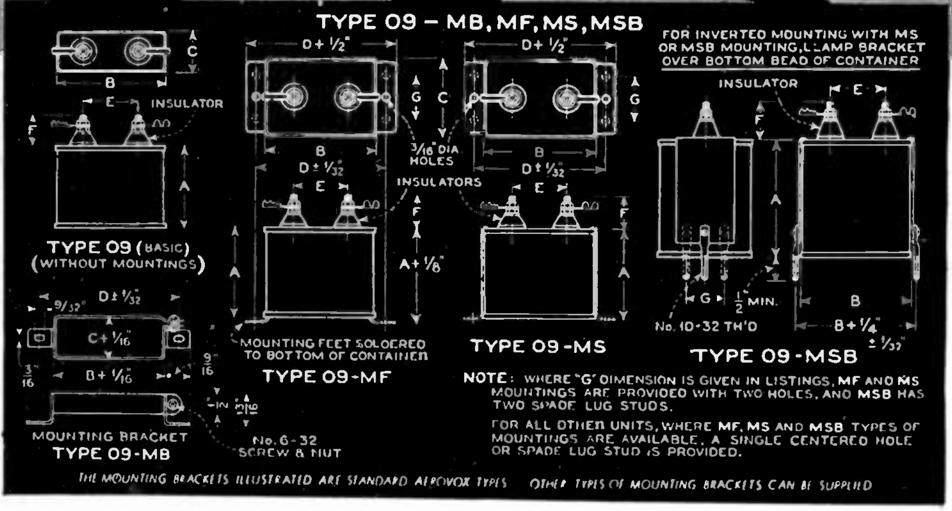
Another important improvement made by Brush has been the development of very simply constructed, low-cost erasing, recording and reproducing heads. These are the very heart of the magnetic recording unit and the intensive research and development work done by Brush has resulted in decided improvements. Of principal interest are their excellent electrical characteristics, extreme simplicity of design to avoid trouble, and the "hum-bucking" characteristics which reduce the effect of extraneous magnetic fields. When required, the head cartridge alone (pole piece and coil unit) may be supplied for incorporation into manufacturers' own head structure.

These latest developments in magnetic recording equipment can now be obtained for radio combinations and other uses. Brush engineers are ready to assist you in your particular use of magnetic recording components.

THE BRUSH DEVELOPMENT CO.
3405 Perkins Avenue • Cleveland 14, Ohio

Write today for further information

HEAVY-DUTY OIL-FILLED CAPACITORS



NOTE: WHERE "G" DIMENSION IS GIVEN IN LISTINGS, MF AND MS MOUNTINGS ARE PROVIDED WITH TWO HOLES, AND MSB HAS TWO SPADE LUG STUDS.

FOR ALL OTHER UNITS, WHERE MF, MS AND MSB TYPES OF MOUNTINGS ARE AVAILABLE, A SINGLE CENTERED HOLE OR SPADE LUG STUD IS PROVIDED.

THE MOUNTING BRACKETS ILLUSTRATED ARE STANDARD AEROVOX TYPES   OTHER TYPES OF MOUNTING BRACKETS CAN BE SUPPLIED

● **Versatility**—with economy of chassis space and assembly operations a prime factor—distinguishes Aerovox Type 09 oil-filled capacitors. Although mass-produced, this type is available in such an outstanding range of voltage and capacitance ratings, as well as mountings, that it is virtually custom-made for most high-voltage heavy-duty applications.

Note particularly the choice of mounting means. Mounting means brackets shown in drawing are Aerovox standard; other types can be supplied.

Voltage ratings from 600 to 7500 D.C.W. Widest selection of capacitance values. Impregnants and fills available are HYVOL (Vegetable) or HYVOL M (mineral oil). The exclusive Aerovox terminal construction means units that pass the standard immersion tests required by various Governmental services. Terminal assembly is non-removable, an integral part of the capacitor.

These capacitors provide maximum capacitance at minimum cost. Widely used for continuous-service in transmitters, amplifiers, rectifier filters and similar applications.

● *Literature on Request*

AEROVOX capacitors

FOR RADIO-ELECTRONIC AND INDUSTRIAL APPLICATIONS

AEROVOX CORPORATION, NEW BEDFORD, MASS., U.S.A.
SALES OFFICES IN ALL PRINCIPAL CITIES • Export: 13 E. 40th St., New York 16, N. Y.
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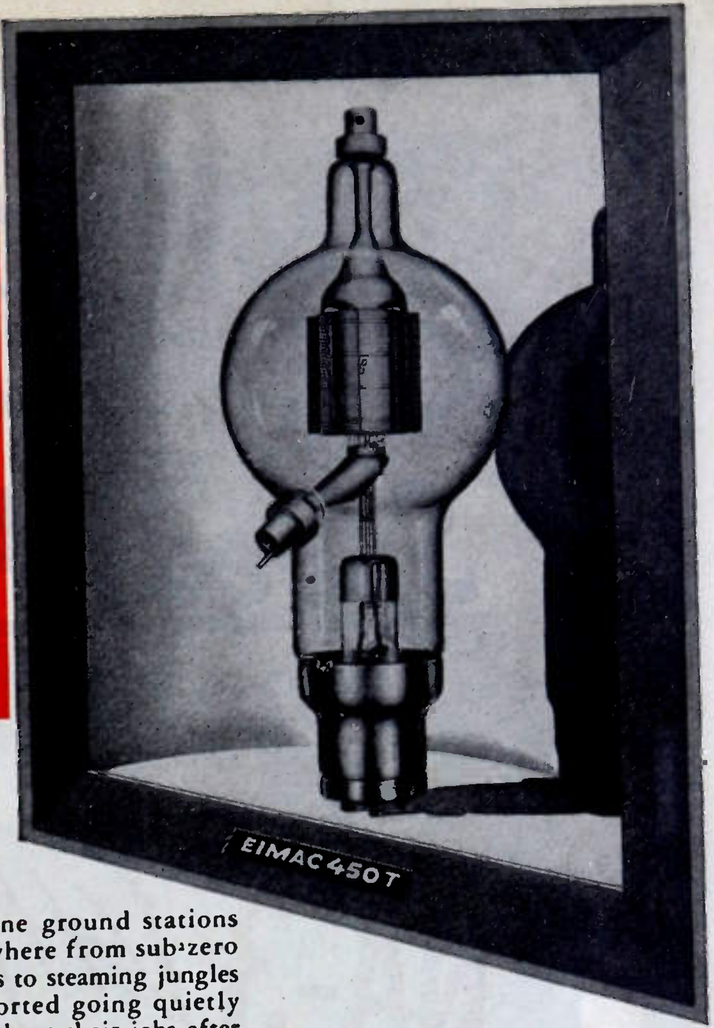
FOR AN HONORED PLACE IN THE HALL OF FAME

Perhaps no other single transmitting tube has such a great and rightful claim to fame as has the Eimac 450T triode.

This tube, one of the original members of the Eimac family, has consistently established records for plus performance in some of the world's most gruelling applications.

Long before the war the Eimac 450T established a high standard of dependability and performance in the ground stations of leading commercial airlines. Because of their outstanding dependability and inherently superior capabilities, these tubes were snapped up for wartime duty in many vital applications.

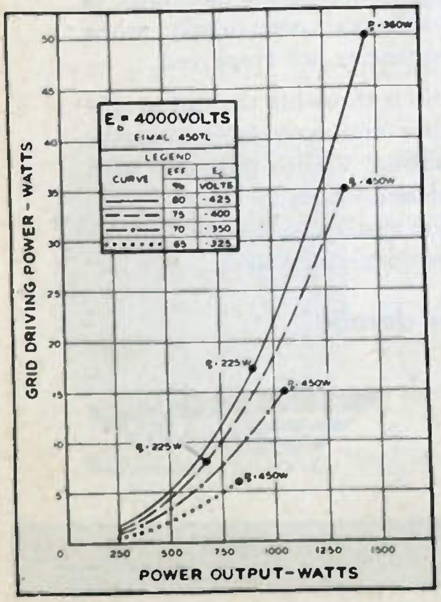


UNUSUAL VERSATILITY

The Eimac 450T is perfectly suited to a wide variety of uses as a modulator, oscillator, or amplifier. It is available as a high-mu (450TH) or low-mu (450TL) type. In every capacity, the Eimac 450T is tops in its power range; stable, rugged, and above all, proven over years of successful use.

LONG DEPENDABLE LIFE

When the first Eimac 450T's were installed in several major broadcasting stations, operators consistently reported better than 15,000 hours of service, top-notch performance. They were astounded to see such a compact tube do a giant's job. Eimac 450T's in airline ground stations located everywhere from sub-zero mountain passes to steaming jungles have been reported going quietly and efficiently about their jobs after 20,000 hours on the air!



PERFORMANCE PLUS

Performance is, after all, the ultimate criterion of electron tubes. The unusual capabilities and low interelectrode capacitances of the Eimac 450T are two of the reasons for its widespread use in 1 Kw to 5 Kw stations at frequencies up to 60 Mc. And even at frequencies up to 150 Mc, the 450T triode will provide a useful output.

HIGH POWER-GAIN

In a class B audio amplifier, a pair of Eimac 450TL's will provide 2200 watts plate power output with a driving power of but 15 watts! Or, in a class-C application, a *single* Eimac 450TL will provide an r-f plate power output of 1800 watts with but 42 watts driving power.

POST-WAR IMPROVEMENTS

The 450T, proven before war and during war, stands today as a greater tube than ever before. Post-war developments, the result of steady, intensive research in Eimac's laboratory, has brought to today's 450T new electrodes for higher thermionic efficiencies and even longer life.

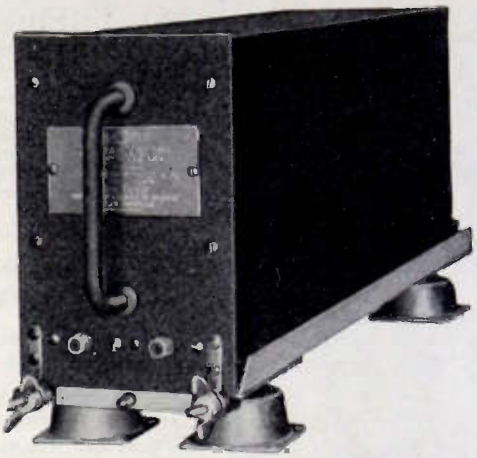
With these facts in mind, it's easy to see why the Eimac 450T is accepted over any other triode of like rating. This veteran tube has stood the acid test of time and rugged duty around the world. Today a still better 450T awaits your order. Inquire!

EITEL-McCULLOUGH, INC.
1305J San Mateo Ave., San Bruno, Calif.
Export Agents:
Frazar and Hansen, 301 Clay St., San Francisco 11, Calif.

Follow the Leaders to

Eimac REG. U.S. PAT. OFF.
TUBES

THE COUNTERSIGN OF DEPENDABILITY IN ANY ELECTRONIC EQUIPMENT



MN-85B — RECEIVER

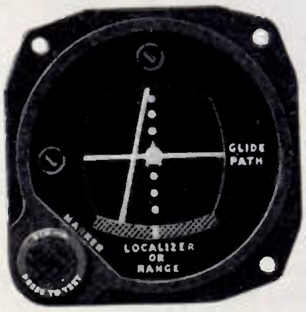
Complete coverage of navigation and communication bands, 280 crystal-controlled channels, 108—136 mc. Standard half-ATR case. "Crystal-Saver" Circuit uses only 11 crystals. Optional Dynamotor Power Supply. Simultaneous operation of *two* receivers on *one* antenna *without interference.* MR-74 Mounting Base, Bendix improved design, is equipped with positive rear plug insertion and ejection mechanism. Weight, 22 lbs.

MN-81B
REMOTE CONTROL UNIT

Concentric, internally-lighted dials indicate frequency selected in megacycles. Toggle switch selects PHASE or 90/150 cycle two-course type of localizer operation. Weight, 1.7 lbs.

MN-82B
COURSE SELECTOR

Standard 3" A-N instrument case, presenting row of easy-to-read numerals representing course selected. Includes built-in ambiguity and no-signal indicator. Weight, 1.8 lbs.



ID-48
CROSS-POINTER INDICATOR

Standard 3" A-N instrument. Vertical pointer indicates right or left deviation from range or localizer course. Horizontal pointer operated by optional glidepath receiver.

MN-69A
COURSE INDICATOR

Standard 3" A-N instrument case. Provides standard Omni-Directional Range course indication and transmits angular data to radio pointer of Radio Magnetic Indicator. Hermetically sealed for long life and low maintenance. Weight, 2 lbs.

MN-72A — RADIO-MAGNETIC INDICATOR

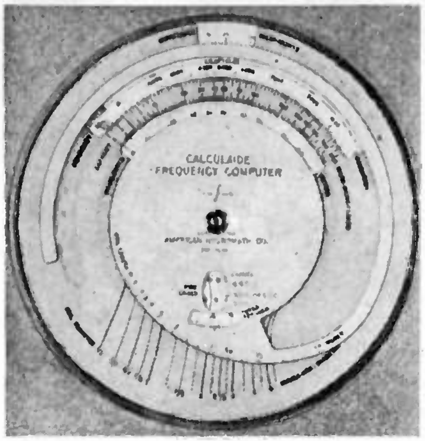
Standard 3" instrument case. Indicates remote magnetic compass and radio bearings. Presents heading-sensitive "ADF-Type" bearing indication from omni-directional range station. Hermetically sealed. Weight, 3 lbs.

MN-84B—ANTENNA Handles simultaneous operation of two receivers without interference Less drag — uniform coverage pattern for all aircraft attitudes—wide band, very low standing wave ratio.

NEWS and NEW PRODUCTS

January, 1947

Frequency Computer



Problems involving frequency, inductance are quickly solved with the "Calculaide" frequency computer devised by American Hydromath Co., 145 W. 57 St., New York 19, N. Y. This new frequency computer correlates, in one setting, the natural frequency and wave length of a circuit comprising a coil and condenser with the physical dimensions of the coil and the capacity of the condenser.

Regulated DC Power Supply

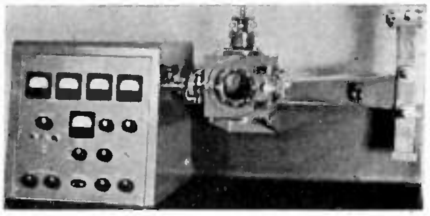
Electronically regulated DC power supplies with sufficient capacity for production and laboratory tests of AC-DC radio receivers, amplifiers and other electronic equipment, and motors and appliances normally designed for DC power line operation, are being produced by Furst Electronics, 800 W. North Avenue, Chicago 22, Ill.

Two models, 310-A and 310-B, are available. The former provides regulated AC and DC while the latter delivers only DC. Provision is made for adjustment of either type of current; and once set, the manufacturer states, the output voltage will stay constant regardless of variations of load or line voltage.

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

Electron Diffraction Instrument

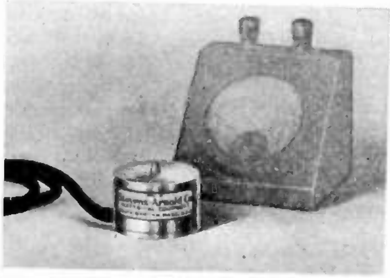
The Electronics Department of the General Electric Company, Syracuse, N. Y., announces that their Electron Diffraction Instrument is now commercially available. This unit provides a method of studying crystal and metal surfaces and is a development from apparatus first designed to establish the wave nature of electrons.



Capable of examining specimens weighing as much as 40 pounds, all the work is electronically done in a vacuum with the aid of electrostatic and magnetic fields. Accelerated electrons are focused upon the the surface, from which they are reflected forming a diffraction pattern on a photographic plate. The sample under examination may be turned in any direction for a complete study of all sides or angles.

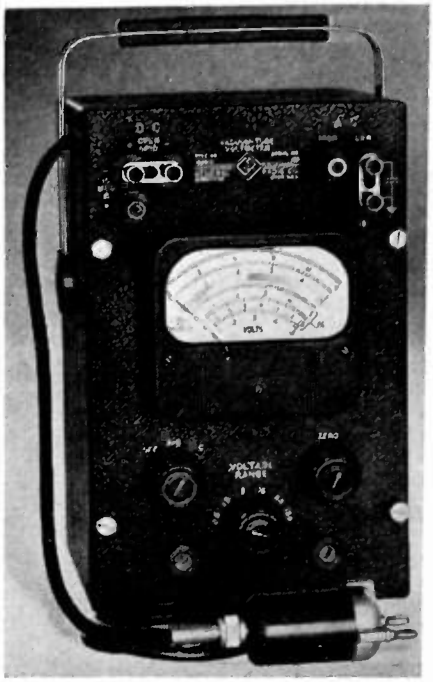
Deflection Pick-up

An hermetically sealed unit which translates minute deflections or pressure variations into lineal DC voltage changes has been announced by Stevens-Arnold Co., 22 Elkins Street, South Boston, Mass. It is claimed that accurate readings are obtained in the range of 0.0005" to 0.1".



This pick-up has an output of 75 millivolts and an internal resistance of less than one ohm so that it may be readily connected to indicating instruments.

Vacuum Tube Voltmeter



General Radio Company, 275 Massachusetts Ave., Cambridge 39, Mass., announces Type 1800-A vacuum tube voltmeter to supersede Type 726-A. This new instrument reads DC (0.01-150) as well as AC (0.1-150) voltages and can be used at much higher frequencies (500 megacycles), with a rated accuracy of ±2%. Another improvement is the single zero setting which serves for all ranges. The probe is furnished with a variety of fittings including both coaxial and banana-plug terminals and a 50-ohm disk resistor for coaxial line measurements.

Television Telemetering

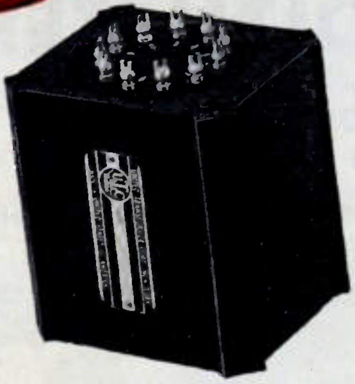
A war-developed method of checking, on the ground, vital test data from an airplane in flight was recently described by the Farnsworth Television & Radio Corp., 3700 E. Pontiac Street, Fort Wayne, Ind.

"Television-telemetering" as worked out by Farnsworth engineers in cooperation with the Curtiss-Wright Corporation and the Navy Department, enables observers at a ground station to see by television all instrument recordings and indicators of pressures, strains, and structural failures of a test plane in the air. Freed from the necessity of keeping a constant check on his test instruments, the pilot can devote his full attention to handling the plane.

(Continued on page 48A)

FOR

THE BROADCAST STATION
THE HIGH FIDELITY AMPLIFIER
THE LABORATORY



LINEAR STANDARD

Linear Standard audio units are the closest approach to the ideal component from the standpoint of frequency response, wave form distortion, efficiency, shielding, and dependability. Guaranteed response ± 1.3 DB, 20—20,000 cycles. The standard of the broadcast industry . . . units available for every audio and power application.

ULTRA COMPACT

For compact, high fidelity equipment, UTC Ultra Compact units are unequalled. Light in weight, yet providing frequency response ± 2 DB from 30 to 20,000 cycles. All units except those carrying DC in primary employ true hum balancing coil structure which, combined with high conductivity outer case, insures good inductive shielding. Units available for all audio applications up to + 10 DB in operating level.

INTERSTAGE FILTERS

UTC Interstage Filters (10,000 ohms impedance) are available in low pass (LPI), high pass (HPI), and band pass (BPI) types for all frequencies from 200 to 10,000 cycles. Designed to effect 6 DB loss at cutoff frequency . . . 35 DB at .75 and 1.5 times cutoff frequency . . . 40 DB at .5 and twice cutoff frequency. Dual alloy magnetic shielding reduces pickup to 150 Mv. per gauss.

VARITRAN

Varitran units provide an ideal means of voltage control for AC equipment. Performance features include high efficiency . . . excellent regulation . . . universal mounting features . . . self-contained fuse protection. Available in 115 volt and 230 volt models with from 1 to 11 Amp. output rating. These units afford stepless adjustment of voltage from 0 to 113% of line voltage.

United Transformer Corp.

150 VARICK STREET NEW YORK 13, N. Y.

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

LABORATORY INSTRUMENTS FOR SPEED AND ACCURACY

Sensational New

VACUUM TUBE VOLTMETER

20 cps to 700 Mc
1.3 mmfd input capacity

...the instrument you have been waiting for!

-hp- Model 410A

Far surpassing any comparable instrument, this new -*hp*- Model 410A High Frequency Vacuum Tube Voltmeter measures voltage over a wider frequency range, and at a higher input impedance than any previously available instrument.

The extremely high input impedance for ac measurements makes possible the testing of video and VHF amplifier circuits without disturbing the circuit under test. The 410A for the first time provides an instrument which will give accurate voltage measurement from audio frequency up through the micro wave regions.

The -*hp*- Model 410A is the instrument the whole electronic industry has been looking for. Your early inquiry will be best assurance of prompt delivery. Write today for more complete information — prices — delivery dates.

SPECIFICATIONS

ac Measurements

Six ranges, full scale readings 1, 3, 10, 30, 100, and 300 volts.

Input impedance, 6 megohms in parallel with 1.3 uuf.

Frequency response, 20 cps to 700 mc ± 1 db.

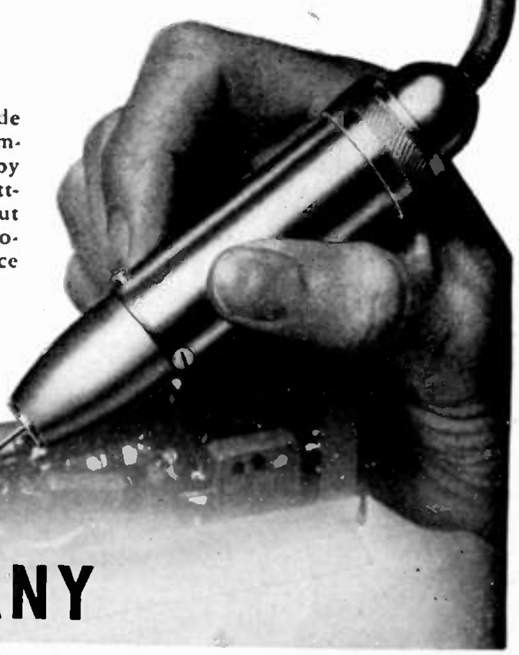
dc Measurements

Seven ranges, full scale readings 1, 3, 10, 30, 100, 300, and 1000 volts.

Input impedance, 100 megohms, all ranges.

Resistance Measurements

Seven ranges, mid scale readings 10, 100, 1000, 10,000, 100,000 ohms, 1 megohm and 10 megohms. Accuracy: ± 3%



The wide range of 410A is made possible by a special probe employing a diode developed by Eimac specifically for Hewlett-Packard. The probe has an input capacity of 1.3 micro-micro-farads, and the input resistance is 6 megohms.

...another hp achievement!

HEWLETT-PACKARD COMPANY

1159D PAGE MILL ROAD, PALO ALTO, CALIFORNIA, U.S.A

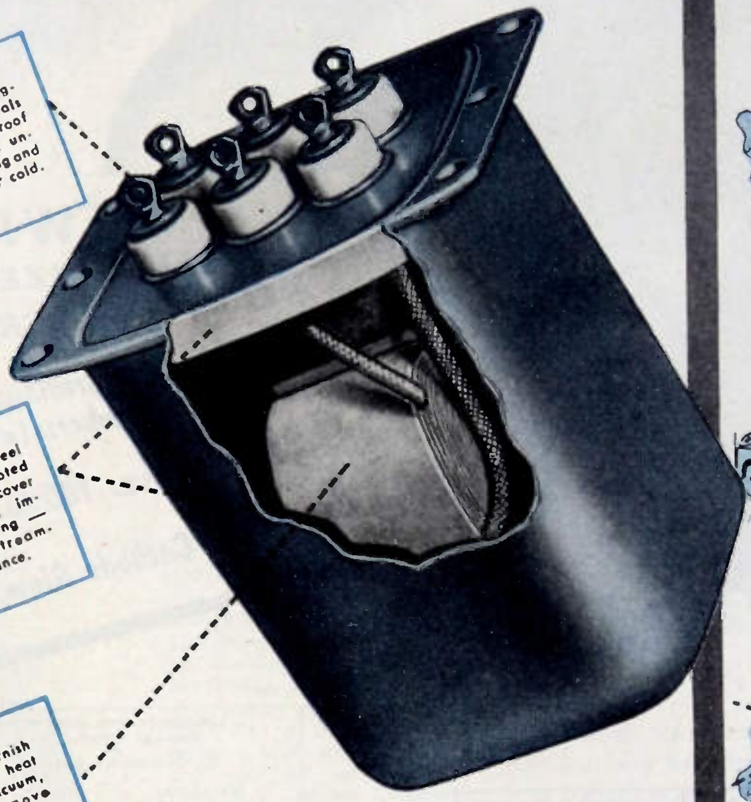
For an Extra Margin of

DEPENDABILITY

UNDER ALL OPERATING CONDITIONS

CASE TYPE

CHICAGO TRANSFORMERS *Sealed in Steel*



Exclusive C. T. bushing-gasket seal at terminals is permanently proof against moisture, is unimpaired by soldering and by climatic heat or cold.

Seamless drawn steel case and C.T.-innovated "Deep-Seal" base cover provide a strong, impenetrable housing — rust-proofed, streamlined in appearance.

Coil is wax and varnish impregnated under heat and alternating vacuum, pressure, to remove moisture, prevent its re-entrance during assembly of unit.

Sealed AGAINST ATMOSPHERIC MOISTURE AND INDUSTRIAL FUMES

THUS *Sealed* AGAINST CORROSION OF COPPER COIL WINDINGS

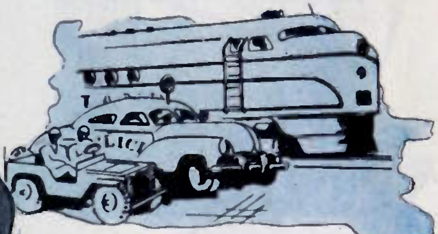
STAY *Sealed* IN EXTREMES OF HEAT AND COLD!

FITTED TO THE APPLICATIONS WHERE COMPONENT DEPENDABILITY IS ESSENTIAL TO

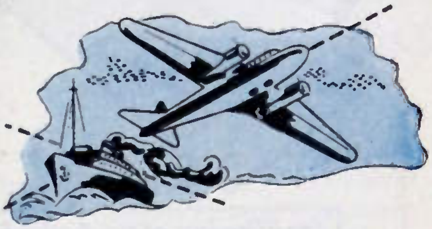
Avoid

COSTLY MAINTENANCE
LOSS OF LIFE
DISRUPTION OF A VITAL SERVICE

RADIO AND TELEVISION BROADCASTING
FIXED, MOBILE, & SATELLITE EQUIPMENT



MILITARY, POLICE, AND RAILROAD COMMUNICATIONS



ELECTRONIC NAVIGATIONAL AIDS FOR SHIPS AND AIRLINES

INDUSTRIAL CONTROLS

★In these and many other transformer applications, economy, as well as efficiency, is best served by Chicago Transformer's *Sealed in Steel* construction.

Let its assurance of long-lasting transformer reliability help make your electronic product free of component replacements and expensive servicing regardless of adverse operating conditions.

TRADE MARK REG.

CHICAGO TRANSFORMER

DIVISION OF ESSEX WIRE CORPORATION

3501 ADDISON STREET • CHICAGO 18, ILLINOIS

The Rauland *VISITRON* 10FP4/R6025

. . . the NEW Picture Tube with Unprecedented Brilliance

NEW!

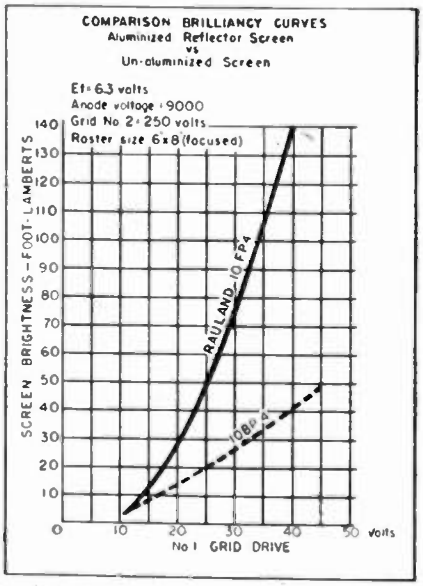
ALUMINIZED REFLECTOR-SCREEN

Doubles the Brilliance
Highlight: 75 Foot Lamberts (avge.)
Contrast Range: Over 100 to 1
No Ion Spot—No Cathode Glow

Specifications of the Rauland Visitron 10FP4/R6025

Heater Voltage	6.3 A.C. or D.C.
Focusing Method	Electromagnetic
Deflection	Electromagnetic
Deflection Angle	50 Degrees
Screen	Phosphor P4 Aluminized Reflector
Bulb Diameter (Max.)	10⅝" at screen end
Length	17⅝" ± ⅜"
Base	Small Shell Duodecal 7 Pin
Anode Terminal	Cavity
Anode Volts (Max.)	13,000
Anode Volts (Operating)	9,000

External Coating (Optional): 500 mmf.



• WRITE FOR INTERESTING BULLETIN •

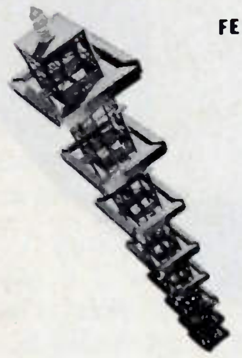
RADIO • RADAR • SOUND • Rauland **COMMUNICATIONS • TELEVISION**

Electroneering is our business

THE RAULAND CORPORATION • CHICAGO 41, ILLINOIS

With Leading Stations from Coast to Coast

it's FM by Federal



FEDERAL'S SQUARE LOOP ANTENNA MULTIPLIES POWER OUTPUT

The 8-Element Square-Loop Antenna — an outstanding Federal development — gives your FM station an effective radiated power more than eight times that of the transmitter rating. A single adjustment per loop tunes for any frequency from 88 to 108 mc. Complete antenna and rugged supporting tower are designed to withstand heavy winds and icing loads.

FEDERAL FM TRANSMITTERS FEATURE "FREQUEMATIC"* MODULATOR

The "Frequematic" modulator—an exclusive feature of every Federal FM transmitter—assures outstanding fidelity and mean carrier stability. Simple all-electronic circuits with standard receiver tubes simplify initial alignment and reduce maintenance expense. Holds center frequency within .001%. Signal-to-noise ratio reduced to 5600-to-1.

*Trade Mark

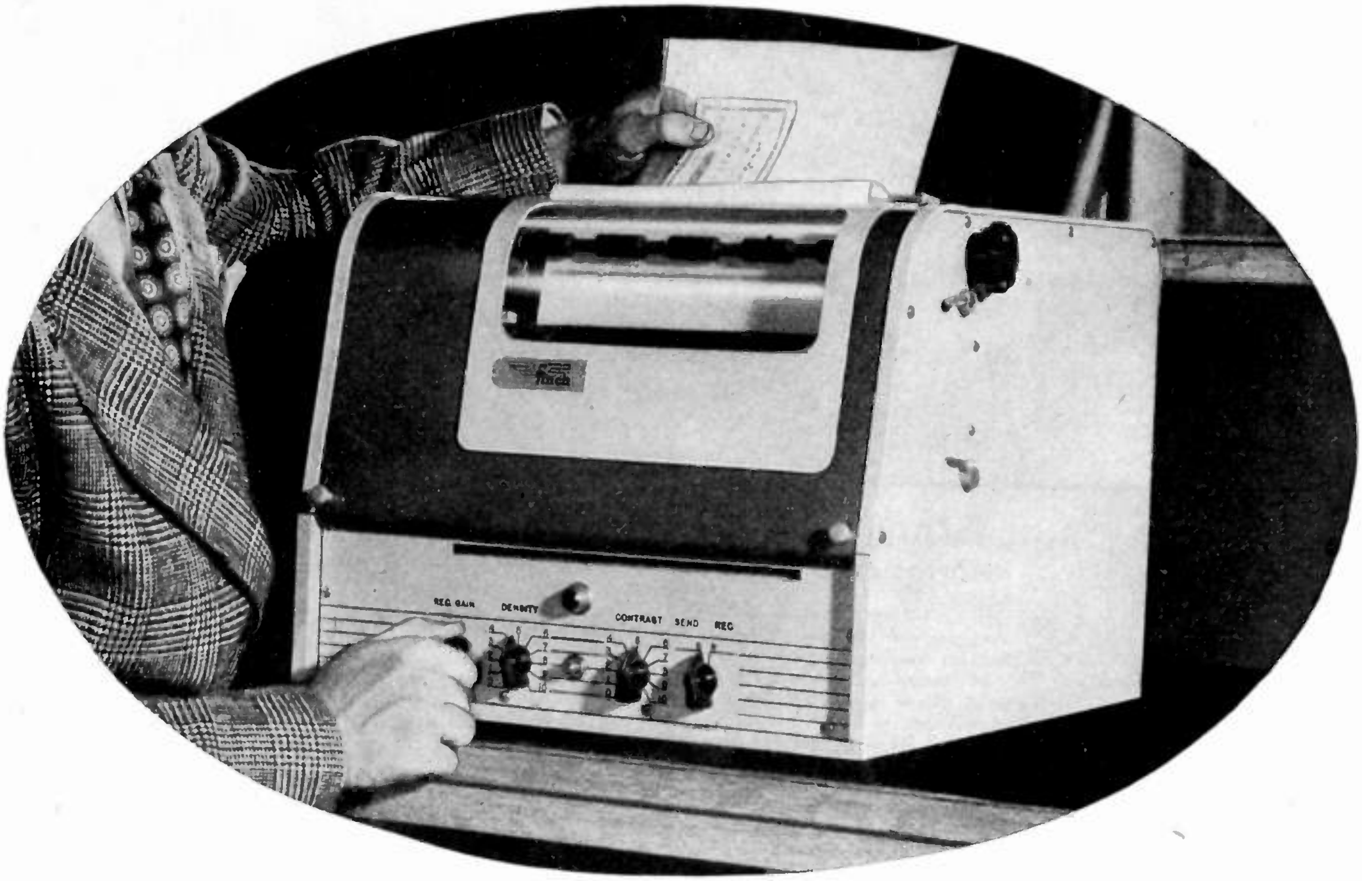
TODAY, station operators everywhere look to FM for finer broadcasting. And they look to Federal for the finest in FM. Already, Federal equipment is on the job in major FM stations throughout the country—setting new records for performance and dependability.

Remember that FM by Federal means complete equipment—and complete service, too. For Federal can supply your entire transmitting system, from microphone to antenna—all components precision engineered and designed to work together. And Federal will see the job through until your station is *on the air*. Factory-trained engineers will supervise installation, tune the equipment, and instruct your personnel in its operation and maintenance—all without extra charge. For complete information, write today to Dept. B237.

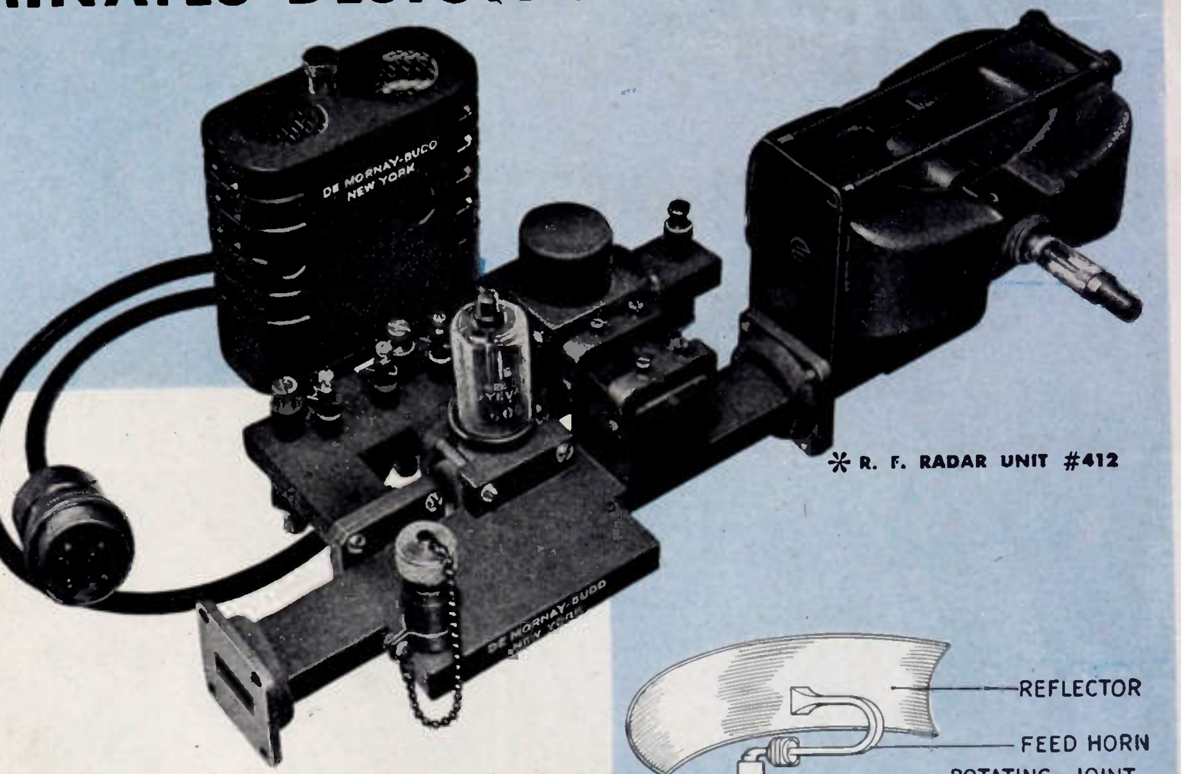
Federal Telephone and Radio Corporation

In Canada:—Federal Electric Manufacturing Company, Ltd., Montreal.
Export Distributors:—International Standard Electric Corp. 67 Broad St., N.Y.C.

Newark 1, New Jersey


finch
REG. GAIN
DENSITY
CONTRAST
SEND
REC

PACKAGED R. F. RADAR ASSEMBLY ELIMINATES DESIGN HEADACHES



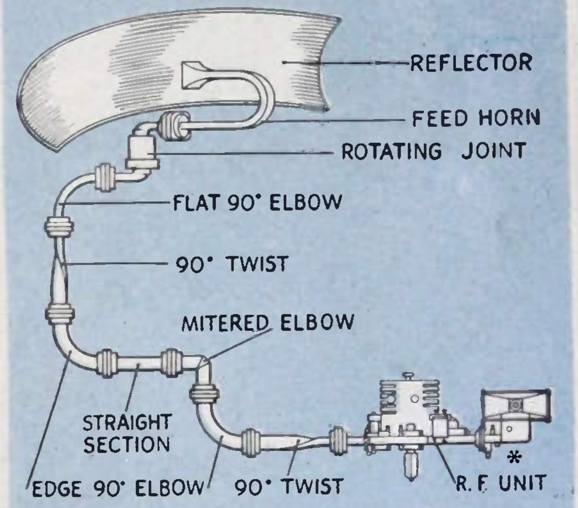
✱ R. F. RADAR UNIT #412

The DeMornay-Budd packaged R. F. Unit provides a complete R. F. assembly for microwave radar. It is now possible to obtain as standard items all the microwave R. F. components necessary in the fabrication of a complete radar—DeMornay-Budd Standard Transmission Line Components plus packaged R. F. Unit.

The R. F. Radar Unit is delivered complete and ready to operate. It is wired and contains all the necessary tubes and crystals. The unit uses a packaged magnetron capable of delivering 20 kw., peak power, at 9375 mc. Two type 2K25 local oscillator tubes are provided, one for receiver and A.F.C. and the other for beacon operation. A type 1B35 A-T-R tube, a type 1B24 T-R tube and the necessary type 1N21 crystals are included in the assembly. A 20 db. directional coupler permits accurate measurements to be made at any time with a maximum of convenience and safety.

Since the use of radar beacons is contemplated in the near future, the unit has been designed with a beacon cavity and crystal mount. The unit can be supplied without the beacon cavity and crystal mount and beacon local oscillator, and a termination supplied in their place so that it becomes a simple matter to convert to beacon operation when necessary.

NOTE: Write for complete catalog of De Mornay·Budd Standard Components and Standard Bench Test Equipment. Be sure to have a copy in your reference files. Write for it today.



R. F. Radar unit #412 (indicated by asterisk) used in conjunction with standard DeMornay-Budd transmission line components.

DE MORNAY
BUDD

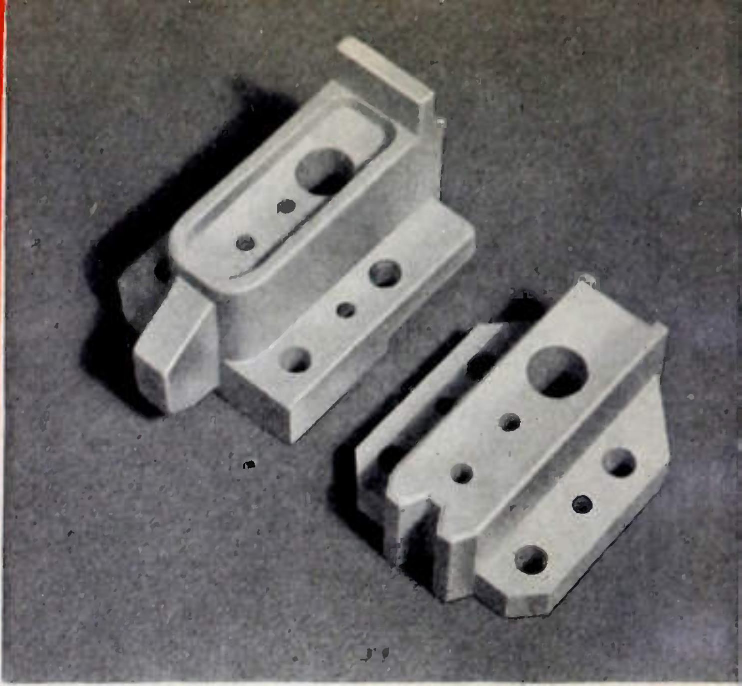
EQUIPMENT
FOR
97% OF ALL
RADAR SETS

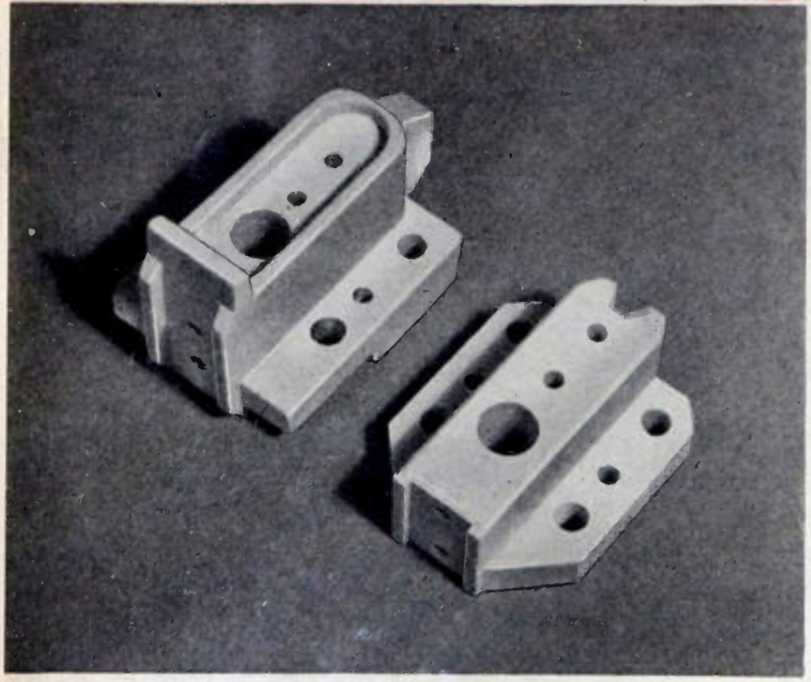
DE MORNAY-BUDD, INC.
475 GRAND CONCOURSE, NEW YORK, N. Y.

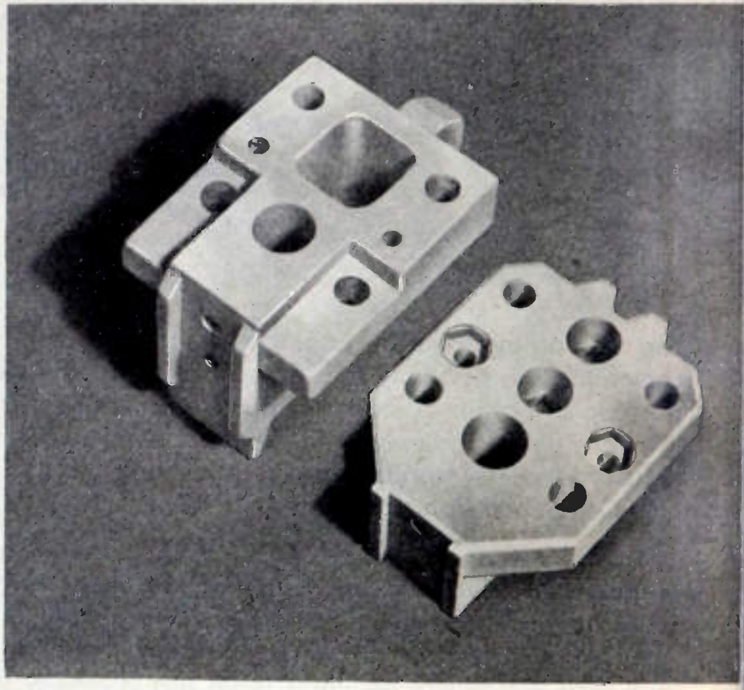
Redesign

IMPROVED PERFORMANCE

CUT COSTS 62%







Photographs courtesy JAMES R. KEARNEY CORP., St. Louis, Mo.

THE designing engineer who specifies ceramics knows exactly what is required, but the special knowledge and experience of the ceramic engineer frequently enables him to make design suggestions for low production cost.

Maximum efficiency and maximum production at lowest cost result when the designing engineer and the production engineer work together. The illustrations above show the results of such collaboration. The original and the final designs are shown in three perspectives.

Redesign of this part cut costs 62%. Maximum deliveries in minimum time were made possible. In addition, the customer said: "Your suggestions on redesign greatly strengthened this component, reduced its size and weight, and increased its utility."

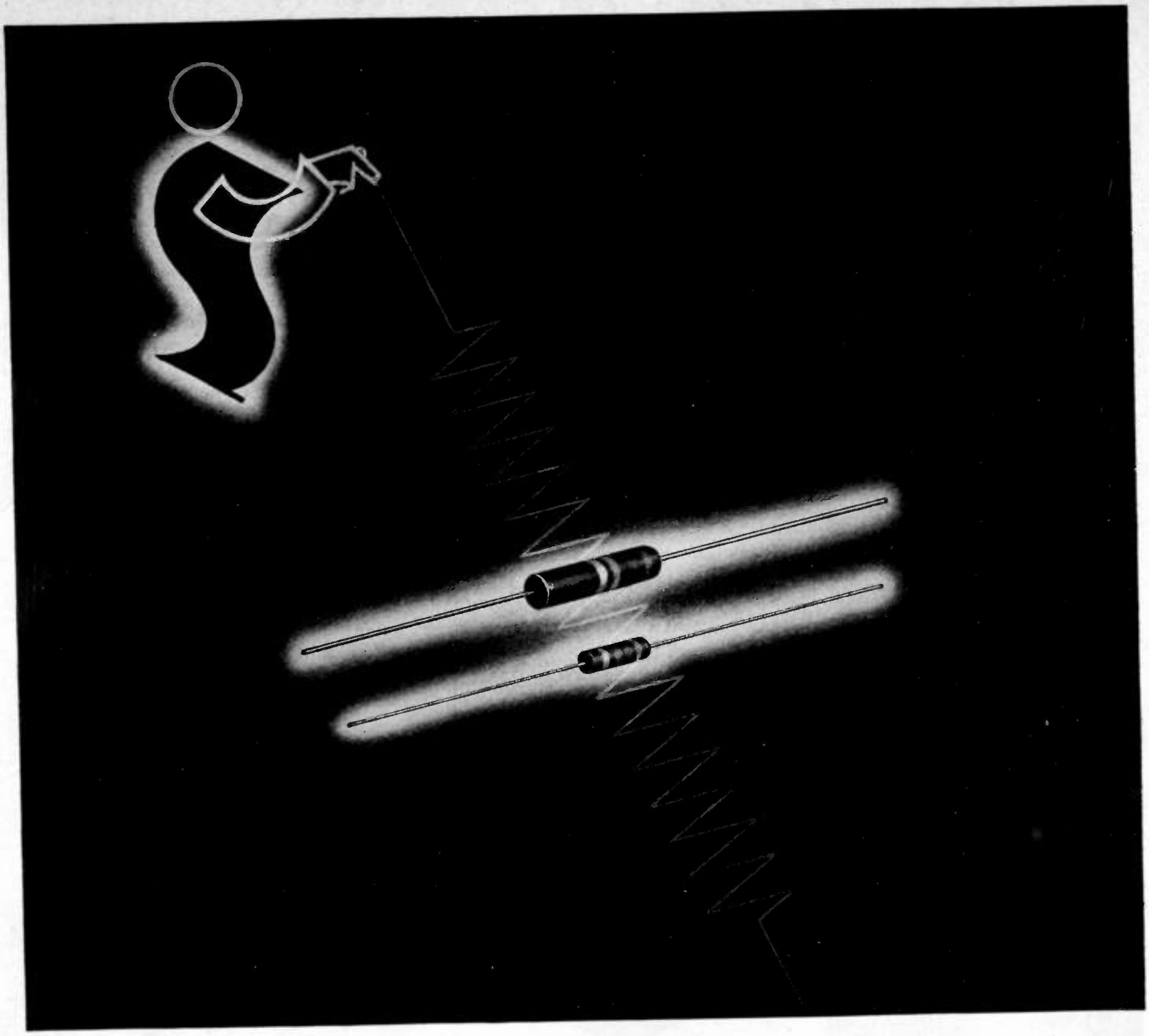
American Lava Corporation engineers will gladly cooperate with you in developing your ideal design in AlSiMag custom made technical ceramics. In their 44th year of practical experience they offer production facilities and techniques that can be most valuable to any user of custom made ceramics.

NEW PROPERTY CHART—A new property chart giving the physical, electrical and mechanical properties of the more frequently used AlSiMag technical ceramic compositions will be mailed without charge on your request.

ALSIMAG
TRADE MARK REGISTERED U.S. PATENT OFFICE

AMERICAN LAVA CORPORATION
CHATTANOOGA 5, TENNESSEE
44TH YEAR OF CERAMIC LEADERSHIP

ENGINEERING SERVICE OFFICES:
ST. LOUIS, Mo., 1123 Washington Ave., Tel: Garfield 4959 • NEWARK, N. J., 671 Broad Street, Tel: Mitchell 2-8159
CAMBRIDGE, Mass., 38-B Brattle St., Tel: Kirkland 4498 • CHICAGO, 9 S. Clinton St., Tel: Central 1721
SAN FRANCISCO, 163 Second St., Tel: Douglas 2464 • LOS ANGELES, 324 N. San Pedro St., Tel: Mutual 9076



REVERE FREE-CUTTING COPPER ROD

...INCREASES ELECTRONIC PRODUCTION

SINCE its recent 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.

REVERE

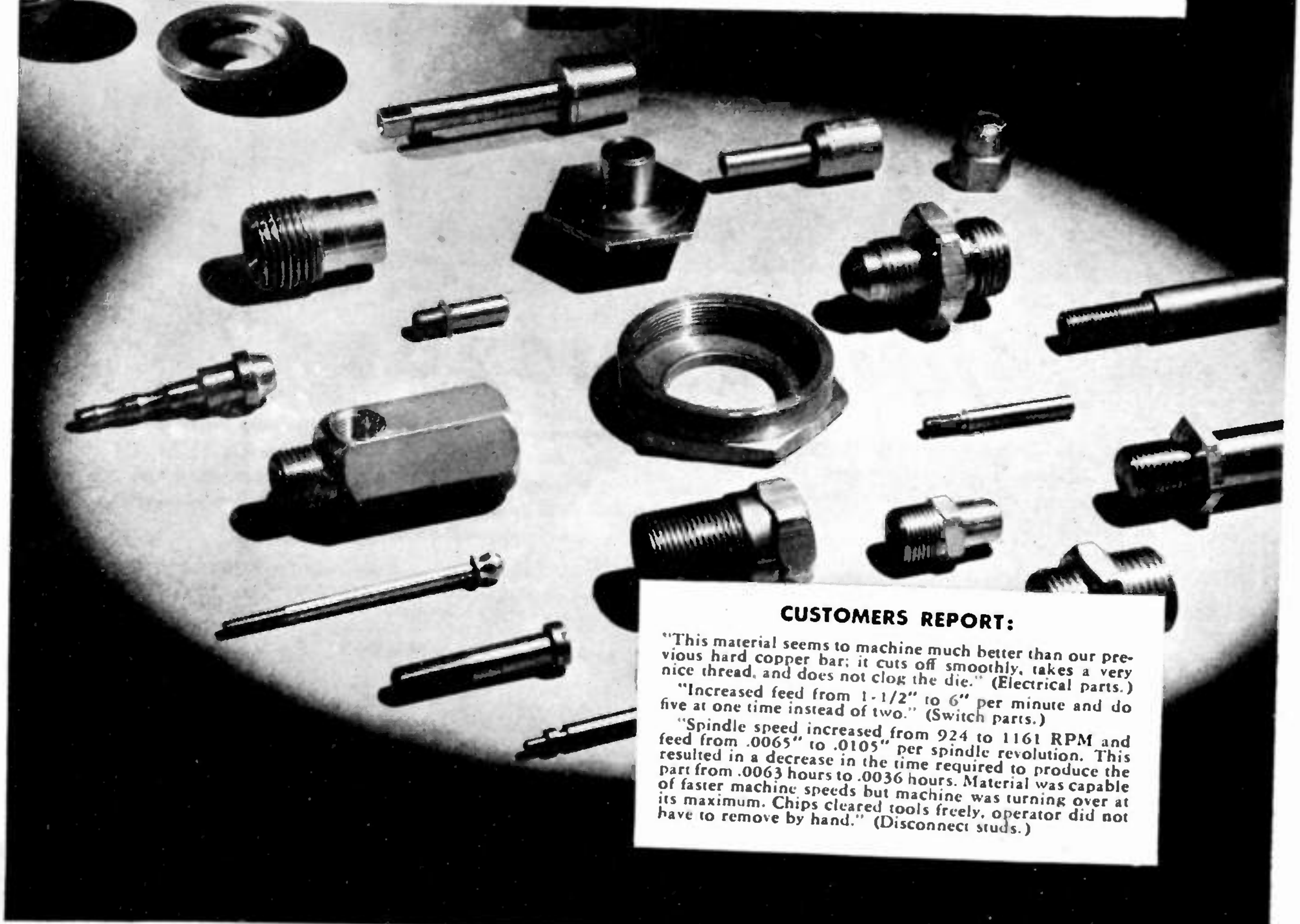
COPPER AND BRASS INCORPORATED

Founded by Paul Revere in 1801

230 Park Avenue, New York 17, New York

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

Listen to Exploring the Unknown *on the Mutual Network every Sunday evening, 9 to 9:30 p.m., EST.*



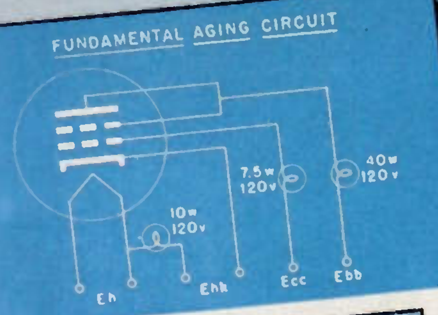
CUSTOMERS REPORT:

"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.)

MAKING TUBES IS EASY.. *If* YOU KNOW HOW!


FUNDAMENTAL AGING CIRCUIT

AGING SCHEDULE FOR HYTRON 50L6GT

Step	Minutes	Eh a-c	Ehk a-c	Ecc d-c	Ebb d-c
1	5	50	110	0	0
2	3	70	110	0	0
3	5	80	110	0	0
4	3	80	110	120	120
5	5	70	0	0	0
6	4	0	0	-10	120
7	5	50	0		

Electrode potentials are varied as shown in the schedule. Actual voltages at the socket depend on currents drawn through the incandescent lamps used as economical, interchangeable current-limiting resistors.

Operations performed in seven steps are: (1) discovery of heater-cathode shorts (2) beginning of cathode processing to stabilize emission (3) further seasoning and burning off of h-k leakage (4) h-k potential increased to eliminate leakage (5) grid, screen, and plate potentials applied to complete de-gassification (6) cooling off period (7) normal potentials applied to pre-heat for test.

AUTOMATIC AGING FOR BETTER TUBES

Yes, radio tubes also must be "aged in the wood." Aging activates the cathode under accelerated life conditions, just before test. In the fundamental aging circuit shown, final seasoning and de-gassification stabilize characteristics in accordance with the carefully planned aging schedule.

Formerly tubes were plugged into long aging racks. An operator, equipped with the schedule and a timer, adjusted electrode potentials throughout the aging cycle. The human element resulted in errors of timing and switch manipulation.

Hytron's new automatic aging wheel minimizes human error. A motor drives a mechanically-indexing horizontal wheel on which 30 radial sections of 12 tubes each are slowly rotated. Brushes contacting commutator segments automatically apply electrode potentials. The wheel itself requires no operator. The final basing machine operator feeds the wheel. Tubes already pre-heated are removed by the test operator.

Other features of the aging wheel are elimination of needless handling, fast and steady pacing of the work, easy servicing, and readily interchangeable load lamps.

To you this automatic aging wheel means economical, more uniform tubes with stable electrical characteristics. Again Hytron know-how takes a forward step by making your tubes easier and better.

SPECIALISTS IN RADIO RECEIVING TUBES SINCE 1921

HYTRON
RADIO AND ELECTRONICS CORP.

MAIN OFFICE: SALEM, MASSACHUSETTS

. . . Impromptu Discussions about Miniature Tubes

"This New Year stuff makes you kind of think of the changes and improvements that are taking place. Now that we are working so much more with miniatures I don't know how we ever got along without them. Take the Tung-Sol 35W4 power rectifier that replaces the old 35Z5GT.

"You know this tube was designed especially for series heater operation in five or six tube ac/dc receivers. The 35W4 is a good job . . . no other rectifier will give you all its features. Heater voltages add up . . . you don't need to throw away power in line cords, ballast tubes, or resistors. And for rectification efficiency . . . say, the 35W4 will make an ac/dc receiver perform just as effectively on dc as on ac power. Since you must have a pilot light, the tapped heater circuit of the 35W4 gives you a fool-proof system which minimizes lamp surges yet gives good illumination.

"High ambient temperature is another point. You have to consider it in regard to its effects on the rectifier as well as other circuit components and on fire underwriters standards. The 35W4 can tolerate an ambient as high as 150° C. This, of course, fits in with the compactness of the resulting equipment. The smaller the tubes, the closer all parts are assembled hence the higher the temperatures.

"The use of two 35W4's is perfectly practical for voltage doubler applications. At 117 volts input, a full-wave doubler delivers 100 ma. at 230 volts to the filter input or 100 ma. at 210 volts with the half-wave doubler. Think what this means in terms of power output in 'transformerless' ac power amplifiers! When you have a circuit demanding up to 4.5 amperes peak current for very short intervals look into the 35W4 . . . there go the whistles, Happy New Year—Everybody!"

TUNG-SOL

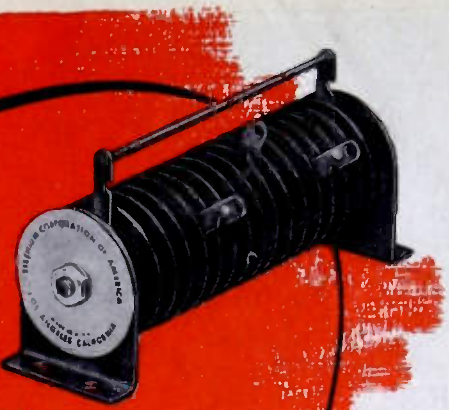
vibration-tested

ELECTRON TUBES

TUNG-SOL LAMP WORKS, INC., NEWARK 4, NEW JERSEY

Sales Offices: Atlanta · Chicago · Dallas · Denver · Detroit · Los Angeles · New York

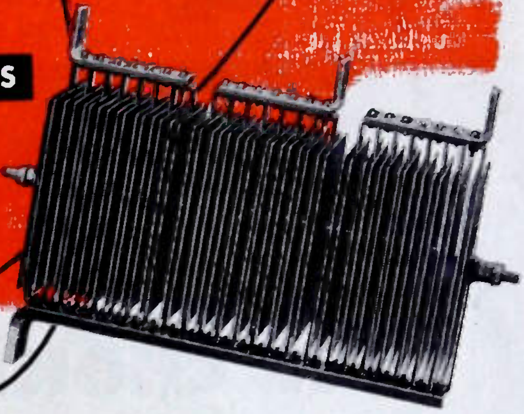
Also Manufacturers of Miniature Incandescent Lamps, All-Glass Sealed Beam Headlight Lamps and Current Intermittors



SCA rectifiers of tomorrow

IN DESIGNS OF TODAY!

POWER RECTIFIERS

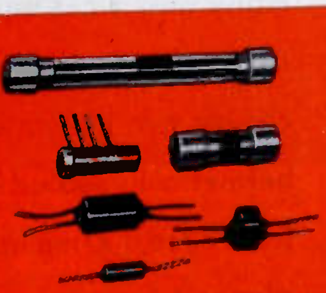


SCA Selenium Rectifiers are ENGINEERED FOR ENGINEERS. Improved performance at lower costs through ENGINEERED adaptability. Selenium Corporation of America meets exacting specifications of modern electronic developments. Manufacturers of a broad line of Selenium Power and Instrument Rectifiers, Self generating Photo-Electric Cells and allied scientific products.

Selenium Rectifiers are rapidly becoming standard in industry for all rectifier applications. Selenium Corporation of America's engineering experience can be called upon for the development and production of special rectifiers for any application.

CHECK THESE OUTSTANDING FEATURES:

- ✓ Permanent characteristics
- ✓ Adaptability to all types of circuits and loads
- ✓ Unlimited life—no moving parts
- ✓ Immunity to atmospheric changes
- ✓ High efficiency per unit weight
- ✓ From 1 volt to 50,000 volts rms.
- ✓ From 10 micro-amperes to 10,000 amperes
- ✓ Economical—simple to install—no maintenance cost
- ✓ Hermetically sealed units available



INSTRUMENT RECTIFIERS

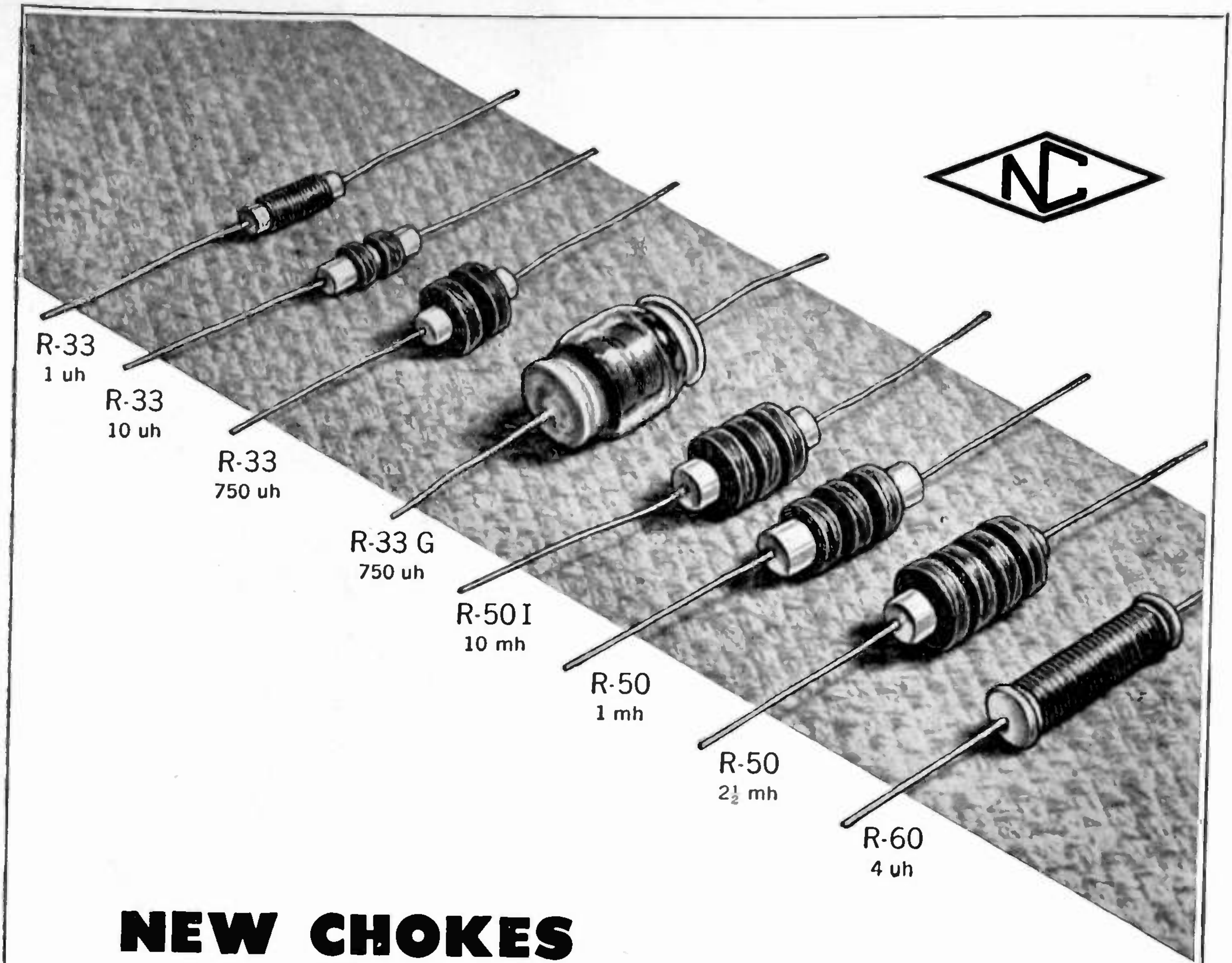
PHOTO-ELECTRIC CELLS

46-E

SELENIUM CORPORATION OF AMERICA

Affiliate of VICKERS Incorporated

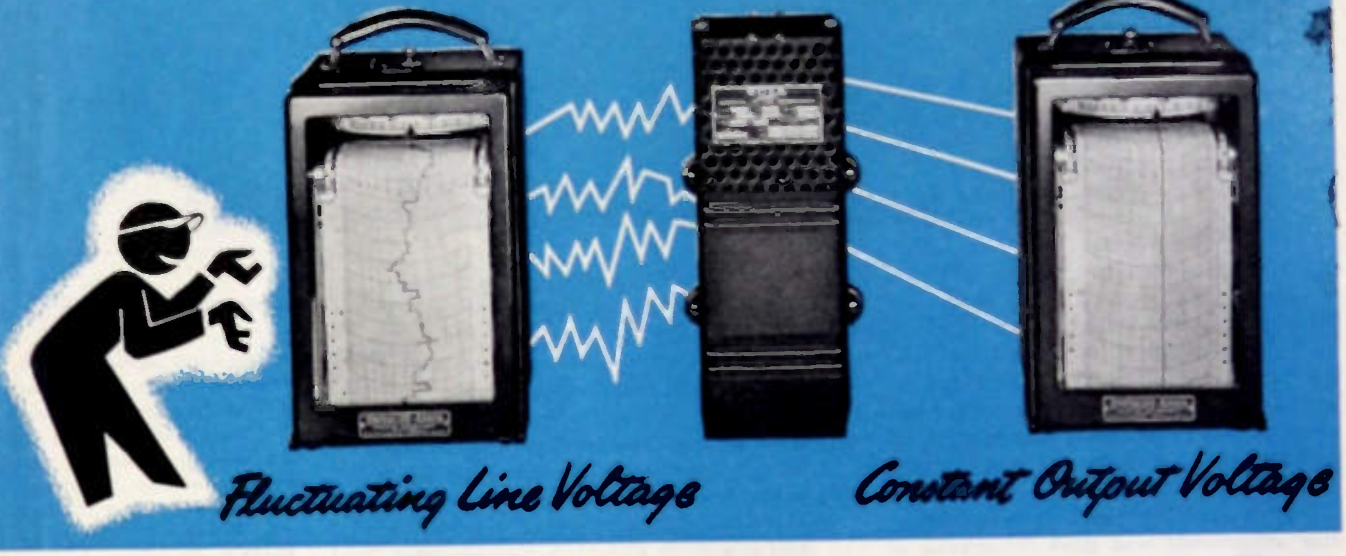
1719 WEST PICO BOULEVARD • LOS ANGELES 15, CALIFORNIA


R-33
1 uh
R-33
10 uh
R-33
750 uh
R-33 G
750 uh
R-50 I
10 mh
R-50
1 mh
R-50
2½ mh
R-60
4 uh

"WHEN EQUIPMENT IS STILL IN THE BLUEPRINT STAGE...

AND VOLTAGE VARIATIONS MAKE YOU RANT AND RAVE

...LET A SOLA 'CV' MAKE THAT VOLTAGE BEHAVE"



It may look good on paper . . . and perform superbly under regulated laboratory voltages, BUT . . . when it encounters the *unstable* voltages that are available to your customers, what happens—

—to costly filaments and tubes?
—to precision parts?
—to sensitive, balanced circuits?
—to over-all efficiency?
—to customer good-will?

The operating voltage you specify will never be consistently available unless you make provision for it. That can be done most economically and satisfactorily by including an automatic, self-protecting SOLA Constant Voltage Transformer as a "built-in" component of your equipment.

There are many standard models in SOLA Constant Voltage Transformers that have been specifically designed for built in use. They are being successfully used today by many manufacturers of electrically energized equipment who have *guaranteed* the availability of constant rated voltage. May we make a recommendation for your equipment?

Write for Bulletin
KCV-102

...Here is the answer to unstable voltage problems.

SOLA *Constant Voltage* TRANSFORMERS

Transformers for: Constant Voltage · Cold Cathode Lighting · Mercury Lamps · Series Lighting · Fluorescent Lighting · X-Ray Equipment · Luminous Tube Signs · Oil Burner Ignition · Radio · Power · Controls · Signal Systems · etc. **SOLA ELECTRIC COMPANY, 2525 Clybourn Avenue, Chicago 14, Illinois**

Manufactured in Canada under license by FERRANTI ELECTRIC LIMITED, Toronto

The remarkable current-generating efficiency of *Western Electric* SOUND-POWERED TELEPHONES is based on critical electronic positioning of magnetic elements visually indicated by the . . .

DU MONT Type 208-B *Oscillograph*

In the manufacture of Sound-Powered Telephones, made for our Army and Navy by Western Electric during the war, one of the most difficult operations was the positioning of the armature *in the exact center* of the magnetic circuit. The balanced armature design means that the armature must be *balanced magnetically among four air gaps.*

During the early years of manufacture, this adjustment was done by means of feeler gauges. This method, however, did not always yield the true magnetic center because of slight differences in magnetic materials. Many units would be rejected in subsequent tests.

To overcome this condition, production engineers at Western Electric devised a method whereby the armature can be accurately located magnetically, with the aid of the cathode-ray oscillograph. This ability *to see electronically* what goes on in these air gaps has been of considerable assistance in overcoming one of the most difficult operations associated with the production of Sound-Powered Telephones.

And so another cathode-ray oscillograph application. Surely in your own production and inspection routine, or in maintenance or servicing, or again in research and engineering, you can use this "electronic seeing" to profitable advantage

Submit your problem.

NO BATTERIES REQUIRED!

As the name implies, Sound-Powered Telephones require no external source of power. The feeble sound power of the voice is converted into electrical energy. Telephonic communication up to 10 miles is feasible over ordinary Army field-telephone wires, and up to 100 miles under laboratory conditions. Such equipment was extensively used by our armed forces afloat and ashore.

DUMONT *Precision Electronics & Television*

ALLEN B. DuMONT LABORATORIES, INC., PASSAIC, NEW JERSEY • CABLE ADDRESS: ALBEEDU, PASSAIC, N. J., U. S. A.

Questions—

1. *Name the ideal dielectric for all-high frequency currents.*
2. *Does Mykroy maintain low loss factor?*
3. *Will it hold to tolerances up to ±.001?*
4. *Is it mechanically stable?*
5. *Can it be machined or molded?*
6. *Will Mykroy warp?*
7. *Does it carbonize under electric arcs?*

Examination

1. Mykroy 2. Yes
3. Yes 4. Yes 5. Both
6. No 7. No

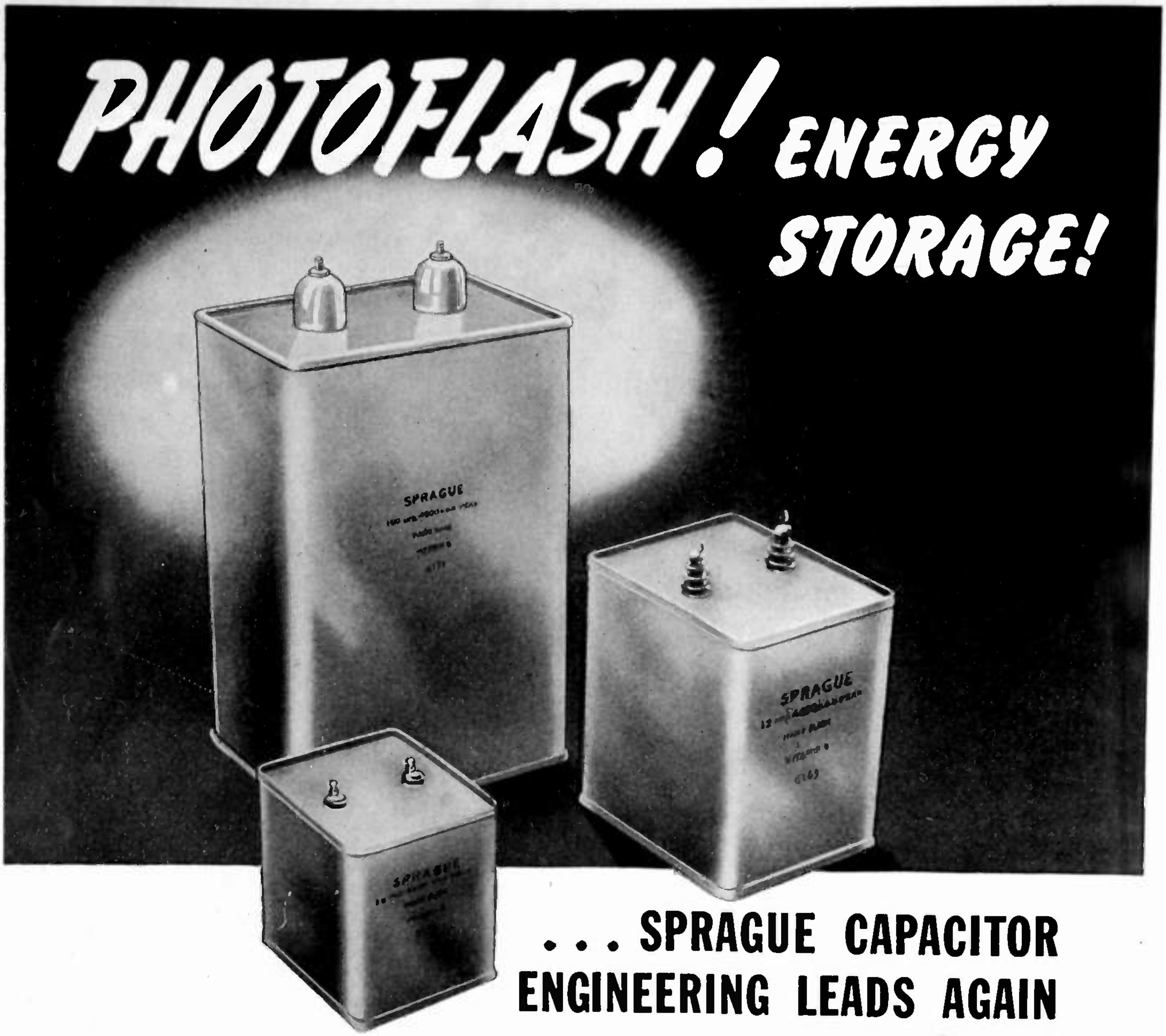
Tests are never dreaded ordeals for Mykroy. With performance characteristics such as these, it's no wonder Mykroy breezes right thru them with a 100% rating every time. That's why this Perfected Mica Ceramic insulation likes to be "put to the test".

Mykroy performance is easily demonstrated . . . will satisfy your most exacting insulation requirements. Just write for a sample . . . submit it to the most critical examination and watch it pass the test *in your own laboratory.*

MYKROY IS SUPPLIED IN SHEETS AND RODS . . . MACHINED OR MOLDED TO SPECIFICATIONS

MADE EXCLUSIVELY BY **ELECTRONIC MECHANICS INC.**

70 CLIFTON BOULEVARD • CLIFTON, NEW JERSEY
Chicago 47: 1917 NO. SPRINGFIELD AVENUE . . TEL. Albany 4310
Export Office: 89 Broad Street, New York 4, N. Y.



. . . SPRAGUE CAPACITOR ENGINEERING LEADS AGAIN

Photoflash units for war applications used Sprague *VITAMIN Q Capacitors—*because only Vitamin Q Capacitors could withstand the severe service conditions encountered.*

Privileged to work with the inventors of photoflash photography from its early inception, Sprague engineers have contributed materially to its post-war development. Not only has the present line of capacitors impregnated with the famous and exclusive Vitamin Q dielectric established new standards of compactness, light weight and dependability for electric flash tube (photoflash) photography; equally important it paves the way for outstanding economies and greater efficiency for capacitors for flash welding and time control circuits where duty cycles other than photoflash conditions prevail.

GET THE FACTS!

PHOTOFLASH ENERGY-STORAGE DATA BULLETIN ON REQUEST

Write for Bulletin #3205. Contains specifications and performance data on Sprague Vitamin Q Capacitors —also helpful, up-to-the-minute information on photoflash problems.

SPRAGUE

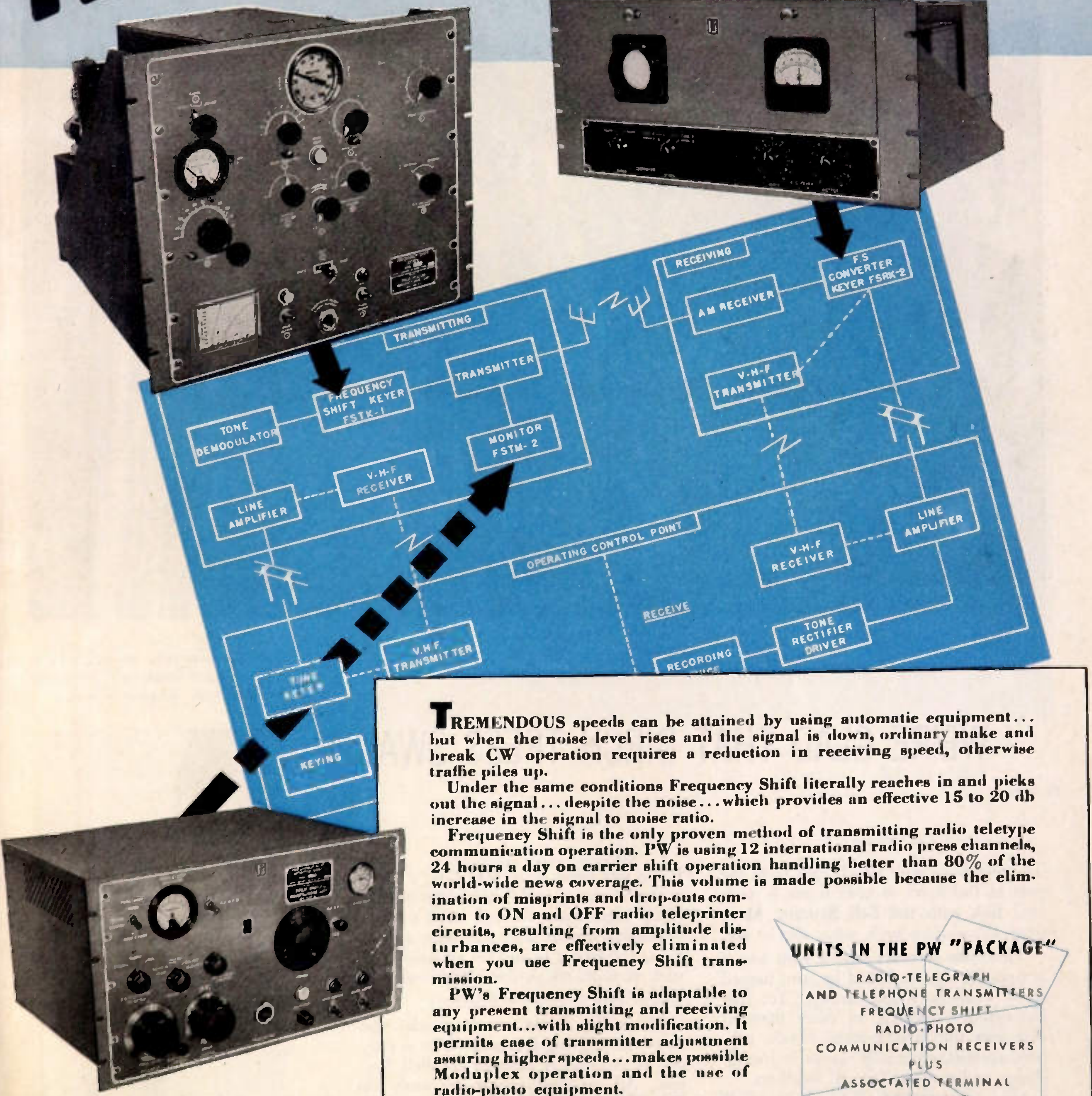
PIONEERS OF ELECTRIC AND ELECTRONIC PROGRESS

SPRAGUE ELECTRIC CO., North Adams, Mass.

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

NEW! PW "FREQUENCY SHIFT"

THE MOST DEPENDABLE RADIO-TELETYPE OPERATION *Known Today*



TREMENDOUS speeds can be attained by using automatic equipment... but when the noise level rises and the signal is down, ordinary make and break CW operation requires a reduction in receiving speed, otherwise traffic piles up.

Under the same conditions Frequency Shift literally reaches in and picks out the signal... despite the noise... which provides an effective 15 to 20 db increase in the signal to noise ratio.

Frequency Shift is the only proven method of transmitting radio teletype communication operation. PW is using 12 international radio press channels, 24 hours a day on carrier shift operation handling better than 80% of the world-wide news coverage. This volume is made possible because the elimination of misprints and drop-outs common to ON and OFF radio teleprinter circuits, resulting from amplitude disturbances, are effectively eliminated when you use Frequency Shift transmission.

PW's Frequency Shift is adaptable to any present transmitting and receiving equipment... with slight modification. It permits ease of transmitter adjustment assuring higher speeds... makes possible Moduplex operation and the use of radio-photo equipment.

For further information concerning Frequency Shift write Dept. 709A, Press Wireless Mfg., Corp., Executive Offices, 1475 Broadway, New York 18, N.Y., USA

UNITS IN THE PW "PACKAGE"

RADIO-TELEGRAPH
AND TELEPHONE TRANSMITTERS
FREQUENCY SHIFT
RADIO-PHOTO
COMMUNICATION RECEIVERS
PLUS
ASSOCIATED TERMINAL
EQUIPMENT

Your installation is engineered from any combination of the above standardized PW units

PRESS WIRELESS INC. PREWI

PW

PRESS WIRELESS

First in "Packaged" Communications Equipment

Drop-wire undergoing abrasion tests in birch thicket "laboratory." Below, the new drop-wire, now being installed.

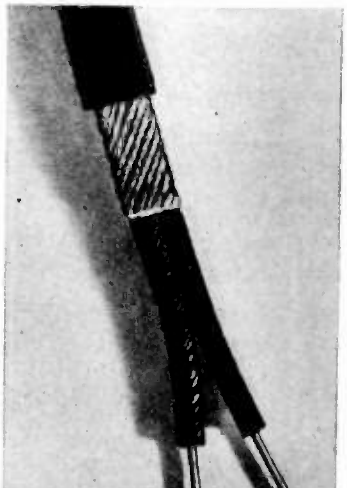
WE'RE GLAD THAT BIRCH TREES SWAY

The telephone wire which runs from the pole in the street to your house is your vital link with the Bell System. More than 17,000,000 such wires are in use.

The wire becomes coated with ice; it is ripped by gales, baked by sun, tugged at by small boys' kite strings. Yet Bell Laboratories research on every material that goes into a drop-wire—metals, rubbers, cottons, chemicals—keeps it strong, cheap, and ready to face all weathers.

Now a new drop-wire has been developed by the Laboratories which lasts even longer and will give even better service. It has met many tests, over 6 or 7 years, in the laboratory and in field experiments. It has been strung through birch thickets—rubbed, winters and summers, against trees, and blown to and fro by winds. In such tests its tough cover lasts twice as long as that of previous wires.

House by house, country-wide, the new wire is going into use. Wire is only one of millions of parts in the Bell System. All are constantly under study by Bell Telephone Laboratories, the largest industrial laboratory in the world, to improve your telephone service.



BELL TELEPHONE LABORATORIES

EXPLORING, INVENTING, DEVISING AND PERFECTING FOR CONTINUED IMPROVEMENTS AND ECONOMIES IN TELEPHONE SERVICE

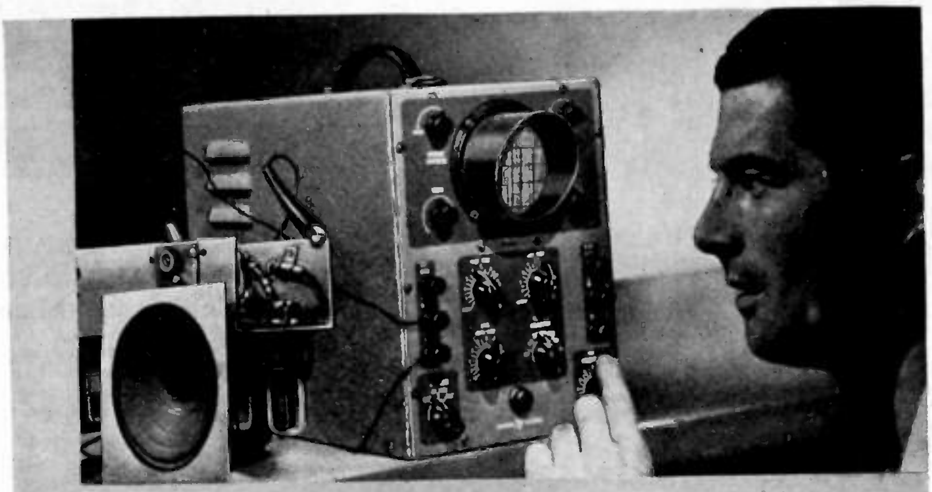
SYLVANIA NEWS

CIRCUIT ENGINEERING EDITION

JAN. *Prepared by* SYLVANIA ELECTRIC PRODUCTS INC., Emporium, Pa. 1947

NEW OSCILLOSCOPE DEVELOPED TO HELP SOLVE PROBLEMS MET IN RADIO AND ELECTRONIC EQUIPMENT

Latest Sylvania Instrument Especially Useful For Rapid Receiver Alignment and Trouble-Shooting



Oscilloscope, Type 131; cabinet is steel constructed, properly ventilated with louvers, finished in durable, attractive pearl-grey baked enamel.

In anticipation of the need for greater accuracy in trouble-shooting, alignment, distortion locating, etc., Sylvania has developed the Oscilloscope, Type 131. This accurate measuring device takes its place beside the numerous radio and electronic devices that have been Sylvania-developed to facilitate the solving of problems encountered in radios and electronic equipment.

CHARACTERISTICS AND SPECIAL FEATURES

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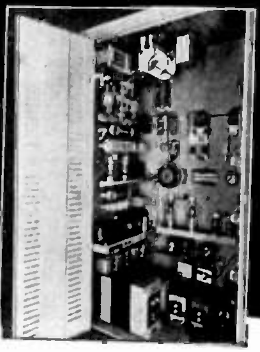

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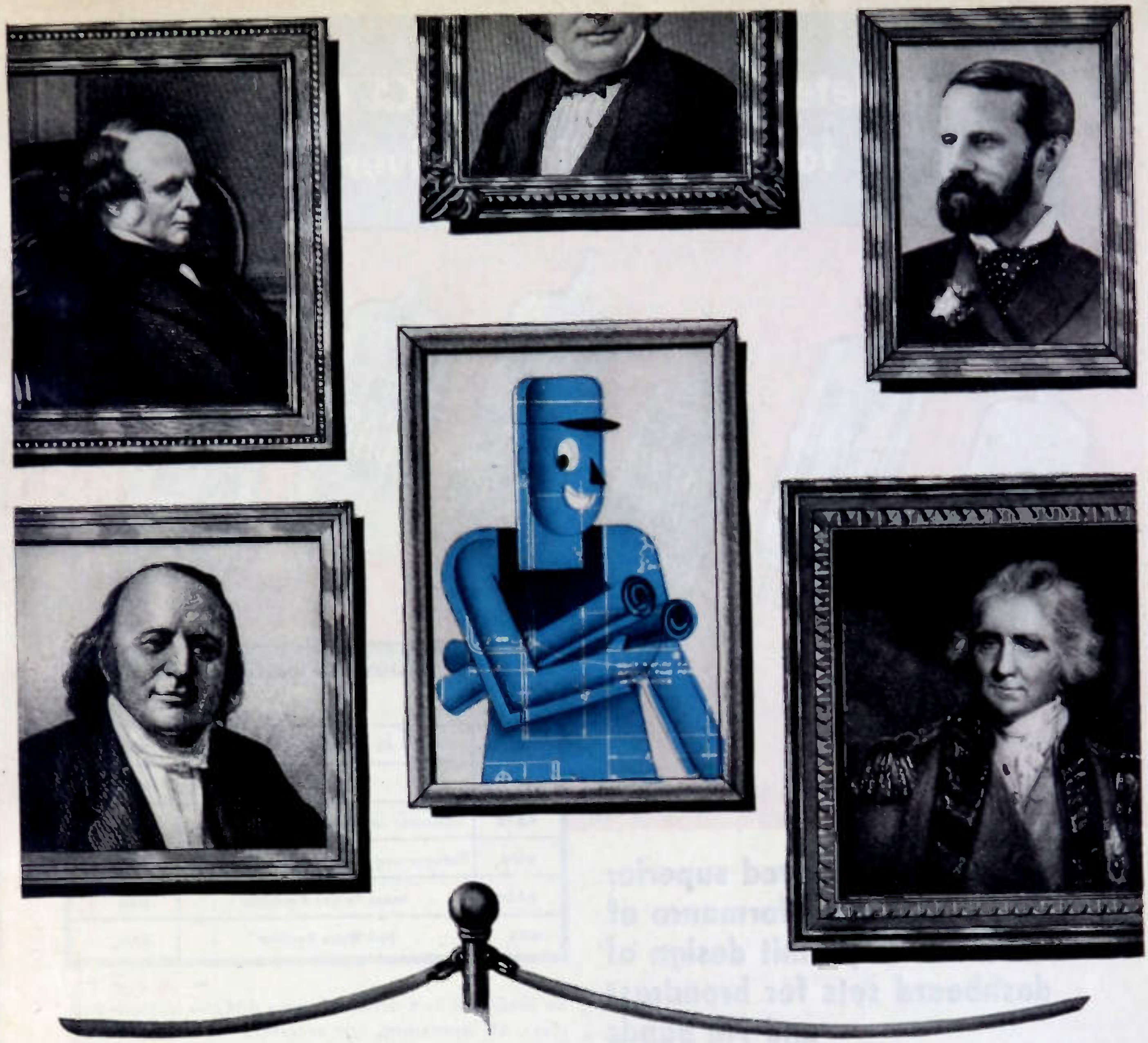

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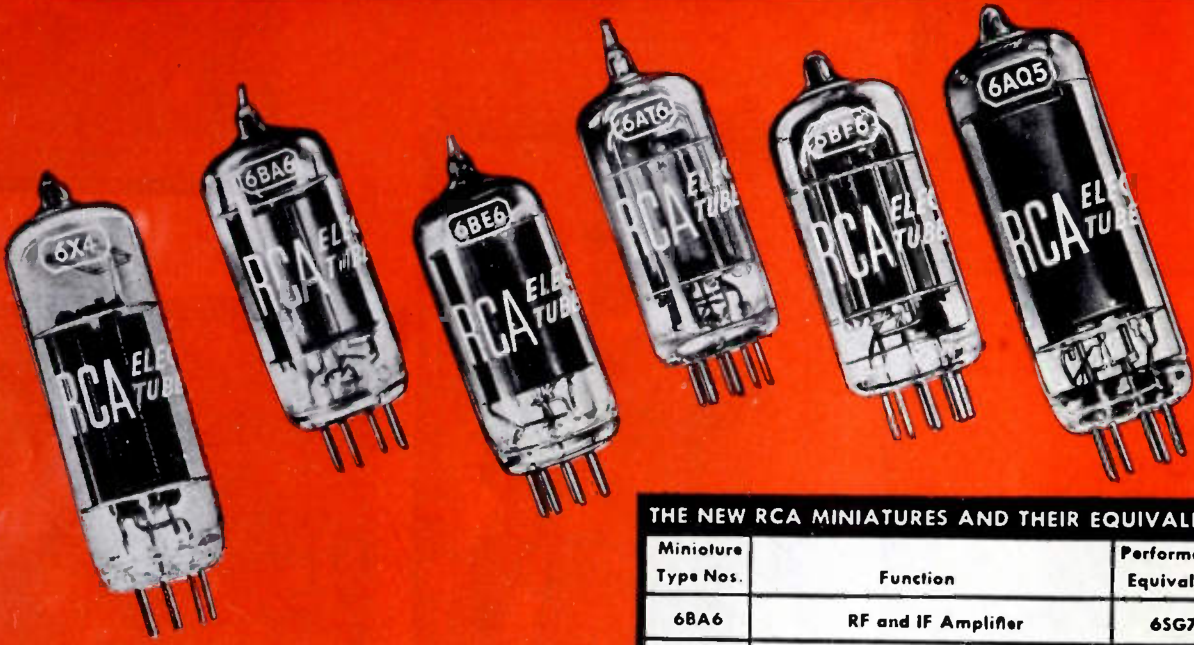

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There have of late been developed close relations between engineers and psychophysicists active in the acoustical and optical fields. Such associations have become increasingly important, not only in the design and utilization of military devices, but also in the intelligent adaptation of peacetime equipment to the needs and capabilities of its users. Readers of the PROCEEDINGS will accordingly find much food for constructive thought in the following guest editorial from an active and prominent worker at Harvard University, in the psychoacoustic field. This editorial is based on an article which appeared in *The American Scientist*, the courtesy of which publication in making this material available is here acknowledged.—*The Editor.*

The Human Factor in Machine Operation

(Based on "Machines Cannot Fight Alone")

S. S. STEVENS

The human side of engineering, the art of gearing machines to the minds and muscles of men, has become an indispensable part of the profession. Psychophysics, which, broadly defined, is the science of the senses in action, has been put to use (sometimes unwittingly, perhaps) by all those whose business it is to adapt machines to men and to get the best out of men by way of technical performance in an age of technological profusion. Whatever the task or the skill required, whatever the human sense employed, there are rules governing the actions and discriminations of men which make one way of doing things better than another.

Stemming from the work of G. T. Fechner, physicist and philosopher, who in 1860 gave us a treatise on the relation between sensation and the stimulus that causes it, the development of this science is the human story of our machine-age war. Implicit in the tale is the brief that the nature of man still determines the shape of his world, and of his wars, and that the science of man and his capacities must run hand in hand with technology if simple effective harmony is to ensue.

This recent war was different from other wars in the peculiar respect that it was fought largely on margin—sensory margin—where the battle hangs on the power of the eyes or the ears to make a fine discrimination, to estimate a distance, to see or hear a signal which is just at the edge of human capacity. Radars do not see, radios do not hear, sonars do not detect, guns do not point without someone making a fine sensory judgment, and the paradox of it is that the faster the engineers and the inventors served up their "automatic" gadgets to eliminate the human factor, the tighter the squeeze became on the power of the operator—the man who must see, hear, judge, and act with that margin of superiority which gives his outfit the jump on the enemy.

For the human story of communication in World War II we must look to the everyday busines of "passing the word." As in many other quests for better harmony between the operators and their gadgets, the methods of psychophysics were turned to account in two different ways: men were trained in the best use of existing equipment, and new devices were developed. Nursed by the urgency of these problems, there grew up and flourished a special area of psychophysics, called psychoacoustics, complete with laboratories and field stations. The claim can be made that the net improvement in certain vital communications added up to anywhere from fifty to several hundred per cent.

It was easy to improve upon the gear we were using in 1941 because it was already obsolescent. The mounting noisiness of guns and machines was making even our simplest communication systems indifferently effective. Our interphones were caught with their vowels down, while at the same time the engineers were bringing up new airplanes, bigger, faster, more powerful—and noisier. "Say again" was a very common phrase in those early days.

Four years of fevered scrambling in testing and development brought the art of communicating more nearly abreast of its scientific possibilities, and we fought the last battles of the war with new earphones, new microphones, new helmets, new oxygen masks, engineered in the light of the all-important human factor.

As with interphones, so with almost every other fighting tool. Wherever men stared searching into the darkness or taxed their eyes at a radar screen, wherever they held their breath to hear the tell-tale echo from the enemy under the sea or twisted their knobs and dials to launch destruction at the foe, the psychophysicist went along in spirit, fretting about the eyes and ears and muscles of the men and scheming to devise a greater harmony between the soldier and his tools. In the final drive to victory there were few important weapons used that bore exact resemblance to the gear employed when we first were jarred from slumber by the rudeness of aggression.

It is a proud story, perhaps, but something of a scandal, too. It depends upon how you want to look at it. However it is judged, we are left at this moment face up against decision. Are we now going to send the psychophysicsts and all their scientific colleagues back to their universities and their peacetime desks and leave our armed forces to enter the next war with the droppings of the last one? Then are we going to rush the scientists back to the frantic task of designing new gadgets and fitting them to human capabilities when the enemy is already upon us? Or are we going to keep science and human engineering alive with encouragement and support, holding one eye on the aspiration to make life more livable while we inhabit our spot of earth, and the other on the business of ensuring that no one takes it from us?

Walter R. G. Baker

PRESIDENT-ELECT, 1947

Walter R. G. Baker (A'19–F'28) was born in Lockport, New York, in 1890. He graduated from Union College, Schenectady, New York, receiving the degree of bachelor of electrical engineering in 1916, and the degree of master of electrical engineering in 1918. In 1935, the honorary degree of doctor of science was conferred upon him by Union College.

Beginning his career in the laboratories of the General Electric Company, Dr. Baker, during 1917 and 1918, took a prominent part in the development of wartime radio, including developmental work and testing of transmitting and receiving apparatus for the government. With the establishment of a separate radio department, he was made designing engineer in charge of transmitters. In 1924, this responsibility was enlarged to include design of all radio products, and in 1926 he was given complete charge of radio development, design, and production. Under his supervision, the broadcast transmitters of WGY, KGO, and KOA were designed, and the South Schenectady developmental laboratory was planned and built. He was also responsible for an extensive investigation of short-wave propagation. Upon the formation of the RCA Victor Corporation in 1929, Dr. Baker went to Camden, New Jersey, to head the radio engineering activities. He was later placed in charge of production, and subsequently became vice-president and general manager. When General Electric resumed radio production activities in 1935, Dr. Baker returned to the company as managing engineer of radio receiver activities in Bridgeport, Connecticut. In 1939, he became manager of the radio and television work in both the Schenectady and Bridgeport plants, and in 1941 he was made vice-president of the newly formed electronics department.

Dr. Baker is a member of the Board of Directors of The Institute of Radio Engineers; director of the engineering department of the Radio Manufacturers Association; member of the Board of Trustees of Union College; member of the Board of Governors of the National Electrical Manufacturers Association; chairman of the Electronics Committee of the American Institute of Electrical Engineers; and a member of the Newcommen Society. Dr. Baker served as chairman of the National Television Systems Committee, and as the first chairman of the Radio Technical Planning Board.

The Maximum Range of a Radar Set[*]

KENNETH A. NORTON[†], FELLOW, I.R.E., AND ARTHUR C. OMBERG[‡], SENIOR MEMBER, I.R.E.

Summary—Formulas are derived which may be used to calculate the maximum range of a radar set. It is shown that the maximum range obtainable with a given radar set depends upon (1) the energy in the pulse, i.e., the peak power times the time duration of the pulse or the average power divided by the pulse-recurrence frequency; (2) the transmitting and receiving antenna gains; (3) the transmission line, antenna, and transmit-receive-box losses; (4) the "effective width" or time duration of the transmitted pulses; (5) the "effective bandwidth" of the receiver; (6) the radio frequency of the transmitted waves; (7) the recurrence frequency of the transmitted pulses; (8) the "noise figure" of the receiver; (9) the cosmic, atmospheric, and man-made noise picked up by the antenna; (10) the attenuation during passage of the radio waves through the atmosphere due to the absorption by the atmospheric gases and rain drops; (11) the "effective echoing area" of the target; (12) the directivity of the transmitting and receiving antennas in elevation and azimuth, and (13) the effect of the ground, which, in turn, is inextricably associated with the particular site used, the height of the target above the ground, and the polarization of the transmitted radiowaves.

I. INTRODUCTION

IT IS THE purpose of this paper to provide a complete discussion of all the factors involved in the determination of one of the important military characteristics of a radar set, namely, its maximum range. It will be shown that the factors determining the maximum range may be conveniently divided into four parts: (1) the free-space maximum range, (2) the effects of the ground on the maximum range, (3) the effective echoing areas of the targets, and (4), at microwave frequencies, the absorption in the atmosphere.

If we let R_{max} denote the maximum range of a radar set for a target of effective echoing area A_e, and R'_{max} the free-space maximum range for a target of ten square meters effective echoing area, then it will be shown that

$$R_{max} = R_{max}' g(\theta)(A_e/10)^{1/4} e^{-\alpha R_{max}} \tag{1}$$

where θ is the elevation angle of the target above the ground and $g(\theta)$ is a function describing the effect of the ground on the maximum range. The function $g(\theta)$ oscillates with increasing values of θ between minimum values near zero and maximum values not greater than four. For most sites the maximum value of $g(\theta)$ is two, although for sets properly situated on sloping sites $g(\theta)$ may reach a maximum value near four for some particular value of θ. The absorption coefficient of the atmosphere is denoted by α; with R_{max} expressed in meters, $\alpha = 0.115$ times the absorption coefficient expressed in decibels per meter. At the higher microwave frequencies, where α is not negligible in magnitude, (1) must be solved for R_{max} by graphical or other convenient methods.

It is shown that cosmic, atmospheric, and man-made noise have the effect of decreasing the free-space maximum range in the lower part of the very-high-frequency band by as much as 50 per cent for receivers with low noise figures. In order to make possible a statement of a single value of maximum range independent of the site on which the radar set is located and independent of external noise and atmospheric absorption, a range index RI is proposed. This range index is defined to be twice the value of the free-space maximum range R_{max}' in the absence of external noise and in the absence of atmospheric absorption.

II. THE FREE-SPACE MAXIMUM RANGE OF A RADAR SET

The maximum range of a radar set is determined when the target is at such a distance that the pulse energy after propagation to and from the target is so small that it can just be detected in the presence of the noise energy always present in the radar receiver. Thus, it is necessary to determine (1) the pulse energy leaving the radar transmitting antenna, (2) the attenuation of this energy in traveling to the target, (3) the fractional part of the energy reflected from the target back in the direction of the radar set, (4) the attenuation of this energy on the return trip, (5) the amount of this energy picked up by the radar receiving antenna, and finally, for comparison with this pulse energy available at the receiver terminals, (6) the external and receiver noise energy at the receiver terminals.

Before starting the derivation of the formula for R_{max}' it will be desirable to define what is meant by pulse energy, peak power, average power, effective pulse duration, and pulse-recurrence frequency. The energy of the pulse E_t expressed in joules is equal to the integral of the instantaneous power p over the time interval occupied by the pulse; thus

$$E_t = \int p\,dt \text{ joules (watt-seconds)} \tag{2}$$

where p is expressed in watts and t in seconds. If the transmitted pulse were exactly rectangular and of duration τ seconds, the instantaneous power would be equal to the peak power P_t and the energy in the pulse would be

$$E_t = \int_0^\tau P_t\,dt = P_t\tau \text{ joules.} \tag{3}$$

Actual transmitted pulses are never exactly rectangular and it is convenient to define an "effective" pulse

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[*] Decimal classification: R357. Original manuscript received by the Institute, March 11, 1946; revised manuscript received, June 28, 1946. Presented, Washington Section of the I.R.E., February 27, 1946. This paper was originally published in February, 1943, by the Office of the Chief Signal Officer in the War Department as Operational Research Group Report ORG-P-9-1, which has recently been declassified. The data shown in Figs. 1 and 2 and Table II correspond to the characteristics of these radar sets as known at that time. In many cases subsequent improvements have been made. In publishing this paper, it has been brought up-to date by including additional material relative to radar echoes from the moon, atmospheric absorption, and cosmic noise received from the sun and the Milky Way.

[†] Central Radio Propagation Laboratory of the National Bureau of Standards, Washington 25, D. C.

[‡] Bendix Aviation Corporation, Baltimore, Maryland.

duration τ which may be used in (3) regardless of the pulse shape. This is accomplished simply by equating (2) and (3) and solving for τ; thus

$$\tau = \frac{1}{P_t}\int p\,dt \text{ seconds} \tag{4}$$

where the integral is to be taken over the total time interval occupied by the pulse. The average power $\overline{P}_t$ is equal to the pulse energy times the pulse-recurrence frequency F expressed in recurrences per second; thus

$$\overline{P}_t = E_tF = P_t\tau F \text{ watts.} \tag{5}$$

Using (3) and (5), we see that the pulse energy may be expressed either in terms of the peak power and the pulse duration or in terms of the average power and the pulse-recurrence frequency, as follows:

$$E_t = P_t\tau = \overline{P}_t/F \text{ joules.} \tag{6}$$

We will now proceed to calculate the energy reflected back from a target at a distance R, expressed in meters, from the radar transmitting antenna. Consider first that the transmitting antenna is an isotropic antenna; such an antenna radiates energy uniformly in all directions. Consider this isotropic antenna to be at a very large distance from the ground; by "very large" is meant a distance at least ten times the distance R. If E_t' represents the energy radiated during each pulse from the antenna, then, since it radiates energy uniformly in all directions, the pulse energy per square meter at the distance R is given by

$$S_0 = \frac{E_t'}{4\pi R^2} \text{ joules/meter}^2 \text{ (isotropic antenna).} \tag{7}$$

Note that $4\pi R^2$ is the area of the surface of a sphere, centered at the antenna and with radius R, through which all of the energy E_t' must pass. Only in the absence of atmospheric absorption, which is important only at the higher microwave frequencies, is (7) applicable; the effects of this absorption will be introduced in part IV. If G_t represents the gain of the actual transmitting radar antenna relative to that of the isotropic radiator above considered, and S_f represents the energy flow per square meter in the direction of the principal maximum of radiation from the actual radar antenna, then, by definition of gain, we have

$$S_f = G_tS_0 = \frac{G_tE_t'}{4\pi R^2} \text{ joules/meter}^2 \text{ (actual radar antenna).} \tag{8}$$

We will now define an area A_e expressed in square meters which will be called the "effective echoing area" of a reflecting obstacle or target. This effective echoing area may be defined as follows: if all the energy incident on this effective echoing area were scattered uniformly in all directions with the same polarization as that of the receiving antenna, the received echo would then be equal to the actual echo from the body. Thus, if the obstacle were in the direction of the principal maximum of radiation from the radar transmitting antenna and if S_r represents the energy per square meter reflected back from the obstacle to the radar receiving antenna at the distance R and with the same polarization as that of the receiving antenna, then by definition of A_e

$$S_r = \frac{A_eS_f}{4\pi R^2} \text{ joules/meter}^2. \tag{9}$$

Here again, atmospheric absorption has been neglected and will be introduced later. Comparing (7) with (9), it is seen that A_eS_f is the energy which would have to be scattered from an isotropic radiator to give the same echoed energy as the actual obstacle.[1] From the above definition, it is evident that the effective echoing area of an obstacle will, in general, be different for different polarizations of the incident wave from the transmitting antenna and for different receiving antenna polarizations and will have a still different value when the transmitting and receiving antennas have different polarizations. Now, if we substitute (8) in (9), we obtain

$$S_r = \frac{A_eG_tE_t'}{(4\pi R^2)^2} \text{ joules/meter}^2. \tag{10}$$

If G_r represents the gain of a matched receiving antenna relative to an isotropic receiving antenna (one which receives energy equally well from all directions) A_r the effective absorbing area corresponding to the direction of the principal maximum of reception of the matched receiving antenna, and λ the wavelength expressed in meters, then it may be shown[2] that

$$A_r = \frac{G_r\lambda^2}{4\pi} \text{ meters}^2. \tag{11}$$

If E_r is the maximum energy that can be dissipated in a receiver properly matched to the receiving antenna by means of a transmission line without loss, then

$$E_r = A_rS_r = \frac{G_rS_r\lambda^2}{4\pi} = \frac{G_rG_tE_t'A_e\lambda^2}{(4\pi)^3R^4} \text{ joules.} \tag{12}$$

The maximum range may now be obtained from (12) by letting R increase until E_r becomes equal to $E_{\min}$. $E_{\min}$ represents the minimum *available* pulse energy from the antenna (i.e., energy in the input impedance of the receiver when the antenna is matched to the receiver) required to make the pulse just visible in the presence of the noise for the particular receiver used and with the particular coupling between antenna and receiver actually employed. Thus we have

$$R_{\max}' = (4\pi)^{-3/4}(G_tG_rA_eE_t'/E_{\min})^{1/4}\lambda^{1/2} \text{ meters.} \tag{13}$$

It is convenient to write the minimum available pulse energy required for a barely visible pulse as follows:

$$E_{\min} = V\overline{NF}kT \text{ joules} \tag{14}$$

---

[1] This definition of "effective echoing area," is the same as that given in several Radiation Laboratory and Radio Research Laboratory reports. In the Telecommunications Research Establishment report M/28, "Elementary survey of scattering and echoing by elevated targets," H. G. Booker defines the "equivalent echoing area" of an obstacle as the area which must be assigned to the obstacle if it scattered power like a Hertzian dipole directed along the electric field of the incident wave. Consequently, due to this difference in definition, the "effective echoing areas" in this report are 1.5 times as large as the "equivalent echoing areas" in this British report.

[2] J. C. Slater, "Microwave Transmission," McGraw-Hill Book Company, New York and London, 1942, p. 260.

where

V = pulse visibility factor corresponding to the particular type of radar presentation used

$k = 1.37\times 10^{-23}$ = Boltzmann's constant

T = absolute temperature in degrees Kelvin

$kT = 4.11\times 10^{-21}$ joules at room temperature ($T = 300$ degrees)

$\overline{NF}$ is the noise figure of the receiver for the particular condition of match or mismatch to the antenna actually employed. For a perfect receiver completely over-coupled to the dummy antenna, $\overline{NF}=1$; for a perfect receiver matched to the dummy antenna, $NF=2$

$\overline{NF}kT$ is the fluctuation noise energy referred to the input circuit of the receiver.[3-6] The visibility factor V is thus the ratio of the minimum available pulse energy required for a barely visible pulse to the receiver fluctuation noise energy referred to the input circuit of the receiver.

The antenna gains in (13) are with reference to an isotropic radiator; since the gain of a half-wave antenna with reference to an isotropic radiator is equal to 1.641, the antenna gain G' relative to that of a half-wave antenna may be written $G' = G/1.641$. If we substitute (14) in (13); multiply (13) by 2 to take account of the effect of the ground, i.e., $g(\theta)$, at the elevation angle corresponding to the maximum of a lobe; introduce the radio frequency f expressed in megacycles, instead of the wavelength λ; replace kT by its value 4.11×10^{-21} joules corresponding to room temperature ($T = 300$ degrees); introduce a value of 10 square meters for A_e; and change the units of R from meters to miles, we have finally the following formula for the range index:

$$RI = 920(G'/f)^{1/2}(E_t'/V\overline{NF})^{1/4}\text{ miles.} \qquad (15)$$

In the above formula G' is the geometric mean of the transmitting and receiving antenna gains referred to that of a half-wave antenna; in the normal case where the same antenna is used both for transmission and reception G' is simply the gain of this antenna. In order to take into account the transmission-line, antenna, and transmit-receive-box loss we may introduce a factor L_t which will be defined as the ratio of the radiated energy to the transmitter pulse energy output; if we write E_t for the transmitter pulse energy output, then $E_t' = L_tE_t$. Furthermore, if we write L_r for the ratio of the available pulse energy at the receiver terminals to the available pulse energy at the receiving antenna terminals, then (15) becomes

$$RI = 920\,(G'/f)^{1/2}(E_tL_tL_r/V\overline{NF})^{1/4}\text{ miles.} \qquad (16)$$

The visibility factor V will depend upon (1) the particular type of presentation used, such as type-A or plan-position indicator (PPI), (2) the effective pulse width τ, (3) the particular shape of the pulse, (4) the effective bandwidth of the receiver B, (5) the pulse-recurrence frequency F, and (6) the rate of rotation and width of the beam of the radar antenna. An empirical formula for V may be obtained by an analysis of some experimental results obtained by A. V. Haeff[7] of the Naval Research Laboratory. Dr. Haeff conducted an extensive series of observations at 200 megacycles using rectangular pulses and type-A presentation. Observations were made to determine the minimum peak power of the pulse $P_{\min}$ required to make it just detectable in the presence of the receiver noise. Pulses were introduced from a pulse signal generator at random positions on the cathode-ray-tube screen and the pulse power increased until the pulse was just detected by the operator. Thousands of observations were made within the following ranges of pulse lengths, bandwidths, and recurrence frequencies:

$$2.5\text{ microseconds} < \tau < 30\text{ microseconds}$$
$$50\text{ kilocycles} < B < 1000\text{ kilocycles}$$
$$30\text{ cycles} < F < 1640\text{ cycles.}$$

An analysis of the above series of observations yields the following empirical formula for the minimum peak pulse power required for a just detectable pulse:

$$P_{\min} = \frac{1}{4}\left(1+\frac{1}{\tau B}\right)^2(1640/F)^{1/3}\overline{NF}kTB\text{ watts.} \qquad (17)$$

When we note that $E_{\min} = \tau P_{\min}$, we may solve for V from (14) and (17):

$$V = \frac{E_{\min}}{\overline{NF}kT} = \frac{\tau P_{\min}}{\overline{NF}kT} = \frac{\tau B}{4}\left(1+\frac{1}{\tau B}\right)^2(1640/F)^{1/3}. \qquad (18)$$

In the above B is the effective bandwidth of the receiver expressed in cycles per second.[3]

We see by (15) that the range increases as V decreases, so that the range will increase as the recurrence frequency F is increased; the reason more pulse energy is required for low pulse-recurrence frequencies is probably associated in some manner with the decay time of the light on the cathode-ray-tube screen between pulses. If we differentiate V with respect to either B or τ we find that V has a minimum, and thus an optimum, value when

$$B = \frac{1}{\tau}\text{ cycles (optimum bandwidth).} \qquad (19)$$

The following table shows V as a function of τB from 0.1 to 10, i.e., for bandwidths from 0.1 to 10 times the optimum value $(1/\tau)$, and for $F = 1640$ cycles.

Table I shows that over three times as much pulse energy is required when a bandwidth 1/10 or 10 times the optimum value is used; or we can also say that the bandwidth may be increased by a factor of 2.66 times the optimum value without increasing the required energy by more than one decibel; a 1-decibel difference in energy corresponds to a 6 per cent difference in the

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[3] H. T. Friis, "Noise figures of radio receivers," Proc. I.R.E. vol. 32, pp. 419–422; July, 1944.

[4] D. O. North, Discussion on "Noise Figures of Radio Receivers," H. T. Friis, Proc. I.R.E., vol. 33, pp. 125–127; February, 1945.

[5] D. O. North, "The absolute sensitivity of radio receivers," *RCA Rev.*, vol. 6, pp. 332–343; January, 1942.

[6] R. E. Burgess, "Noise in receiving aerial systems," *Proc. Phys. Soc.*, (London), vol. 53, pp. 292–304; May, 1941.

[7] A. V. Haeff, "Minimum detectable radar signal and its dependence upon parameters of radar systems," Proc., I.R.E., vol. 34, pp. 857–861; November, 1946.

free-space maximum range. The reason more pulse energy is required when the bandwidth is made too narrow is the fact that only a small portion of the pulse energy passes through the receiver under these conditions; on the other hand, when the bandwidth is made much greater than $(1/\tau)$ the output pulse energy does not continue to increase with increasing bandwidth whereas the output noise energy increases continuously with increasing bandwidth. It should be noted that, when $F=1640$ cycles and $B=1/\tau$ cycles, V is equal to one and the pulse energy required for a barely visible pulse is just equal to the receiver fluctuation noise energy under these optimum bandwidth conditions, so that the pulse-plus-noise-energy to noise-energy ratio is equal to 2. When we consider the experimental method used in arriving at the above formula for the visibility factor, we see that it is applicable only to the case where the radar antenna is searchlighting, i.e., pointed in the direction of the target. In order to generalize the expression for V so that it may be used for other types of presentation, for rotating radar antennas, and for other than rectangular pulses, we may write

TABLE I

Visibility Factor for Square Pulses with Type-A Presentation and Recurrence Frequency of 1640 Cycles.

τB	V	decibels
0.1	3.025	4.81
0.2	1.8	2.55
0.5	1.125	0.51
1	1	0
2	1.125	0.51
5	1.8	2.55
10	3.025	4.81

$$V = \frac{k_1\tau B}{4}\left(1+\frac{k_2}{\tau B}\right)^2(1640/F)^{1/3}. \qquad (20)$$

In the above k_1 is a factor which will differ with the type of presentation, rate of rotation, and width of the beam of the radar antenna, and possibly with the type of detection, e.g., linear, square-law, etc., being equal to unity for type-A presentation and searchlighting antennas; while k_2 is a factor which will differ with the shape of the pulse, being equal to unity for rectangular pulses. When (20) is differentiated with respect to τ or B we find that the optimum bandwidth for arbitrary pulse shapes is given by

$$B = k_2/\tau \text{ cycles per second.} \qquad (21)$$

If we substitute (20) into (16) we obtain the following general formula for the range index, i.e., the maximum range in the maximum of one of the ground-reflection lobes for a large fighter or small bomber aircraft in the absence of external noise or atmospheric absorption:

$$RI = 494.9(E_tL_tL_rG_t'G_r'/f^2\overline{NF})^{1/4}$$
$$\cdot\left[\frac{k_1\tau B}{4}\left(1+\frac{k_2}{\tau B}\right)^2\right]^{-1/4} F^{1/12} \text{ miles} \qquad (22)$$

where E_t is the transmitter output pulse energy expressed in joules (6); $\overline{NF}$ is the noise figure of the receiver corresponding to the particular condition of coupling employed between it and the radar receiving antenna; G_t' and G_r' are the gains of the transmitting and receiving antennas, respectively, relative to that of a half-wave antenna; L_t and L_r are factors less than one which allow for the transmission-line, antenna and transmit-receive-box loss for transmission and reception, respectively; τ is the effective pulse duration expressed in seconds as defined in (4); B is the effective bandwidth of the receiver expressed in cycles per second[3]; F is the pulse-recurrence frequency expressed in recurrences per second; f is the radio frequency expressed in megacycles; k_1 is a factor dependent on the type of presentation, rate of rotation, and width of the beam of the radar antenna, and the type of detection; and k_2 is a factor dependent on the shape of the pulses, being equal to unity for rectangular pulses.

Equation (22) contains all of the factors that are under the control of the designer of a radar set and which affect the maximum range. The gains of the transmitting and receiving antennas clearly should be made as large as possible; however, high gain means narrow beams, either in azimuth or elevation, and this may result in a loss of coverage or in operational complications. Furthermore, high gain means large antennas which may not be sufficiently portable or may be difficult to rotate. Thus we see that there are definite limits to the amount of gain obtainable in practice. Next, the pulse energy E_t should be made as large as possible; for a given value of pulse-recurrence frequency, which is usually fixed by the range within which observations are to be made, E_t is directly proportional to the average transmitter power $\overline{P}_t$, so that this provides an equally good measure of radar-set performance in most cases. The maximum average power which can be used is limited either by the requirement of portability or by the capabilities of transmitting tubes. Next, the noise figure of the receiver should be made as small as possible. This is a particularly important requirement, since a set with a low noise figure requires no more tubes, weighs no more, and occupies no more space; yet it makes possible an increase in the range of the radar set. The pulse width τ and the bandwidth B occur in the formula only as the product τB, and the range is a maximum when $\tau B=1$; once the pulse width has been settled upon from operational considerations, the bandwidth should be made equal to $(1/\tau)$ or at least no greater than twice this value for optimum performance. Adjustment of the bandwidth to this optimum value is also particularly important, since this adjustment is comparatively inexpensive and yet increases the maximum range as much as an equivalent increase in transmitter power. The pulse-recurrence frequency should be made as large as possible consistent with the maximum range expected of the set. When all of the other factors in the above formula are fixed, the maximum range of the set decreases with increasing frequency; thus, the frequency of an early-warning set should be as low as possible consistent with the possibility of properly designing an antenna with sufficient gain and directivity and which is not too large to be portable or rotatable.

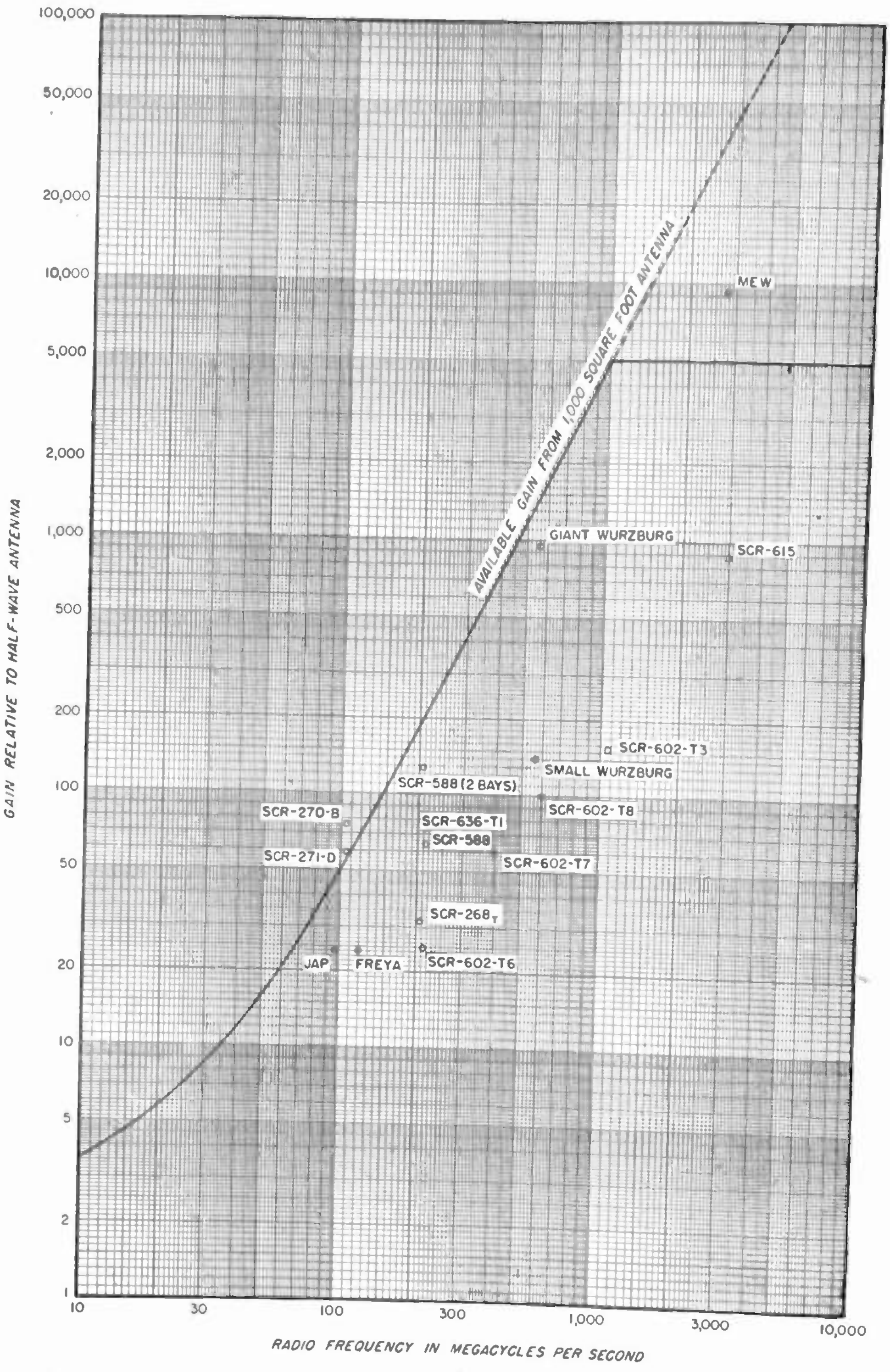

Fig. 1—Antenna gain assumed in determining early-warning radar optimum frequency; shown for comparison are the antenna gains used on several radar sets.

In view of the fact that the antenna gain, noise figure of the receiver, and available pulse energy from vacuum tubes all vary with frequency, the determination of the optimum frequency for a radar set designed to perform a given function probably can best be determined only by calculating the range for various frequencies by means of (22) and then picking the frequency providing the greatest coverage. As an example of the above suggested procedure, consider the problem of determining the optimum frequency for a fixed early-warning radar set for which size and weight are of little importance. At low frequencies, the maximum antenna gain which can be obtained will be limited by the size of an antenna which can be conveniently mounted and rotated; we will assume that the maximum area for such an antenna will be about the same as that of the SCR-270, i.e., about 1000 square feet. If we reduce this value of area by a factor of 2/3 to allow for a distribution of current designed to cut down the size of the side lobes of radiation, then we obtain, from (11), $G'=0.005\ f^2$ where f is the frequency expressed in megacycles. At very high frequencies, on the other hand, the maximum gain which can be used will be limited by the requirements that (1) the beam must not be so narrow in elevation that it will not illuminate all targets up to altitudes of at least 50,000 feet, and (2) that the beam must not be so narrow in azimuth that it cannot cover its sector of search in a reasonably short period of time; experience indicates that antenna gains greater than about 5000 cannot be usefully used. From the above we see that the maximum available or useful values of G' are given by $G'=0.005f^2$ for 100 megacycles $<f<1000$ megacycles and $G'=5000$ for $f>1000$ megacycles; for frequencies greater than 1000 megacycles, the required area will be less than 1000 square feet, being given by the formula $A=10^9/f^2$ square feet. The antenna gain obtainable from a 1000-square foot antenna is slightly greater than $0.005f^2$ for frequencies less than 100 megacycles. The antenna gain for this ideal early-warning radar set is shown in Fig. 1 as a function of frequency throughout the range from 10 to 10,000 megacycles. In order to arrive at reasonable values for the noise figure of the receiver, we will assume that it contains a GL 446 (lighthouse radio-frequency amplifier) at frequencies less than about 1500 megacycles and will use the data given in Fig. 7 of Radiation Laboratory Report 43-14, reproduced in Fig. 2. At frequencies above 1500 megacycles data are given for crystals as well as for the lighthouse tube.[8] The available information on the available pulse energy from transmitters as a function of frequency is rather meager. However, it is known that the SCR-270 has a pulse energy of about 1.5 joules while that of the SCR-615 has about 0.5 joules; in view of the lack of any better data, we will assume that the available pulse energy is given by the equation $E_t=7/f^{1/3}$ joules; this is also shown in Fig. 2.[9] We will use a bandwidth B equal to twice the optimum value $(1/\tau)$ so as to allow for some spreading of the pulse energy due to unintentional frequency modulation and to permit the receiver tuning to be somewhat less critical; we will use a pulse-recurrence frequency $F=200$, and thus $V=2.26$. We will assume that the transmission-line and transmit-receive-box loss is independent of frequency and will take $L_t=0.9$ and $L_r=0.8$. Using all the above data in (22), we obtain the resulting curve for range index shown in Fig. 2. For frequencies above 1000 megacycles the solid range-index curve corresponds to a constant gain of 5000, while the dotted curve corresponds to the gain obtainable with a 1000-square-foot antenna as shown by the dotted gain curve. For frequencies above 1500 megacycles, a crystal detector was used, with the noise figure as shown in Fig. 2, to calculate the range-index curves. The figure shows clearly that there is an optimum frequency for an early-warning set in the neighborhood of 1000 megacycles; it also appears from the data shown that the use of this optimum frequency will more than double the range index over values obtainable with present radar sets, even after these sets have been improved so that they will operate with maximum efficiency. Also shown in Fig. 1 and 2 for comparison with the optimum curves are the values of antenna gain, transmitter pulse energy, receiver noise figure, and the visibility factor for several American radar sets and for the Japanese early-warning radar set captured in the South Pacific area. It will be noted that the calculated range of the microwave early-warning radar (MEW) is greater than the optimum value shown by the curve. The reason for this is that the antenna gain for this set is more than 8000 and it is believed that this gain is too large to be operationally advantageous; this point can be settled conclusively only by means of adequate operational tests.

III. THE EFFECTS OF EXTERNAL NOISE ON THE FREE-SPACE MAXIMUM RANGE

All of the factors in (22) may be determined by means of laboratory measurements and are independent of the particular conditions which may be encountered at a particular site; thus the range index as determined by (22) and shown graphically as a function of frequency in Fig. 2 is an appropriate laboratory measure of the expected performance of a radar set. When the radar set is installed at a particular site, in addition to the effects of the ground and the effects of atmospheric absorption and bending, which will be discussed in the following part of this report, the cosmic, atmospheric, and man-made noise picked up by the antenna will have the effect of increasing the noise energy at the first circuit of the receiver and will thus reduce the range somewhat with respect to its value as given by (22). In order to obtain a formula for the free-space maximum range $R'_{\max}$ which may be substituted in (1) to give the actual maximum range to be expected in the field, the effects of this external noise will now be introduced.

When measurements are made of the noise figure of the receiver at a particular radar site and the first circuit of the receiver is coupled to the actual radar receiving antenna rather than to a lumped impedance of the same resistance and reactance as that of the antenna, then the measured value of effective noise figure may be either larger or smaller due to the external noise picked up by the antenna. It should be noted that a radiation resistance is not a real resistance and thus introduces no noise into the receiver except to the extent that it absorbs noise radiation from its surroundings. If we write $N_a \equiv \overline{EN}kT(T=300$ degrees) for the available noise energy picked up by the antenna, then

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[8] Recent advances in crystal techniques now permit noise figures of 9 to 10 decibels from 1000 to 10,000 megacycles.

[9] New magnetrons permit pulse energies of 2 joules at 3000 megacycles and 1 joule at 10,000 megacycles (4-J-50 magnetrons).

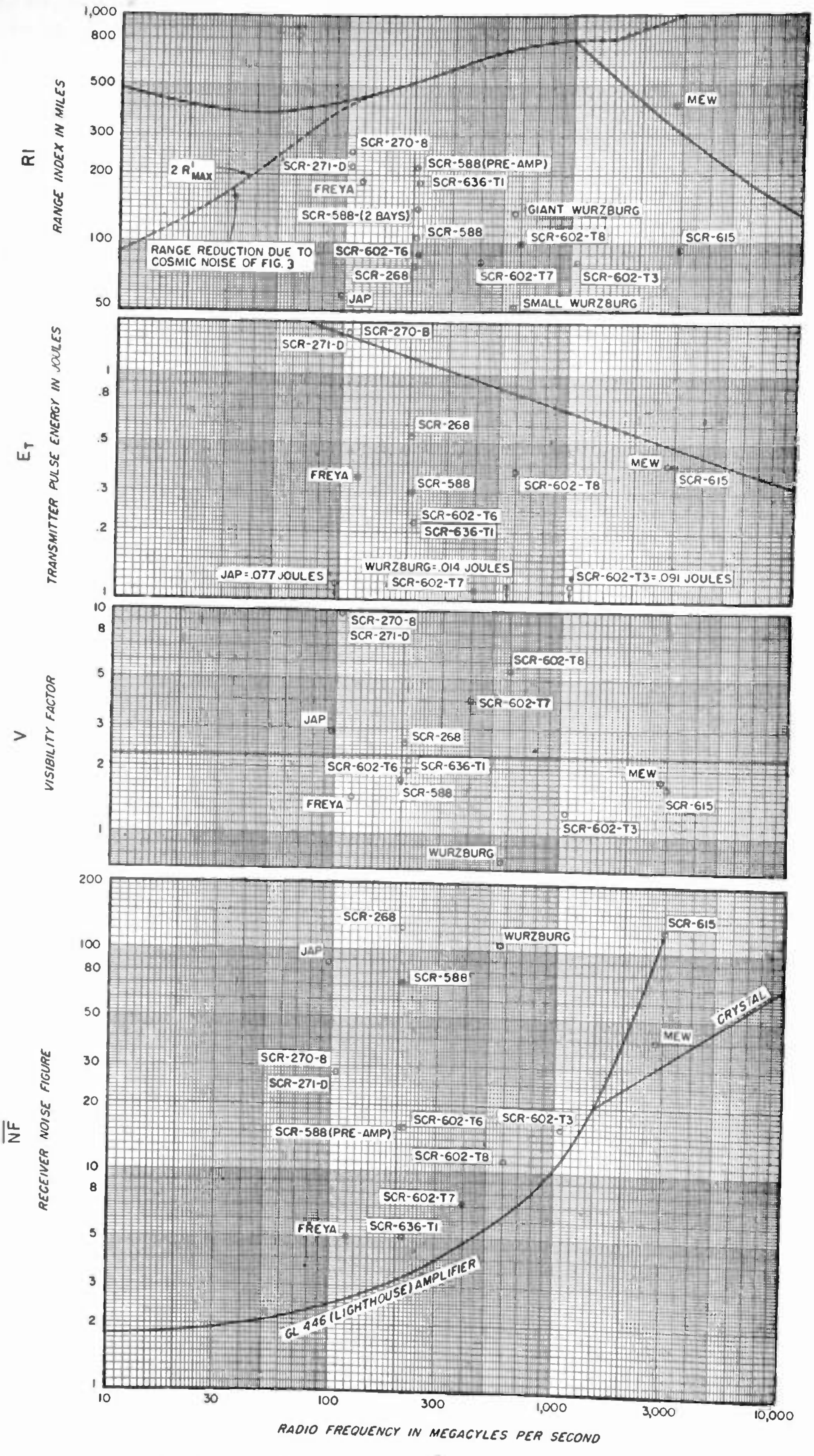

Fig. 2—Receiver noise figure, radar visibility factor, transmitter pulse energy, and resulting range index for an assumed ideal early-warning radar set compared with these factors as measured for several existing radar sets.

the dimensionless external noise factor $\overline{EN}$ as defined by this relation provides a convenient measure of the external noise energy. Thus, using $\overline{EN}$ as a measure of the external noise energy, the effective noise figure of the receiver-antenna combination as measured with the actual antenna at the particular radar site becomes equal to

$$\overline{NF'} = \left(\frac{\overline{NF}}{L_r} + [\overline{EN} - 1]\right) \tag{23}$$

and the corresponding value of free-space maximum range thus may be written

$$R_{\max}' = 458.6[E_tL_tL_rG_t'G_r'/f^2V(\overline{NF} + L_r[\overline{EN} - 1])]^{1/4} \text{ miles.} \tag{24}$$

The factor L_r enters into the expression as an allowance for the reduction in the signal (and external noise) energy caused by the transmission-line, antenna, and transmit-receive-box losses. It should be noted that the effective noise figure $\overline{NF'}$ of the receiver-antenna combination will be equal to the receiver noise figure $\overline{NF}$ only when there are no transmission-line losses and the antenna happens to pick up the same amount of noise energy as would be available from a resistance at room temperature, i.e., kT, so that $\overline{EN}=1$. By definition of $\overline{NF}$ and $\overline{EN}$, T is taken to be 300 degrees. Instead of introducing the dimensionless external noise factor $\overline{EN}$, some investigators[5,22] have simply introduced the concept of an effective absolute temperature T_a of the antenna resistance. In that notation, the available external noise energy picked up by the antenna would simply be equal to kT_a, and when this is equated to $\overline{EN}kT$ we see that

$$\overline{EN} \equiv T_a/T = T_a/300. \tag{25}$$

The concept of an effective absolute antenna temperature is quite useful, as will be seen in the sequel, for the description of the nature of some external noise sources. However, the dimensionless factor $\overline{EN}$, since it implies no special origin for the external noise, would appear to be more desirable for describing such diverse external noises as those arising in thunderstorms, man-made noise such as auto ignition, cosmic noise, and the noise associated with sun spots, as well as the temperature noise due to the fluctuation-noise energy in the matter surrounding the antenna.

With the exception of temperature noise, which is frequency independent, all of the external noise sources decrease very rapidly in intensity with increasing radio frequency and become negligibly important above 300 megacycles. Thus, in free space (for example, in an aircraft flying at high altitudes) $\overline{EN}$ for thunderstorm noise will be inversely proportional to the fourth power of the radio frequency. For antennas nearer the ground the $\overline{EN}$ for thunderstorms will differ somewhat from the fourth-power law due to the frequency dependence of the propagation of the thunderstorm noise over the surface of the earth; this propagation factor prevents all except very local thunderstorms from being heard at very-high frequencies and higher. Cosmic noise was first identified and measured by Jansky[10–13]. From the data presented in Jansky's 1937 paper, we find that the cosmic noise at 18 megacycles corresponded to values of $\overline{EN}$ varying between about 11 and 110 as measured on a highly directional rhombic antenna, and varying between about 72 and 350 as measured on a horizontal half-wave doublet. Other measurements of cosmic noise have been made by Reber,[14–17] Fränz,[18] Sander,[19] Hey, Parsons and Phillips,[20] and Moxon,[21] and an analysis of some of these results has been made by Scott[22] which indicates that cosmic noise, when averaged over the sky, or alternatively received on an isotropic receiving antenna, corresponds to an external noise factor

$$\overline{EN} \cong 1.8(10^6)/f_{Mc}{}^3 \tag{26}$$

within the frequency range from 18 to 160 megacycles; some of the later measurements seem to indicate that the rate of decrease of cosmic noise with frequency is less than that estimated by Scott and that it varies more nearly as the inverse square rather than the inverse cube of the radio frequency. Scott's estimate is shown on Fig. 3 and we see that $\overline{EN}<1$ for $f>120$ megacycles and cosmic noise will thus be practically of very little importance above 120 megacycles but, since it is always present, will form the lower limit to the sensitivity of radio reception in the lower part of the very-high-frequency band. The sources of this cosmic noise are not uniformly distributed over the sky but tend to be concentrated in several regions on the celestial sphere, the principal source being in the region Scorpio-Sagittarius near the center of the galaxy. Consequently, when received on a directional antenna, the maximum values of the noise may be expected to be several times greater than the average values shown in Fig. 3 and may be expected to vary in a characteristic manner from hour to hour and from day to day.

At frequencies above 300 megacycles temperature noise is usually the principal source of noise external to

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[10] K. G. Jansky, "Directional studies of atmospherics at high frequencies," Proc. I.R.E., vol. 20, p. 1920; December, 1932.

[11] K. G. Jansky, "Electrical disturbances apparently of extraterrestrial origin," Proc. I.R.E., vol. 21, p. 1387; October, 1933.

[12] K. G. Jansky, "A note on the source of interstellar interference," Proc. I.R.E., vol. 23, pp. 1158–1163; October, 1935.

[13] K. G. Jansky, "Minimum noise levels obtained on short-wave radio receiving systems," Proc. I.R.E., vol. 25, pp. 1517–1530; December, 1937.

[14] G. Reber, "Cosmic static," Proc. I.R.E., vol. 28, pp. 68–70; February, 1940.

[15] G. Reber, "Cosmic static," Proc. I.R.E., vol. 30, pp. 367–378; August, 1942.

[16] G. Reber, "Cosmic static," *Astrophys. Jour.*, vol. 91, p. 621; 1940.

[17] G. Reber, "Cosmic static," *Astrophys. Jour.*, vol. 100, pp. 279–287; 1944.

[18] K. Fränz, *Elek. Nach. Tech.*, vol. 16, p. 92; 1942. *Hochfrequenz. und Electroakustic*, vol. 59; 1942.

[19] K. F. Sander, Radar Research Development Establishment Report No. 285, 1945.

[20] J. S. Hey, S. J. Parsons, and J. W. Phillips, "An investigation of cosmic noise at 64 mc/s," Army Operational Research Group Report, 1945.

[21] L. A. Moxon, Admiralty Signal Establishment Extension Report, reference XRC 3/45/9, dated 26/11/1945.

[22] J. M. C. Scott, "The intensity of cosmic noise, a survey of the data available," Radar Research Development Establishment Report No. 286, August, 1945.

the receiver. This is the only type of external noise for which an absolute temperature T_a, as given by (25), is an appropriate measure. A directional receiving antenna will absorb different amounts of temperature noise energy as it is pointed in different directions. This directional dependence of T_a, and consequently of that part of $\overline{EN}$ due to temperature noise, may be expressed as follows:

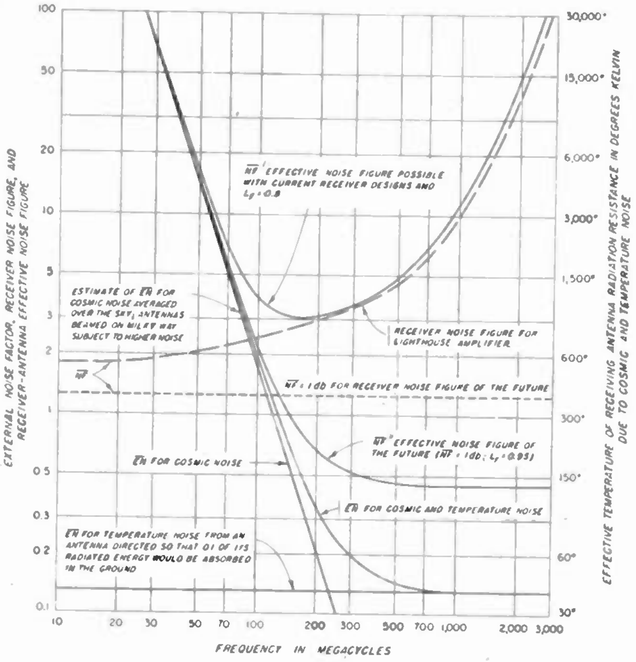

Fig. 3—Effective noise figures for receivers, including effects of cosmic and antenna temperature noise.

$$T_a = 300\,\overline{EN} = \frac{1}{4\pi}\iint T(\theta,\phi)\cdot G(\theta,\phi)d\omega. \qquad (27)$$

The integration is performed over a surface surrounding the antenna. $T(\theta, \phi)$ is an effective absolute temperature of the material in space as properly weighted and averaged with respect to distance along the beam in an elementary solid angle $d\omega$ centered about the direction θ, ϕ. $G(\theta, \phi)$ is the gain of the antenna in the direction θ, ϕ relative to that of an isotropic radiator. The proper method of determining the effective value of $T(\theta, \phi)$ may be seen most readily from the reciprocal problem in which energy radiated from the antenna is absorbed as it is propagated from the antenna out to a distance such that it is completely absorbed and remembering that good absorbers are correspondingly good radiators. The following artificial example will serve to clarify the problem. If one-third of the total energy radiated in the elementary solid angle $d\omega$ centered on the direction θ, ϕ were absorbed in a gas of uniform absolute temperature of 300 degrees extending from 0 to 1000 miles from the antenna, another one-third absorbed in a gas of uniform absolute temperature 30 degrees extending from 1000 to 1250 miles, and the final one-third of the energy absorbed in a black body at the distance 1250 miles with a surface temperature of 600 degrees then $T(\theta, \phi)$ for that direction would be equal to $1/3(300+30+600)=310$ degrees absolute temperature. The definition of T_a as given by (27) and as explained above arises from the principle of detailed balancing in statistical mechanics, according to which in thermal equilibrium each infinitesimal process must be balanced by its inverse process.[23,24] Thus, since T_a has been confined by definition to refer only to temperature noise arising from fluctuations in the matter surrounding the antenna, and since the radiation resistance of the antenna must be in thermal equilibrium with its surroundings, it is only necessary to apply the principle of detailed balancing of absorption and radiation processes to the matter in each part of the space surrounding the antenna and to assume that $G(\theta, \phi)$ is the same for transmission and reception in order to derive (27). It should be noted that, when $T(\theta, \phi)$ equals a constant value T_c in all directions, then T_a will simply be equal to T_c since the constant T_c may then be taken from under the integral signs and the integral is, by definition of $G(\theta, \phi)$, simply equal to 4π. On the other hand, for a high-gain antenna, if $T(\theta, \phi)$ has a very large value T_e over the effective beam of the antenna and a very small value in other directions, then T_a will again be nearly equal to T_e since the contributions to the integral for directions θ and ϕ far removed from the maximum of the antenna will be negligible. Measurements of the effective noise temperatures of antennas in the ultra-high-frequency band beamed on the open sky are of the order of 10 degrees absolute, corresponding to the very low value of $\overline{EN}=0.033$. When these antenna beams are directed horizontally along the ground, a small part, say one-tenth of the energy which could be transmitted from such an antenna would be absorbed in the ground, and, since the earth is at a temperature approximating 300 degrees, the effective noise temperature of such an antenna used for reception and directed horizontally along the ground would be equal to $T_a = [(1/10)300+(9/10)10]=39$ degrees corresponding to a value of $\overline{EN}=0.13$ (see Fig. 3). In the future, when receivers with very low noise figures become available in the ultra-high-frequency band, it may turn out to be desirable to discriminate against the ground-reflected waves in order to reduce the received noise; discrimination against the ground-reflected wave by means of highly directional receiving antennas has already proved on microwave relay circuits to be a valuable method for reducing, and in some cases practically eliminating, the adverse effects of within-line-of-sight fading due to the interference between the direct and ground-reflected components.[25,26]

At microwave frequencies the large antenna gains which are available have made possible the direct

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[23] J. C. Slater, "Microwave Transmission," McGraw-Hill Book Co., Inc., New York and London, p. 256, 1942.

[24] J. C. Slater, "Introduction to Chemical Physics," McGraw-Hill Book Co., Inc., New York and London, p. 91, 1939.

[25] R. Bateman and T. J. Carroll, "Dish Tilt Experiment—Propagation Report No. 1," in connection with "Comparative Tests of Radio Relay Equipment," Office of the Chief Signal Officer, The Pentagon, Washington, D. C.

[26] Ross Bateman, "Elimination of interference-type fading at microwave frequencies with spaced antennas," Proc. I.R.E., vol. 34, pp. 662–663; September, 1946.

observation of temperature noise from the sun by Southworth.[27] In the region 1 to 10 centimeters, Southworth has been able to show that the radiation received from the sun is quantitatively only slightly greater than that to be expected from a black body the size of the sun with a surface absolute temperature of 6000 degrees. Since Southworth used an aperture angle on his receiving antenna at his highest frequency comparable to that of the sun, it follows from (5) that $\overline{EN}$ should be approximately equal to $6000/300=20$ when the antenna is pointed directly at the sun. At the two lower frequencies the receiving antenna apertures were larger and this resulted in smaller values of $\overline{EN}$; in other words, the antenna "sees" some of the space adjacent to the sun, which has a very much lower effective temperature.

Recently several observers[28,29] have reported the reception, at very-high frequencies, of unusually intense noise from the sun which appears to be associated with sun spots since it occurs only when large groups of spots are visible on the sun. As an indication of the possible practical importance of this solar noise a description will be given of some measurements of it made at the Federal Communications Commission monitoring station at Laurel, Maryland, from January 31 to February 12, 1946, during which time several large groups of sun spots were crossing the face of the sun. The solar noise was recorded at 44.9 megacycles on a half-wave horizontal dipole 30 feet above the ground. The sensitivity of the receiver, which had a noise bandwidth of approximately 120 kilocycles, corresponded to a noise figure of about 10, and the Esterline-Angus recorder used for indicating the amplified levels of the incoming signals at the intermediate frequency of the receiver would respond to signal-generator voltages less than 0.8 microvolt. When coupled to the antenna the usual noise voltage V_n appearing across the antenna terminals was approximately equal to 1.5 microvolts, but this increased during the daytime on each day of the above period, reaching a maximum equal to 15 microvolts at 1300 Eastern Standard Time on February 6, which lasted a few minutes. The usual level of this solar noise as recorded during this period corresponded to something less than 4 microvolts. To convert these data into effective antenna temperatures we may equate the available noise power from the half-wave antenna ($V_n^2/4R$ where R is the radiation resistance of the antenna, say 73 ohms) to the available noise power kT_aB in the 120-kilocycle receiver band B to be expected from the antenna resistance at the temperature T_a. Thus, with V_n expressed in microvolts,

$$T_a = V_n^2 \cdot 10^{-12}/292kB = 2080V_n^2. \tag{28}$$

We see by the above that the noise temperature of the half-wave dipole radiation resistance due to this solar noise reached a peak value of more than 450,000 degrees. Furthermore, the equivalent temperature of a black body with the projected surface area of the sun must be much greater than T_a in order to produce this much noise power. This may be seen from (27), which may be written in this case

$$T_a = \frac{\Delta\omega}{4\pi} T_s(\theta, \phi)G(\theta, \phi) \tag{29}$$

where $G(\theta, \phi)$ is the gain of the dipole antenna in the direction of the sun and T_s is the surface temperature of the sun. Over the small aperture angle $\Delta\omega$ occupied by the surface of the sun, $G(\theta, \phi)$ will be constant and, since T_s is practically equal to zero in other directions, (27) becomes equal to (29). Since the aperture of the sun $\Delta\omega$ is equal to 6×10^{-5} steradians and $G(\theta, \phi)$ will never be larger than 1.64, we see by (29) that T_s would have to be at least 4×10^{10} degrees in order to account for the observed solar noise. We are thus forced to the conclusion that this solar noise observed at very-high frequencies and associated with sun spots is of an entirely different character than that observed at microwave frequencies. Observations[30] on various frequencies in the very-high-frequency band indicate that the intensity of this sun-spot noise decreases rapidly with increasing frequency. This suggests that the source of this sun-spot noise is similar to that of the cosmic noise from interstellar space.

Man-made noise is also known to decrease rapidly with increasing frequency. Thus, for example, George[31] has measured the field intensities of auto ignition noise, which is one of the worst offenders in the very-high-frequency band, and these measurements indicate that $\overline{EN}$ for such noise decreases with frequency at a rate somewhere between inverse square and inverse cube of the frequency.

By an appropriate choice of receiving location, man-made noise may often be largely avoided; thunderstorm and sun-spot noise are of importance for only a small percentage of the time. Thus, we are left with cosmic and temperature noise as the most important sources of external noise because they are always present and thus set the ultimate limit to the sensitivity of radio reception at very-high frequencies and above. In Fig. 3, the dashed curve corresponds to the measured noise figure of a good receiver employing a GL-446 (lighthouse) amplifier and represents very nearly the *present* optimum performance of radio receivers of modern design. If we assume, as before, that the transmission-line loss corresponds to $L_r=0.8$, then this value of $\overline{NF}$ may be combined by means of (23) with the value of $\overline{EN}$ shown for cosmic and temperature noise to give the effective value of noise figure labelled $\overline{NF}'$ on Fig. 3. This curve is believed to be a reasonably accurate measure of the maximum possible sensitivity to be expected from

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[27] G. C. Southworth, *Jour. Frank. Inst.*, vol. 239, p. 285; 1945.
[28] E. V. Appleton, *Nature*, vol. 156, p. 534; November, 1945.
[29] J. S. Hey, *Nature* (London), vol. 157, p. 48; January, 1946.

[30] J. L. Pawsey, R. Payne-Scott, and L. L. McCready, "Radio frequency energy from the sun," *Nature*, vol. 157, pp. 158–159; February 9, 1946.

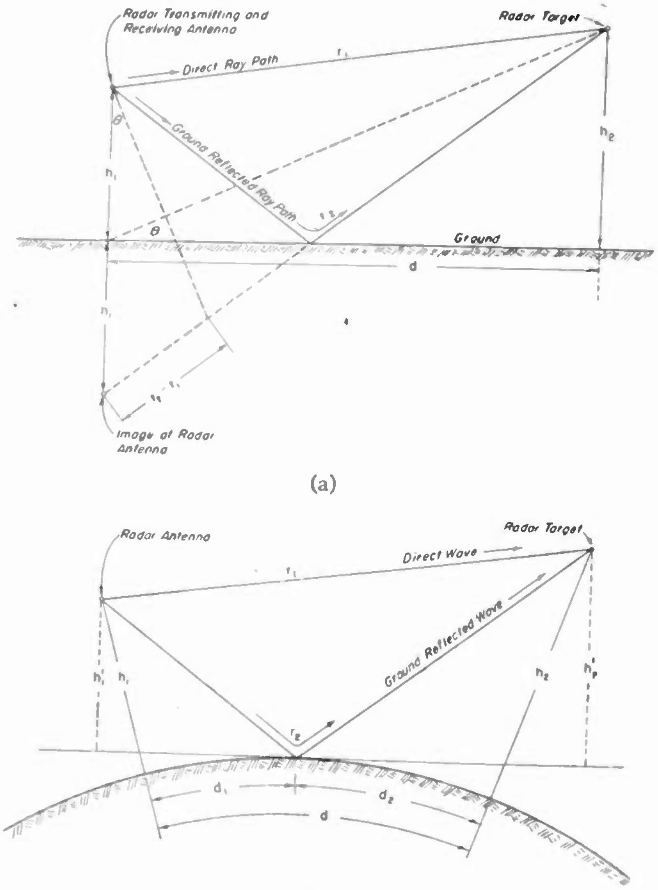
[31] R. W. George, "Field strength of motor car ignition between 40 and 450 megacycles," PROC. I.R.E., vol. 28, pp. 409–412; September, 1940.

currently available receiver designs employing radio-frequency amplifiers, although, as previously mentioned, the use of crystal mixers may result in somewhat lower noise figures for frequencies above 1000 megacycles. $\overline{NF}'$ with the receiver coupled to an antenna is very nearly the same, near 200 megacycles, as the noise figure $\overline{NF}$ measured in the laboratory. On Fig. 2, the dashed curve labelled $2R'_{\max}$ shows the average effect of the external cosmic noise in reducing the maximum range of a radar set. We see by this curve that this external noise would be expected to have very little effect on the effective range of even a very sensitive radar at frequencies above 100 megacycles where most American radar sets were operated during the past war, but would be of some importance for lower-frequency radar sets such as those used in some other countries. As an estimate of what may conceivably be expected in the future, we have shown on Fig. 3 the results to be expected from a receiver with a noise figure of 1.25, i.e., about one decibel. The curve labelled $\overline{NF}''$ gives the corresponding value of the effective noise figure for such a receiver coupled to an antenna with a transmission-line loss $L_r = 0.95$. The large differences between $\overline{NF}'$ and $\overline{NF}''$, which exceed 14 decibels at frequencies above 1000 megacycles, represent a challenge to the receiver engineers; reductions in these differences usually may be immediately translated into increased radar ranges or into reduced transmitter powers.

IV. THE EFFECT OF THE GROUND AND ATMOSPHERE ON THE MAXIMUM RANGE OF A RADAR SET

The ground reflects a part of the energy radiated from the radar antenna and this combines at the target with the energy flowing directly from the radar antenna to produce an illumination of the target which may be either greater or less than that in the absence of the ground, depending upon the relative phases of these direct and ground-reflected waves. These relative phases depend in turn upon the relative optical lengths of the direct and ground-reflected wave paths and the change in phase of the ground-reflected wave upon reflection at the earth. The energy received back from the target at the receiving antenna travels by identical direct and ground-reflected paths and is thus increased or decreased again at the receiving antenna by the same amount as at the target. Thus, if we write $g(\theta)$ for the ratio of the *field intensities* to be expected at the target in the presence of the earth relative to that which would be obtained at this distance in the absence of the earth, then the ratio of the received *energies* at the target with and without the earth will be modified by a factor $g^2(\theta)$, while the energies received back at the radar receiving antenna will be modified by a factor $g^4(\theta)$. Thus the effect of the ground can be included by multiplying the right-hand side of (12) by the factor $g^4(\theta)$, or by multiplying (13) or (24) by the factor $g(\theta)$, and this forms the basis for (1), giving the maximum range in the presence of the earth $R_{\max}$ equal to the free-space maximum range $R'_{\max}$ multiplied by the factor $g(\theta)$.

The above discussion has reduced the problem of determining the effect of earth reflections to that of determining the ratio $g(\theta)$ of the field intensity to be expected at the target in the presence of the earth relative to the expected free-space field intensity at the target. The general nature of this ratio $g(\theta)$ may be understood most simply by evaluating it for an assumed plane, perfectly reflecting earth and a homogeneous atmosphere, and then later introducing the effects of the earth's curvature, its finite conductivity, its roughness, and its actual atmosphere. Fig. 4(a) gives the geometry of the idealized problem. The radar antenna is at a height h_1 above the earth, while the target is at a distance d at an angular height θ and at an actual height h_2 above the earth. If we write E_1 for the free-space field intensity to be expected at a unit distance from the radar antenna in the direction of the direct-ray path to the target and E_2 for the free-space field intensity to be expected at a unit distance from the radar antenna in the direction of the ground-reflected-ray path, then the electric field F_g to be expected at the target in the presence of the earth is given by



(a)

(b)

Fig. 4—The geometry of radar propagation over a plane and spherical earth.

$$F_g = \frac{E_1 \cos\left[2\pi(r_1 - ct)/\lambda\right]}{r_1} + \frac{E_2 \cos\left[\pi + 2\pi(r_2 - ct)/\lambda\right]}{r_2}. \tag{30}$$

The first term in the above equation corresponds to the direct wave from the radar antenna which travels a distance r_1 to the target, while the second term above corresponds to the wave reflected at the ground and which travels a distance r_2 in going from the radar antenna to the target. The above equation corresponds to horizontal polarization and the phase term π in the ground-reflected wave term allows for the phase reversal on reflection at the earth. Equation (30) may also be written

$$F_0=\frac{E_1}{r_1}\left\{\cos\left[2\pi\left(\frac{r_1+r_2}{2}-\frac{[r_2-r_1]}{2}-ct\right)/\lambda\right]\right.$$
$$\left.-\frac{E_2r_1}{E_1r_2}\cos\left[2\pi\left(\frac{r_1+r_2}{2}+\frac{[r_2-r_1]}{2}-ct\right)/\lambda\right]\right\}. \qquad (31)$$

When the distance d is very large E_1 and E_2 become equal to the field at the maximum of the radar beam, r_1 is approximately equal to r_2, while the path difference $(r_2-r_1)\cong 2h_1 \sin\theta$. When these approximations are used in (31) and the formulas for the cosine of the sum and difference of two angles applied to the two cosine terms in (31), we obtain

$$F_0=\frac{E_1}{r_1}\sin\left[2\pi\left(\frac{r_1+r_2}{2}-ct\right)/\lambda\right]2\sin\left[\frac{2\pi h_1\sin\theta}{\lambda}\right]. \qquad (32)$$

The expected field intensity F_1 in the absence of the earth would be given by the magnitude of the first term in (31), and thus the ratio $g(\theta)$ of the field intensity to be expected at the target in the presence of the earth relative to the expected free-space field intensity may be written

$$g(\theta)=|F_0|/|F_1|=2\sin\left[\frac{2\pi h_1\sin\theta}{\lambda}\right]$$
$$\text{(plane earth; horizontal polarization).} \qquad (33)$$

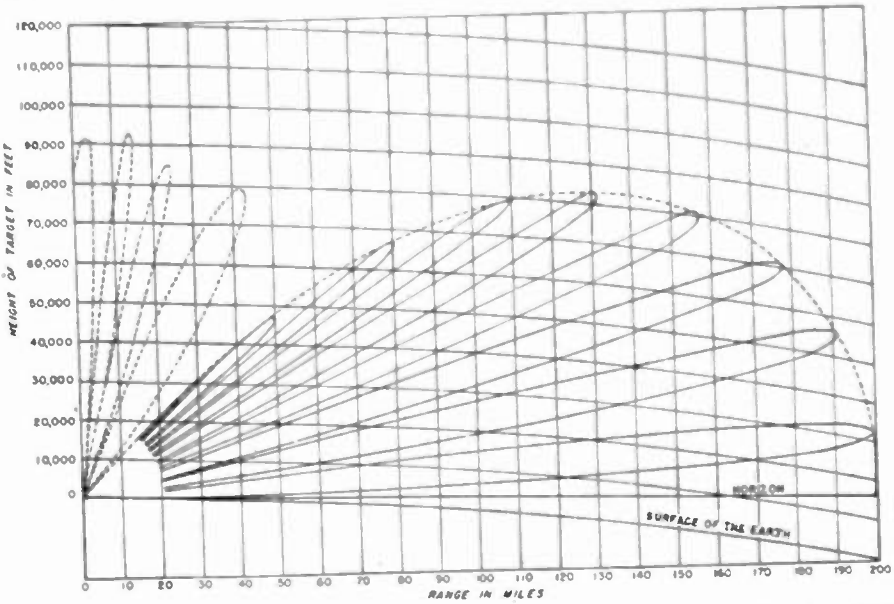


Fig. 5—Effect of the ground and the directivity of the radar antenna on the maximum range of an SCR-270 radar set. The dotted curve shows the effect of the radar antenna directivity and represents the envelope of the maximum ranges for any height of antenna. The solid curve gives lobes resulting from ground reflection and represents the actual expected maximum range for an antenna 200 feet above the sea, with the radar antenna oriented so that the maximum radiation is in the horizontal direction.

We see by the above that the value of $g(\theta)$ oscillates sinusoidally between 0 and 2 as the angular height θ is increased or, since $\sin\theta=h_2/d$, as the actual height h_2 is increased at a fixed distance. Figs. 5 and 6 show examples of the lobe structure resulting from the application of this factor $g(\theta)$ to calculations of the maximum range of an SCR-270; in these examples the effects of the curvature of the earth and of the polar radiation characteristic of the radar antenna are also included. If we write $(2\pi h_1 \sin\theta/\lambda)=n\pi/2$, then $g(\theta)$ will be equal to zero for even values of $n=0, 2, 4, 6, \cdots$ and will be equal to 2 for odd values of $n=1, 3, 5, 7, \cdots$. Thus, the angles θ at which $g(\theta)$ has a value of 0 or 2 may be determined by

$$\sin\theta=\frac{n\lambda}{4h_1}\ (g(\theta)=0 \text{ for } n \text{ even and } g(\theta)=2 \text{ for } n \text{ odd}). \qquad \textbf{(34)}$$

The total number of lobes is equal to the number of half-wavelengths that are contained in the height h_1 of the radar antenna above the ground. For example, with an SCR-270 radar at a height of 200 feet above the sea as shown on Fig. 5, since $\lambda=10$ feet there will be a total of 40 ground-reflection lobes. Only the first 9 of these lobes are shown on Fig. 5; the remaining 31 lobes are contained in the high-angle dotted lobes of Fig. 5, these dotted lobes arising from the free-space directional characteristics of the radar antenna array and not the effect of the ground.

We see by (34) that the first null corresponds to $n=0$ and this corresponds to $\theta=0$, i.e., at the horizon. This region of short radar ranges near the horizon is of great practical importance, as may be seen by reference to Fig. 5. Thus this SCR-270 radar might be expected to detect an aircraft flying at an altitude of 30,000 feet at twice the value of its free-space maximum range, that is, at 200 miles, but this same radar would be able to detect an aircraft flying at 1000 feet altitude at a range of only 50 miles. This gap in effective radar coverage may be narrowed by decreasing the angular height of the maximum of the first lobe; reference to (34) indicates that this may be achieved either by increasing the height of the radar antenna or by increasing the radio frequency at which it is operated.

Fig. 5 indicates the importance of considering earth curvature in calculations of the effective coverage of a radar set. The appropriate methods of calculation in this case are discussed in an earlier paper by one of the authors,[32] in which the following formula is given:

$$g(\theta) = [1 + (DR)^2 + 2DR \cos \phi]^{1/2} \text{ (spherical earth)} \quad (35)$$

where ϕ is the relative phase between the direct and ground-reflected waves at the target, D is the divergence factor measuring the decrease in intensity of the ground-reflected wave due to the divergence of the reflected energy at the curved surface of the earth, and R is the factor indicating the effect of polarization, and the finite conductivity and roughness of the ground on the reflected wave intensity. The principal difference between vertical and horizontal polarization is the smaller reflection coefficient R for vertically polarized waves reflected near Brewster's angle;[33] this has the effect of reducing the range somewhat in the maximum of the lobes and of filling in the null zones near Brewster's angle. The largest effect of the curvature of the earth on the effective coverage of a radar set is the modification of the actual heights h_1 and h_2 of the radar antenna and the target, respectively, to new effective heights h_1' and h_2', as shown on Fig. 4(b). Methods for calculation of h_1', h_2', D, R, ϕ, d_1 and d_2 are all given in the reference cited and so will not be repeated here.

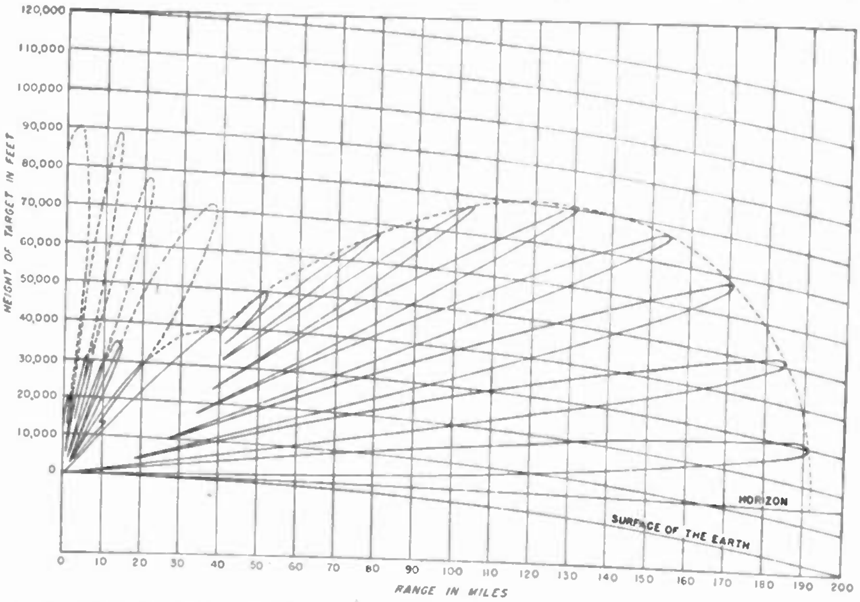


Fig. 6—Maximum range of SCR-270 with array tilted back 2 degrees. The dotted curve represents the envelope of the maximum ranges for any height of antenna; the solid curve represents actual range for an antenna 200 feet above the sea.

The systematic effects of air refraction in reducing the effects of the curvature of the earth are also given in the reference just cited[32] where it is shown that these systematic effects may be included in the calculations by using an effective radius of the earth equal to 4/3 of its actual radius. This method is applicable up to heights of the order of 30,000 to 40,000 feet, the value 4/3 being appropriate in temperate latitudes.[34]

Another effect of the atmosphere of especial importance for the detection of ships or of low-flying targets is the formation of ducts near the surface of the earth which tend to guide the waves along the surface with abnormally low attenuation. The phenomenon is now known as superrefraction, although it was originally called "anomalous propagation." These ducts are of increasing importance at the higher radio frequencies and sometimes make possible the detection of targets near the surface of the sea at distances of the order of 100 miles or more. These ducts are layers usually occurring in the first few hundred feet of the lower atmosphere where the downward curvature of a radio ray exceeds the earth's curvature. A sharp decrease of the humidity with height associated with a temperature inversion is the usual cause of such an atmospheric duct. At microwave frequencies, ducts are sometimes responsible for a substantial improvement in the radar coverage. Since these ducts are not always present, this additional coverage is not reliable.

Since all of the calculations and examples in this paper were confined to frequencies below 3000 megacycles, it has been unnecessary to include the effects of

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[32] K. A. Norton, "The calculation of ground wave field intensity over a finitely conducting spherical earth," Proc. I.R.E., vol. 29, pp. 623–639; December, 1941.

[33] K. A. Norton, "Ground wave propagation," Federal Communications Commission Report No. 47475, presented at the Fourth Annual Broadcast Engineering Conference, February 10–21, 1941, Ohio State University.

[34] Work now in progress at the Central Radio Propagation Laboratory of the National Bureau of Standards is expected to provide a method of calculation for all heights and at any geographic location.

atmospheric absorption in the calculations of maximum ranges. At still higher frequencies such effects become of increasing importance. Van Vleck[35] predicted as early as 1942 that the atmosphere would absorb these higher frequencies, and further work by Van Vleck[36] and others has established an absorption resonance frequency due to water vapor in the neighborhood of 23,000 megacycles and two such absorption bands due to oxygen in the neighborhood of 60,000 and 120,000 megacycles. Rain also causes attenuation by absorbing and scattering the radio energy. Summaries of the practical effects of this attenuation are given in reports by Goldstein[37] and Fich.[38] If we write F_1 and F_2 for the expected field intensities with and without the atmosphere, respectively, for one-way transmission for a distance R meters, then the absorption coefficient α in (1) is defined by the relation

$$\alpha R \equiv \log_e (F_2/F_1). \qquad (36)$$

Thus the previous equations in this report can be modified so as to include the effects of atmospheric absorption by multiplying (7) and (9) by the factor $e^{-2\alpha R}$ and (10) and (12) by the factor $e^{-4\alpha R}$. Note that $R'_{\max}$ is, by definition, the free-space maximum range in the absence of atmospheric absorption; thus (1) is obtained by solving (12) after the factor $e^{-4\alpha R_{\max}}$ has been introduced and E_r replaced by $E_{\min}$. The absorption coefficient α is expressed in nepers per meter and is equal to 0.115 times the absorption coefficient expressed in decibels per meter. Below 150,000 megacycles, the atmosphere absorbs the most energy at frequencies near 60,000 megacycles where $\alpha \cong 1.4 \times 10^{-3}$; the great importance of this absorption may be seen by noting that the maximum range is reduced to a value $1/e$, i.e., to 0.368 of its value without absorption at a distance $R_{\max} = 1/\alpha = 714$ meters. On the other hand, the corresponding values at 10,000 megacycles are $\alpha = 1.1 \times 10^{-6}$ and $1/\alpha = 900{,}000$ meters with no rain; at 10,000 megacycles, a heavy rain may increase the absorption so that $\alpha = 8 \times 10^{-5}$ and $1/\alpha = 12{,}500$ meters. At still lower frequencies atmospheric absorption becomes increasingly less important and is of practically negligible importance below 3000 megacycles.

Fig. 6 illustrates the effect on the radar coverage of tilting the radar antenna array back a few degrees so that the ground-reflected waves will be weakened relative to the direct waves. This has the effect in (31) of making E_2 smaller than E_1 even when d is large. In this example, an SCR-270 broadside array (consisting of 32 half-wave dipoles in an array 4 dipoles wide and 8 dipoles high) is tilted back by 2 degrees. Comparing the coverages shown on Figs. 5 and 6, we see that the maximum range is reduced from 200 to 190 miles but the gap in coverage for aircraft flying at 30,000 feet, which occurs on Fig. 5 between 22 and 29 miles, is eliminated on Fig. 6. Thus, by sacrificing some maximum range, it is possible to gain more nearly solid coverage at the shorter distances by tilting the array by only 2 degrees. At microwave frequencies, where higher antenna gains are more readily obtainable, it is possible to so shape the radiation characteristic in the vertical plane as to eliminate objectionable ground reflections almost completely; this practice does, of course, eliminate the two-fold increase in the free-space range which occurs at the lobe tips in the presence of ground reflection.

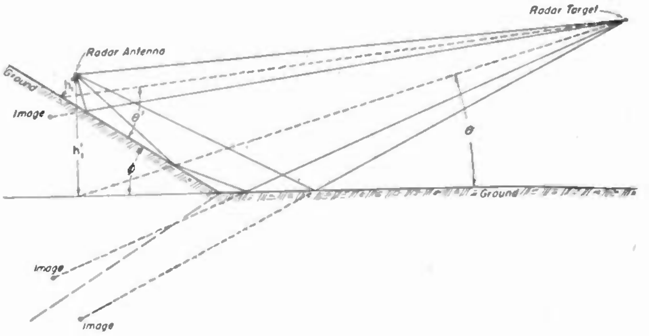

Fig. 7—The number of ground-reflected waves can be tripled by placing the radar on a sloping site. The energy received from a distant target will be increased nearly 16-fold by the slope if the radar antenna is placed at a height $h_1 = \lambda/4 \sin \theta'$ and then moved back along the slope until $h_1' = \lambda/4 \sin \theta$.

Sometimes advantage can be taken of a sloping site to increase the maximum range of a radar set. The geometry of this case is shown on Fig. 7, which is self-explanatory. The maximum range may be increased for targets at certain angles to nearly twice the values obtainable without the slope or to four times the maximum free-space range. This will occur at an angle such that the four waves from the radar antenna and its three images all are in phase at the target. The slope, however, must be sufficiently extensive to support at least one complete Fresnel zone, as discussed in the Appendix.

V. THE NATURE OF ECHOING AREAS OF RADAR TARGETS[39]

If S_i represents the energy flow per square meter incident on the target and S_r represents the energy flow per square meter at a distance R from the target, reflected back in the direction of the radar antenna with the same polarization as that of the incident wave, then, by definition of echoing area A_e, we have

$$S_r = \frac{A_e S_i}{4\pi R^2}. \qquad (37)$$

If we write A_s for the total scattering area of the target and g for the ratio of the actual energy per square square meter scattered back in the direction of the

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[35] J. H. Van Vleck, "Atmospheric absorption of microwaves," Radiation Laboratory Report No. 43-2, April 27, 1942.

[36] J. H. Van Vleck, "Further theoretical investigations on atmospheric absorption of microwaves," Radiation Laboratory Report No. 664, March 1, 1945.

[37] L. Goldstein, "Absorption and scattering of microwaves by the atmosphere," Report WPG-11, Wave Propagation Group, Columbia University, Division of War Research, New York, N. Y., May, 1945.

[38] S. Fich, "Wave propagation study for line of sight communication and navigation," AAF Project AC 230,04, Wave Propagation Group, Columbia University, Division of War Research, New York, N. Y., September, 1945.

[39] More complete formulas are given in the Radio Research Laboratory Report No. 4, Harvard University, October 22, 1942.

radar antenna to that which would be scattered in this direction if the scattering were the same in all directions, then, by definition of A_s, we also have

$$A_e = gA_s. \qquad (38)$$

Thus g is simply the gain in the direction of the radar transmitter of the re-radiating target relative to an isotropic radiator.

Now consider the scattering area of a large, flat, perfectly conducting circular disk, normal to the direction of the incident radiation. A disk is considered to be large if its radius a is large compared with $\lambda/2\pi$. As the radius of the plate is increased without limit or, conversely, the wavelength of the incident radiation decreased without limit, the scattering area of the disc A_s approaches its actual visual area $A = \pi a^2$ since edge effects become negligible in comparison to the total energy scattered. Since the disk is assumed to be large, flat, and perfectly conducting, equal currents with equal phases will flow throughout the entire face of the disk exposed to the incident radiation in such a manner that its scattering polar radiation characteristic will be the same as that of a large broadside antenna with reflecting screen and covering the same area as that of the disk; thus the ratio g will be simply the maximum gain G of such a broadside antenna relative to that of an isotropic radiator

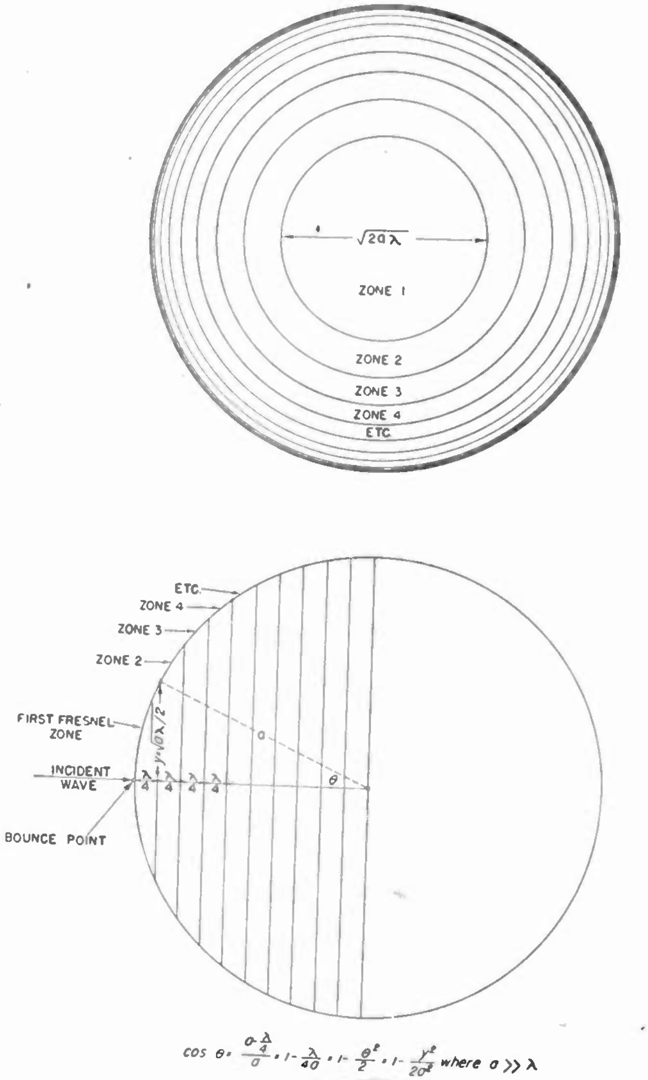
$$g = G = \frac{4\pi A}{\lambda^2}\cdot \qquad (39)$$

The right-hand side of (39) above is the expression for the maximum gain of a large broadside array with reflecting screen and with uniform currents in all of the elements in terms of its area A. Now, since the total scattering area $A_s = A$, we obtain from (38) and (39) the following expression for the echoing area of a flat, perfectly conducting disk normal to the direction of the incident radiation and with a radius large in comparison to $\lambda/2\pi$:

$A_e = 4\pi A^2/\lambda^2$ (large, flat, perfectly conducting disk normal to the direction of the incident radiation). (40)

All of the above arguments are equally applicable to flat plates of other shapes with the same visual area as the disk, provided both linear dimensions of these plates are large compared to $\lambda/2\pi$. We see by the above expression that the echoing area of such a plate increases directly in proportion to the square of its visual area A and is directly proportional to the square of the radio frequency. It is this reflecting characteristic of targets which is responsible for the very large echoes sometimes observed at the higher frequencies. However, since the target must be flat within a small fraction of a wavelength over the entire area A in order to produce the echo represented by (40), such large echoes are not often encountered. Furthermore, since the echo from a large flat plate is concentrated in a narrow beam, a small change in the aspect of the target reduces the echo considerably; thus, for the disk, a change in aspect $\theta = 7.3(\lambda/a)$ degrees will reduce the echoed energy to half of that reflected at normal incidence. In the limit, this highly directional reflection from large flat plates is simply the specular reflection of geometrical optics and consequently, in the radio case, is often called specular reflection.

The more usual target consists of curved surfaces. As an introduction to this problem, we will consider the echoing area of a perfectly conducting sphere of radius a very much larger than $\lambda/2\pi$. Fig. 8 gives the geometry of the problem and shows two cross sections of the sphere with its surface divided into Fresnel zones each of depth $\lambda/4$ along the direction of the incident radiation.[40] For a very large sphere, the energy scattered back in the direction of the source of the incident radiation will be essentially determined by that reflected from the first zone, since the waves reflected from zone 2 will have an opposite phase to and will thus very nearly cancel those reflected from zone 3 etc.; in fact, it can



$$\cos\theta = \frac{a-\frac{\lambda}{4}}{a} = 1-\frac{\lambda}{4a} = 1-\frac{\theta^2}{2} = 1-\frac{y^2}{2a^2} \text{ where } a \gg \lambda$$

$$\therefore y = \sqrt{a\lambda/2}$$

Fig. 8—Geometry of the echoing area of a sphere.

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[40] A discussion of the Fresnel zones involved in the reflection from a plane earth is to be found in the Appendix of this paper.

be shown that the total echoed field intensity from all except the first zone has the correct magnitude and phase to cancel just half of the field from the first zone, leaving for the total echoed field intensity from the entire sphere just half of that to be expected from the first zone alone. Furthermore, the echoed field intensity from the first zone considered as a curved surface will be equal to $(2/\pi)$ times the value to be expected from a flat plate of the same visual area; thus, the *energy* reflected from a large sphere will be equal to $(1/\pi)^2$ times that to be expected from a flat plate with the same visual area as that of the first Fresnel zone. This visual area of the first Fresnel zone may be determined from the geometry of Fig. 8

$$A_1 = \pi a\lambda/2 \text{ (visual area of the first Fresnel zone).} \qquad (41)$$

Since, for a large sphere, the first Fresnel zone has a radius which is large with respect to the wavelength, we may obtain the echoing area of this zone considered as a flat plate by substituting (41) in (40), and after multiplying this result by the factor $(1/\pi)^2$, we obtain finally for the echoing area of a large perfectly conducting sphere with a radius $a >> \lambda/2\pi$:

$$A_e = (4\pi/\lambda^2)\cdot(\pi a\lambda/2)^2\cdot(1/\pi)^2 = \pi a^2. \qquad (42)$$

We see by the above that the echoing area of a large, perfectly conducting sphere is just equal to its visual area A and independent of the radio frequency. A similar argument shows that the echoing area of any curved surface is equal to

$$A_e = \pi\rho_1\rho_2 \qquad (43)$$

where ρ_1 and ρ_2 are the principal radii of curvature at the bounce point, each being assumed to be large relative to $\lambda/2\pi$. As may be seen in Fig. 8, the bounce point of a curved surface is the point nearest the source of the incident radiation. From the above discussion of the sphere, it is clear that (43) is approximately correct for surfaces containing at least one Fresnel zone. Equation (43) provides the explanation for the effective echoing area of an aircraft target which is known to be substantially independent of the radio frequency in the range of frequencies ordinarily used for radar. Thus, the effective echoing area of an aircraft may be expressed:

$$A_e = \sum_i \pi\rho_{1i}\rho_{2i} \text{ (effective echoing area of an aircraft target).} \qquad (44)$$

The effective echoed energy from an aircraft is thus considered to be the sum of the energy echoed from the various curved surfaces encountered by the incident radio wave. As the aspect of the aircraft changes, the effective echoing area will change due to changes in the radii of curvature ρ_{1i} and ρ_{2i} at the various bounce points. It is of interest to note that, to the extent that (44) provides an adequate measure of the echoing area of an aircraft target, this echoing area is independent of the radio frequency and polarization; this agrees with experience.

If we compare (42) with (38) and remember that the scattering area of a large, perfectly conducting object is just equal to its visual area, we see that $g=1$, i.e., that the energy reflected from the sphere is equal to that scattered from an isotropic scatterer. Although the proof will not be given here, it can be shown that the sphere will scatter energy equally in all directions and thus, when illuminated from a large distance, acts as an isotropic radiator.

We will turn now to a discussion of radar targets with dimensions small in comparison to the wavelength. For a perfectly conducting small sphere, the solution was first obtained by Lord Rayleigh in 1881.[41]

$$A_e = 4.41(10)^4 a^6/\lambda^4 \quad (a \ll \lambda/2\pi). \qquad (45)$$

The above expression shows that the echoing area for a small sphere is proportional to the fourth power of the radio frequency. This frequency dependence of the scattering from small objects is true irrespective of their shape or electrical constants and forms the basis for the explanation of the blue color of the clear sky as first given by Lord Rayleigh. Thus, for a small, flat, perfectly conducting disk of radius a we have

$$A_e = (3.53)10^3 a^6/\lambda^4 \quad (a \ll \lambda/2\pi). \qquad (46)$$

When the dimensions of the object are neither very large nor very small relative to the wavelength, recourse must be had to series expansions for computing the echoing area. Fig. 9 gives results obtained in this way for perfectly conducting spheres and flat disks normal to the direction of the incident radiation.[42] The envelope of the maxima and minima for the large sphere may be expressed

$$A_e = \pi a^2(1 \pm 0.144\lambda/a)^2. \qquad (47)$$

Finally, we will determine the echoing area of tuned linear antennas for which the current distribution may be assumed to be approximately the same for transmission as for reception. The effective absorbing area of a tuned matched antenna is given by (11) and, since there will be an equal amount of energy scattered by the radiation resistance as is absorbed by the load, the scattering area of such a matched tuned antenna oriented relative to the incident wave in such a manner that it absorbs a maximum of the incident radiation is also given by (11); thus

$$A_s = \frac{G\lambda^2}{4\pi} \text{ (scattering area of matched tuned antenna parallel to the incident electric field).} \qquad (48)$$

This scattered energy is reflected back in the direction of the incident radiation in a beam with a gain G so that the echoing area is given (according to (38)) by

$$A_e = GA_s = G^2\lambda^2/4\pi \text{ (matched tuned antenna parallel to the incident electric field).} \qquad (49)$$

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[41] Lord Rayleigh, Scientific Papers, I, pp. 518–536, or Philosophical Magazine, XII, pp. 81–101, 1881. This is one of a series of papers on scattering beginning with one on "The Light From the Sky, Its Polarization and Colour," Phil. Mag. XLI, pp. 107–120, 274–279; 1871.

[42] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Company, New York, N. Y., p. 563, 1941.

For a matched tuned Hertzian dipole, $G=1.5$ and $A_e=0.18\lambda^2$. For a matched half-wave dipole, $G=1.641$ and $A_e=0.22\lambda^2$. Parasitic antennas have four times the above scattering and echoing areas for matched antennas, since the incident field causes twice as much current to flow and thus four times as much energy is scattered, none being absorbed. Thus, a parasitic half-wave dipole has an echoing area $A_e=0.86\lambda^2$. The very large echoes obtainable from parasitic half-wave dipoles with the expenditure of a very small amount of material formed the basis for the radar countermeasure known as chaff, which was used by all participants during the later phases of World War II for the purpose of producing large confusing echoes on the enemy radar screen which would also tend to screen aircraft at the same slant range.

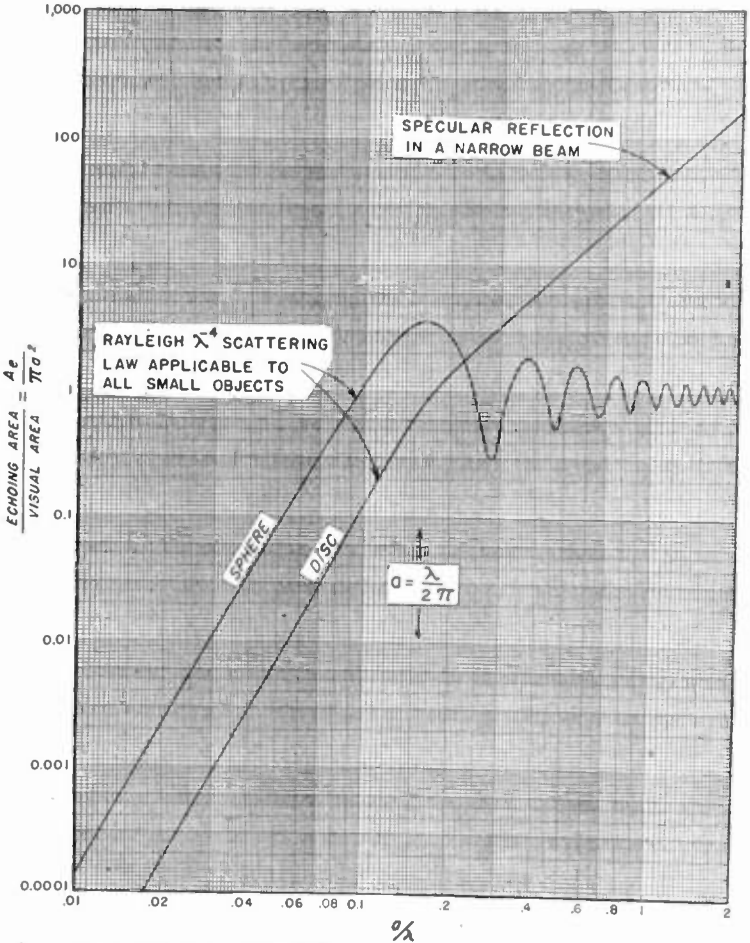

Fig. 9—The echoing area of perfectly conducting spheres and flat disks normal to the incident radiation (a = radius of sphere or disk; λ = wavelength of incident radiation).

VI. Moonshining

As an example of the utility of these maximum-range formulas, they will be applied to a discussion of the radar echoes from the moon recently observed on a modified SCR-270 early-warning radar set at the Signal Corps Engineering Laboratories. The echoing area of the moon can be determined by considering it to approximate the echoing area of a large, perfectly conducting sphere and then correcting this value for the expected effects of the actual electrical constants of the surface material of the moon and the irregularities in its terrain. The radius of the moon is equal to 1080 miles, so that its echoing area, as approximated by (42), is equal to 9.5×10^{12} square meters, which is nearly 10^{12} times the value of 10 square meters assumed in this paper for a typical aircraft target. Since the maximum radar range is proportional to the fourth root of the echoing area, the effective range of a radar set used to obtain echoes from the moon should be approximately $(10^{12})^{1/4}=1000$ times the values given in the right-hand column of Table II. Reference to Table II indicates that the only sets shown therein which might be expected without modification to detect such echoes are the SCR-270 or the MEW, since the distance to the moon is 239,000 miles. A modified SCR-270 radar set was used in the successful moonshining experiments conducted at the Fort Monmouth Signal Corps Laboratories.

Actually, because of its finite conductivity, the echo from the moon will be less than the above value because some of the incident energy will be absorbed. The effects of this absorption on the echoing area may be approximated by multiplying the echoing area by the square of the magnitude of the plane-wave Fresnel reflection coefficient R for a normally incident wave:

$$R = \frac{\left[\dfrac{\sin b'}{\epsilon}\right]^{1/2} e^{-i(\pi/4 - b'/2)} - 1}{\left[\dfrac{\sin b'}{\epsilon}\right]^{1/2} e^{-i(\pi/4 - b'/2)} + 1} \qquad (50)$$

$$\tan b' = \epsilon/x \qquad (51)$$

$$x = 1.8 \cdot 10^{15}\, \sigma/f_{Mc} \qquad (52)$$

where ϵ = dielectric constant of the moon and σ = conductivity of the moon (in electromagnetic units). If we assume that $\sigma = 10^{-14}$, corresponding to a rather poor ground conductivity on the earth, and take $f_{Mc} = 106$ megacycles, corresponding to that of the SCR-270, then $x = 0.17$. Since ϵ will very probably be greater than 4, $\tan b'$ will be very large and we can approximate to the value of R by setting $b' = \pi/2$, whence

$$R \cong \frac{1 - \sqrt{\epsilon}}{1 + \sqrt{\epsilon}} \quad (x \ll \epsilon). \qquad (53)$$

With ϵ varying from 4 to 36, the square of R varies from 0.11 to 0.51, which indicates the amount of echoing-area reduction to be expected from the above choice of electrical constants for the moon.

The possible effects of roughness must still be considered. Since we have seen in connection with the discussion of the echoing area of the sphere that the energy scattered back is essentially determined by that reflected from the first Fresnel zone, it is of interest to determine the area of this zone on the face of the moon. Using (41) we obtain, for the SCR-270 frequency, $f = 106$ megacycles and $A_1 = 3$ square miles, the radius of this zone being slightly less than one mile. Photographs

TABLE II

Type of set	Peak power (kilowatts)	P_t Avg. power (watts)	E_t[19] Pulse energy (joules)	G_t' Trans. ant. gain rel. to λ/2 ant.	G_r' Rec. ant. gain rel. to λ/2 ant.	G[34] Ant. gain rel. to isotropic radiator	$\overline{NF}$ Noise figure (ratio)	$\overline{NF}$ Noise figure (db)	L_t[16] TR box & line loss (ratio)	L_r[16] TR box & line loss (ratio)	$L_t L_r$ (db)	τ Pulse width (μs)	B Band width (Mc.)	τB	V Visibility factor (ratio)	f_r Recurrence freq. (cycles)	f Carrier freq. (Mc.)	R1 Range index (miles)
SCR-268	67.5	2210	0.54	33	40	66	126	21	0.95	0.95	0.5	8	1.5	12	2.50	4098	205	79.5
SCR-270-B	84.4	950	1.52	76	76	125	28	14.5	0.95	0.80	1	10	0.55	5.54	4.01	625	106	251
SCR-270-B	30.4	950	1.52	76	76	125	29	14.5	0.95	0.80	1	50	0.53	26.5	9.85	625	106	200
SCR-271-D	84.4	950	1.52	50	50	97	28	14.5	0.95	0.80	1	10	0.55	5.54	4.01	625	106	220
SCR-271-D	30.4	950	1.52	50	50	97	28	14.5	0.95	0.80	1	50	0.53	26.5	9.85	625	106	176
SCR-588 (One bay)	125	125	0.31	64	64	105	70.8	18.5	0.80	0.80	1.5	2.5	0.74	1.85	1.76	400	200	106
SCR-588 (Both bays)	125	125	0.31	128	128	210	70.8	18.5	0.9	0.70	2.0	2.5	0.74	1.85	1.76	400	200	141
SCR-588 (Pre-amp.)	125	125	0.31	128	128	210	15.8	12.0	0.9	0.70	2.0	2.5	0.74	1.85	1.76	400	200	212
SCR-602 T-3	130	22.8	0.091	153	153	250	15.8	12	0.95	0.80	1.5	.7	1.7	1.2	1.28	[illegible]	1073	82
SCR-602 T-6	120	95	0.24	25	25	41	15.8	12	0.95	0.80	1	2	1.7	3.4	1.92	400	212	89
SCR-602 T-7	27	35	0.11	50	50	97	7.1	8.5	0.95	0.80	1	4	1.8	7.2	4.01	324	600	83
SCR-602 T-8	250	75	0.375	100	100	164	11.2	10.5	0.95	0.80	1	1.5	5.75	8.65	5.65	200	600	100
SCR-615	400	310	0.4	880	880	1440	126	21	0.90	0.70	2	1	2.7	2.7	1.63	775	3000	93
SCR-636 T-1	120	95	0.24	64	64	105	5	7	0.95	0.80	1	2	1.5	3	2.13	400	212	184
MEW	400	160	0.4	9330	9330	15300	40	16	0.90	0.70	2	1	2	2	1.70	400	2800	400
Japanese	6.5	77	0.077	24	24	39	88	19.5	0.95	0.95	0.5	12	0.45	5.65	2.9	1000	97	58
Small[35] Wurzburg	7	52.5	0 0.014	140	140	230	107	20.3	0.98	0.98	0.2	2	0.5	1	0.739	3750	560	52
Giant[35] Wurzburg	7	52.5	0 0.014	950	950	1560	107	20.3	0.98	0.98	0.2	2	0.5	1	0.739	3750	560	135
Freya	100	175	0.35	24	24	39	5	7	0.95	0.80	1	3.5	0.29	1	1.485	500	120	178
Freya	100	350	0.35	22	24	39	5	7	0.95	0.80	1	3.5	0.29	1	1.18	1000	120	189

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[1] These values of peak power were obtained from "A survey of radar equipment in production and development," compiled by Section D-1, of National Defense Research Committee, J. G. Trump, Secretary.

[2] These values of peak power were computed from the average power output.

[3] These values of average power were computed from the peak powers.

[4] Average power output measured by Equipment Unit, Operational Research Group. It was found that the average power output was approximately independent of pulse width for a constant power input.

[5] Theoretical antenna gains computed on the assumption that the current in each antenna element is the same. These gains are possibly somewhat optimistic.

[6] This antenna gain is a theoretical value reduced by a factor of 0.6 to take account of the distribution of current. The antenna was assumed to be 10×25 feet.

[7] These values of $\overline{NF}$ were obtained from measurements made by the Bell Telephone Laboratories.

[8] Estimated noise figures based on measurements made by Radiation Laboratory on similar type of equipment at these frequencies.

[9] These "effective pulse widths" were obtained from "A survey of radar equipment in production and development," compiled by Section D-1 of the National Defense Research Committee.

[10] These values of effective pulse widths were obtained from measurements made by the Equipment Unit, Operational Research Group, and represent the minimum and maximum pulse widths which can conveniently be obtained using external controls.

[11] These bandwidths were obtained from "A survey of radar equipment in production and development," compiled by Section D-1 of National Defense Research Committee. They may not represent the effective bandwidths used in the equations in this appendix.

[12] Effective bandwidths measured by Equipment Unit, Operational Research Group.

[13] These data obtained from "Survey of radar equipment in production and development," compiled by Section D-1 of National Defense Research Committee.

[14] These data obtained from A. V. Haeff of the Naval Research Laboratory.

[15] This figure was obtained by assuming that the operation would be practically the same as with the SCR-268, which uses a similar type of first circuit.

[16] These transmission-line, antenna and transmit-receive-box losses are "educated guesses."

[17] Since the antenna on this set is tilted upward, in order to obtain complete vertical coverage, there will be no ground reflections in the direction of the maximum of the beam, so that the maximum range is the same as if the antenna were in free space; i.e., half the value of the range index.

[18] This is a theoretical value which has been checked experimentally. It was obtained from RCA report TR-8-C, prepared by George H. Brown.

[19] These values of pulse energy were computed from the average power and the pulse-recurrence frequency.

[20] Based on measurements made at Bell Telephone Laboratories using the same type tubes.

[21] Effective pulse width obtained by integrating the pulse as traced from the monitor oscilloscope.

[22] Obtained by integrating the over-all frequency-response curve.

[23] These values obtained from R. E. Crane of Bell Telephone Laboratories.

[24] Measured by Dakin of Research Enterprises, Ltd.

[25] Rough estimate by Bridgeland of Research Enterprises, Ltd.

[26] These values obtained from C. G. Pierce of General Electric Company.

[27] These values obtained from Mantsetti of Camp Evans Signal Laboratory.

[28] This value obtained from K. T. Bainbridge of Radiation Laboratory.

[29] This transmitter is an SCR-602 T-6 revised for 60-cycle operation, and it is assumed that the performance will be the same.

[30] This value was estimated since no data were available.

[31] These values obtained from TRE Report 6/R/25, titled, "Final technical report on the German RDF equipment captured at Bruneval (Wurzburg) on 28th February, 1942."

[32] These values obtained from TRE Report No. 5/37, "Report on enemy 120 Mc/s RDF chain."

[33] These values obtained from E. G. Schneider of Radiation Laboratory.

[34] *Note on Antenna Gain.* The antenna gain referred to that of an isotropic radiator is given in the chart as well as the value relative to that of a half-wave antenna. In view of the more fundamental nature of the gain relative to an isotropic radiator, it is urged that it be adopted universally. If we write G' for the gain relative to that of a half-wave antenna, $G = 1.641G'$. If we write G'' for the gain relative to that of an infinitesimal electric or magnetic dipole, $G = 1.5G''$.

[35] *Note on German Equipment.* The TRE Report 6/R/25, "Final technical report on the German RDF equipment captured at Bruneval on February 28, 1942," brings to light two very interesting points. The first is that the Germans were using intermediate-frequency bandwidths of $1/\tau$, which is shown to be optimum in this report. The second point of interest is the statement by TRE: "... This shows that the receiver gain was usually operated at a high level, giving about 1 cm of noise on the table. Owing to the high recurrence frequency used (3750 pulses per second), it would be possible to see signals below the noise level. ..." This "seeing below the noise" is equivalent to saying that the visibility factor as described in this report is less than 1. The actual calculated value is 0.78. This is of especial interest when it is noted that this German equipment was designed at least as early as 1939!

of the moon show irregularities which are much larger than this, so that it seems likely that the actual echo to be expected from the moon will consist of an effective echo made up of the reflections from the innumerable bounce points at various elevated places on the surface of the moon in somewhat the same way that an aircraft echoes the incident radiation, as explained in connection with (44), plus some specular reflection from possible flat plateaus normal to the direction of the incident radiation. The echoes so far obtained from the moon have not been stable but, surprisingly enough, have varied considerably, even between successive pulses. Often, for several weeks at a time, no echoes are obtained; this suggests that the echoes so far obtained are due to specular reflection from flat or slightly concave surfaces which may become normal to the line of sight as the moon varies its aspect to the earth. A comparatively slow variation in received pulse intensity is to be expected and can be explained as due to (a) the variations in the echoed field intensity due to the factor $g^2(\theta)$ arising from the systematic effects of earth reflections, and (b) the added variations in intensity due to irregularities in the earth's terrain over the varying area of the first Fresnel zone; on the days when echoes are obtained, this slow variation due to the lobe structure is observed. The echoes so far obtained have all been received near the time when the moon is rising or setting, since the antenna array could not be tilted, and this permitted advantage to be taken of the extra 12 decibels obtainable in the maxima of these ground-reflection lobes. Thus the echoes from the moon are only seen when it is at an angular elevation θ above the horizon in the neighborhood of one of the values of θ given by (34) with odd values of n and as shown graphically on Fig. 5.

As a check on the possible influence of the earth's atmosphere on the echoes from the moon, the antenna was directed on the sun and measurements made of its noise radiation. At sunrise and sunset, as the sun passed through the first few ground-reflection lobes of the antenna pattern, the received external noise varied quite regularly and smoothly as would be expected from a noise source with the surface area of the sun; the fact that this noise was received at all, and the further fact that it did not vary erratically with time, indicates that atmospheric absorption is probably not the explanation for the variable echoes from the moon as received at 106 megacycles. The surface temperature of the sun required to explain the observations was of the order of 10^6 degrees Kelvin.

From the above discussion we see that the actual echo expected from the moon would probably be somewhat weaker than the value from a perfectly conducting sphere of the same radius even with no atmospheric absorption along the path, so that it is necessary to modify the SCR-270 to obtain echoes with a satisfactory intensity. Since there is no problem of discriminating between echoes from the moon and other celestial bodies, the simplest method of accomplishing this is to increase the number of joules radiated in each pulse by increasing the length of the pulses. The reported pulse energy used in the SCR-270 experiments was equal to 300 joules, i.e., a peak power of 3000 watts with a pulse duration of 0.1 second; this is 23 decibels more pulse energy than that used in the calculations of Table II. Full advantage of this large pulse energy cannot be obtained, of course, unless the bandwidth of the receiver is decreased so as to satisfy approximately the relation $\tau B=1$. In the SCR-270 experiments the receiver bandwidth was narrowed to about 40 cycles, so that $\tau B=4$, resulting in a mismatch of only 2 decibels from the optimum value but being an improvement of 2.7 decibels over the value in Table II. In the SCR-270 experiments two antenna arrays were combined with a net gain relative to a half-wave antenna of about 100, which results in an improvement of about 2 decibels relative to the value assumed in Table II. The receiver noise figure was about 6 decibels better than that given in Table II. Finally, assuming a moon dielectric constant $\epsilon=6$, corresponding to an energy loss of 7.5 decibels, and adding and subtracting the above estimates, we find that the expected echo should be $(-7.5+6+23+2.7+2)=26.2$ decibels stronger than the minimum detectable value. The above calculations do not take into account the very low value of pulse-recurrence frequency ($F\cong 0.2$ cycles) used in the moon experiments, and thus may be optimistic by a few decibels. However, since the observed echoes were considerably stronger than the noise, there is very little margin for possible absorption in the atmosphere, and this is in agreement with the observations made on solar noise with the same antenna and receiver.

It was hoped that the successful detection of radio waves reflected from the moon might lead to a method of transmission of intelligence, including such complex forms as those involved in television, between points widely separated on the earth's surface, e.g., between London and New York. It was expected that such transmissions on very-high-frequencies or higher would be free from the multipath effects of ionospheric transmission so disastrous in the high-frequency band. However, experiments to date appear to indicate that the irregularities of the surface of the moon will cause correspondingly large multipath effects in this case, too. Further work is now in progress at the Signal Corps Engineering Laboratories which will further clarify the problems involved in radio "moonshining."

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APPENDIX

FORMULAS FOR THE CHARACTERISTICS OF ELLIPTICAL GROUND-REFLECTION FRESNEL ZONES ON A PLANE EARTH

In Figs. 4 and 7 the ray reflections are shown as though they occurred at a point. Actually the entire surface of the earth is illuminated and, in accordance with Huygens' principle, re-radiates elementary waves in all directions. In any particular direction (in the present case in the direction of the target) all of these elementary waves reflected from a smooth earth arrive with intensities and phase relations such that they very nearly cancel each other, leaving only the waves from a small elliptical zone in the neighborhood of the ray reflection point as determined by the laws of geometrical optics. This small reflecting zone may be called a Fresnel zone, since it is closely related to the Fresnel zones of diffraction theory;[43] thus the length of the reflected ray path at the edge of the first Fresnel zone is one-half wavelength greater than the geometrical ray path and, more generally, the length of the reflected ray path at the outer edge of the nth Fresnel zone is n half-wavelengths greater than the geometrical ray path.

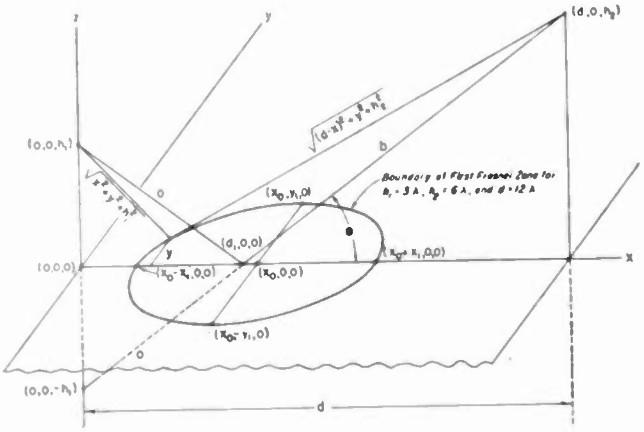

Fig. 10—Geometry of Fresnel zones on a plane earth.

Fig. 10 shows the geometry of the problem. A rectangular co-ordinate system (x, y, z) is chosen with the xy plane representing the plane of the earth and the origin of co-ordinates on the ground immediately below the transmitting antenna. Thus, the transmitting antenna at a height h_1 above the ground has the co-ordinates $(0, 0, h_1)$, while the receiving antenna, at a height h_2 above the ground and at a distance $x=d$, has the co-ordinates $(d, 0, h_2)$. The co-ordinates of the geometrical ground-reflection point are $(d_1, 0, 0)$. The length $(a+b)$ of the geometrical ray path will be called $R\equiv(a+b)$. By definition, the length of the propagation path via some other reflection point $(x, y, 0)$ lying on the boundary of the nth Fresnel zone will be equal to $(R+n\lambda/2)\equiv R_n$. The elementary waves scattered from points within the first Fresnel zone $n=1$ all have phases such that they increase the total field intensity at the target; elementary waves scattered from points within the second zone $n=2$ are out of phase with those from the first zone, etc., alternate zones adding and subtracting from the total field at the target. Fig. 11 shows the first five Fresnel zones for the geometrical arrangement of antennas shown in Fig. 10. We see by Fig. 10 that R_n may be expressed in terms of x, y, h_1, h_2, and d as follows:

$$R_n = \sqrt{x^2 + y^2 + h_1^2} + \sqrt{(d - x)^2 + y^2 + h_2^2}. \qquad (54)$$

If we subtract $\sqrt{x^2+y^2+h_1^2}$ from both sides of (56) and then square both sides, we obtain

$$2R_n\sqrt{x^2 + y^2 + h_1^2} = R_n^2 - d^2 + h_1^2 - h_2^2 + 2dx. \qquad (55)$$

We will write $R_n^2-d^2+h_1^2-h_2^2=K^2$ and square both sides of (55), obtaining

$$Ax^2 - 2Bx + C + 4R_n^2y^2 = 0 \qquad (56)$$

in which $A=4(R_n^2-d^2)$, $B=2K^2d$, $C=(4R_n^2h_1^2-K^4)$. Since (56) defines a second-degree curve on the x, y plane, and since physical considerations indicate that it is a closed curve, it must be an ellipse. If we let x_0 denote the center of the ellipse (which is, in general, not at the geometrical reflection point $(d_1, 0, 0,))$ and let x_1 denote the length of the semi-major axis of the ellipse, we may determine the points (x_0-x_1) and (x_0+x_1) at which the ellipse crosses the x axis by setting $y=0$ in (56) and solving the remainder as a quadratic in x; thus

$$x_0 - x_1 = \frac{B - \sqrt{B^2 - AC}}{A} \qquad (57)$$

$$x_0 + x_1 = \frac{B + \sqrt{B^2 - AC}}{A}. \qquad (58)$$

Thus x_0, x_1 are half of the sum and difference, respectively, of (57) and (58):

$$x_0 = \frac{B}{A} = d_1\left[1 + \frac{(h_2 - h_1)}{2h_1[1 + (h_1 + h_2)^2/n\lambda(R + n\lambda/4)]}\right] \qquad (59)$$

$$x_1 = \frac{\sqrt{B_2 - AC}}{A} = \frac{\left(1 + \dfrac{n\lambda}{2R}\right)\left(1 + \dfrac{n\lambda}{4R}\right)}{\sin\theta\left[1 + \dfrac{n\lambda\left(R + \dfrac{n\lambda}{4}\right)}{(h_1 + h_2)^2}\right]}\sqrt{\frac{abn\lambda}{a + b + \dfrac{n\lambda}{4}}\left[1 + \frac{n\lambda\left(R + \dfrac{n\lambda}{4}\right)}{4h_1h_2}\right]}. \qquad (60)$$

The above equation of x_0 determines the centers of the Fresnel zones in terms of the location, $d_1=dh_1/(h_1+h_2)$,

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[43] C. F. Meyer, "The diffraction of light, x-rays, and material particles," University of Chicago Press, Chicago, Illinois.

of the geometrical reflection point. When $h_2 = h_1$ all of the elliptical Fresnel zones have coincident centers at the geometrical reflection point $d_1 = d/2$. When h_2 is greater or less than h_1, x_0 will be greater or less, respectively, than d_1. The centers of the successive Fresnel zones are not coincident but lie progressively further from the geometrical reflection point d_1 and approach the midpoint $d/2$ in the limit as the Fresnel zone number n becomes very large.

The above equation for x_1 is an exact expression for the major semi-axis of the ellipse. The positions of the two ends of the ellipse along the x axis are obtained by subtracting and adding x_1 and x_0.

To determine the minor semi-axis of the ellipse y_1 we may set $x = x_0$ in (56) and solve for y:

$$y_1 = \left(1 + \frac{n\lambda}{4R}\right) \sqrt{\frac{abn\lambda}{a + b + \frac{n\lambda}{4}} \left[1 + \frac{n\lambda\left(R + \frac{n\lambda}{4}\right)}{4h_1h_2}\right] \bigg/ \left[1 + \frac{n\lambda\left(R + \frac{n\lambda}{n}\right)}{(h_1 + h_2)^2}\right]}. \tag{61}$$

In most radar applications both h_2 and R are usually very much larger than λ and h_1, and if we limit our considerations to the first Fresnel zone, we obtain

$$x_0 \cong d_1\left[1 + \frac{\lambda}{2h_1 \sin\theta}\right] \qquad \left\{\begin{matrix} h_2 \gg \lambda \\ h_2 \gg h_1 \end{matrix}\right\} \tag{62}$$

$$x_1 \cong \sqrt{\frac{h_1\lambda}{\sin^3\theta}\left[1 + \frac{\lambda}{4h_1 \sin\theta}\right]} \qquad \left\{\begin{matrix} h_2 \gg \lambda \\ h_2 \gg h_1 \end{matrix}\right\} \tag{63}$$

$$y_1 \cong x_1 \sin\theta \quad (h_2 \gg \lambda \text{ and } h_2 \gg h_1). \tag{64}$$

An important application of the above formulas is the determination of the location of the first Fresnel zone for transmission at the angle θ corresponding to the maximum of the first ground-reflection lobe. By (34) we see that $\sin\theta = \lambda/4h_1$, and when this is substituted in the above (noting that $d_1 \cong h_1/\sin\theta$), we obtain

$$x_0 \cong 12h_1^2/\lambda \qquad (h_2 \gg \lambda) \tag{65}$$

$$x_1 \cong 8\sqrt{2}h_1^2/\lambda \qquad (h_2 \gg \lambda;\ h_2 \gg h_1) \tag{66}$$

$$y_1 \cong 2\sqrt{2}h_1 \qquad (h_2 \gg \lambda \text{ and } h_2 \gg h_1). \tag{67}$$

Combining (65) and (66) we find that the distances from the transmitter to the nearest d_n and furthest d_f points on the first Fresnel zone for transmission at an angle corresponding to the maximum of the first lobe are equal to

$$d_n = x_0 - x_1 \cong 0.688h_1^2/\lambda \quad (h_2 \gg \lambda;\ h_2 \gg h_1) \tag{68}$$

$$d_f = x_0 + x_1 \cong 23.3h_1^2/\lambda \quad (h_2 \gg \lambda;\ h_2 \gg h_1). \tag{69}$$

We may apply (69) to the problem of extending the maximum range by the use of a sloping site, as shown on Fig. 7. It is desirable that the slope support at least one complete Fresnel zone if it is to contribute substantially to the distant field, and thus the distance from the radar antenna to the end of the slope should be greater than or at least equal to d_f as given by (69). For a distant target on the horizon $\theta' = \phi$, and upon substituting the value $h_1 = \lambda/4 \sin\theta$ corresponding to the lobe maximum in (69), we obtain

$$d_f = 1.46\lambda/\sin^2\phi \quad (\theta' = \phi). \tag{70}$$

If the radar antenna is placed back along the slope at a distance greater than that given by (70), then the slope will support at least one complete Fresnel zone for radiation in a horizontal direction and the maximum range in this direction may be very nearly doubled.

Other applications of the above Fresnel-zone theory arise in connection with the problem of estimating the size and location of the areas in which the ground reflection takes place. This problem is important in connection with siting radar sets or other types of radio direction finders. For a well-developed ground-reflection maximum lobe, the ground should be flat over an area which includes, as a minimum, that within the first Fresnel zone. The degree of flatness required for a radio reflection depends upon the wavelength and the angle of incidence; assuming that phase changes corresponding to path-length differences less than $\lambda/16$ are of negligible importance, then, according to Rayleigh's limit, height deviations in terrain from a smooth surface must have a magnitude less than $\Delta h = \lambda/16 \sin\theta$ throughout the area of the first Fresnel zone for waves at an angle θ. Substituting (34) in the above, we find that the permissible height deviation to permit a well-developed kth lobe is $\Delta h = h_1/4(2k-1)$ for a transmitting antenna at a height h_1; it is evident that Δh is independent of the frequency in this case and, for the first few lobes, rather large irregularities in terrain are allowable within the first Fresnel zone provided h_1 is large.

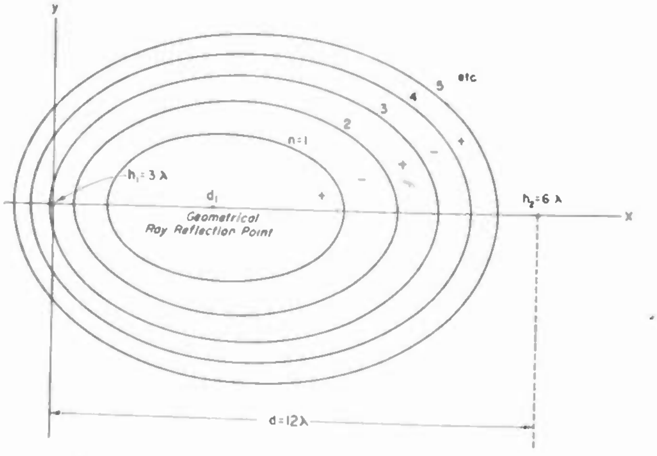

Fig. 11—Plan view of first five Fresnel zones on the ground for the antenna arrangement of Fig. 10.

A New System of Frequency Modulation*

ROBERT ADLER†, ASSOCIATE, I.R.E.

Summary—**The development of a new phase-modulator tube is described. In a concentric structure of conventional dimensions, a radial electron stream is shaped into a wave-like pattern which progresses continuously around the cathode. Phase shift, 12 to 16 times larger than in conventional modulators, is produced by electromagnetic deflection.**

The failings of early models are analyzed and the steps are reviewed which resulted in the development of a satisfactory tube structure.

THE PHASITRON system of frequency modulation, which has recently come into practical use, is based on a new tube structure. It is the object of this paper to review the development of this structure from its initial concept to the point where all technical requirements for high-fidelity broadcasting as well as for communication transmitters were met by laboratory models. The step from the laboratory model to the production model and the design of complete high-fidelity transmitters incorporating the new modulator have already been described in the literature.[1]

Two different types of frequency-modulation transmitters have heretofore been widely used, based on reactance modulators and phase modulators. In transmitters using a reactance modulator, devices are required for continuous automatic correction of the average carrier frequency with respect to a crystal oscillator. The frequency actually transmitted is then a function of the correction-system parameters and is not solely dependent on the crystal oscillator. Transmitters employing conventional phase modulators use a total frequency multiplication of several thousand times to obtain full rated deviation at the lowest audio frequencies. This high factor of multiplication requires two separate multiplier chains with frequency conversion between them, involving large numbers of tubes and circuits which present a number of problems with respect to random noise and spurious beat notes.

While none of these difficulties could be regarded as insurmountable, it appeared that crystal-controlled frequency-modulation transmitters could be simplified and perhaps improved in some respects if a phase modulator were known which would produce phase excursions far in excess of those obtainable with conventional modulators.

This thought stood at the beginning of the development. Analysis of an idea by C. W. Carnahan, who had proposed a mechanical phase modulator, indicated that inertia forces would make such a device impractical if not impossible; it appeared that only a vacuum tube could do the job. To operate efficiently in connection with conventional circuits and loads, such a tube would have to work with anode voltages and currents comparable to common receiver tubes. It seemed that this could be most easily accomplished if the usual concentric arrangement of anode and cathode were retained. With these rather broad directions in mind, let us now study the fundamental concept of the new phase modulator.

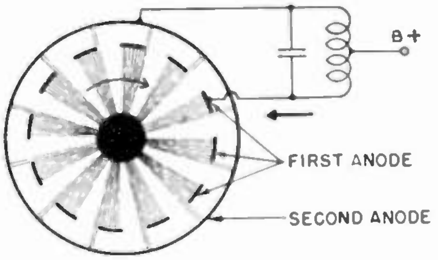

Fig. 1—Original concept of the phasitron modulator: a rotating bundle of electron beams.

Let us assume that by some device in the black central portion of Fig. 1 the electrons which fly radially away from the cathode are split into discrete radial beams—twelve are shown in this figure—and that the first anode consists of twelve bars, all connected. Let us further assume that the device in the center rotates with uniform speed, so that the electron beams alternately hit the twelve bars which constitute the first anode, or pass between them to strike the second anode which is a full cylinder. The tuned circuit connected from plate to plate is then excited at a constant frequency which depends on the speed of rotation of the device in the center.

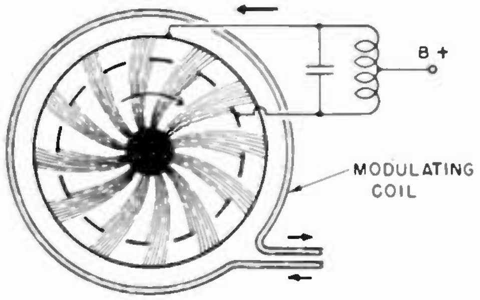

Fig. 2—Phase advanced 180 degrees by axial magnetic field.

Now let us apply a magnet field parallel to the cylinder axis. We can produce it by a concentric coil wound around the tube, as shown in Fig. 2. All electron beams are now deflected clockwise and thus advanced in the direction of rotation, so that they strike all plate bars ahead of schedule. In Fig. 2 the deflection has been

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† Zenith Radio Corporation, Chicago, Illinois.

[1] F. M. Bailey and H. P. Thomas, "Phasitron FM transmitter," *Electronics*, vol. 19, pp. 108–112; October, 1946.

made equal to the width of one bar, so that the plate currents which drive the push-pull circuit are advanced in phase by 180 degrees. If the magnet field were reversed the deflection would be counterclockwise and would retard the arrival of the beams at the anodes. The amount of phase shift obtained for a given plate voltage depends only on the magnetic field intensity.

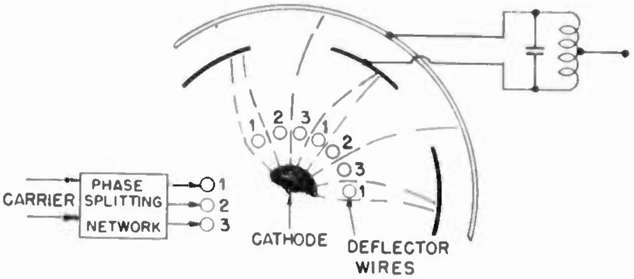

Fig. 3—Generation of rotating electron beams by a deflector structure carrying three-phase potentials.

Let us now turn to the device in the center which generates the rotating beams. Fig. 3 shows schematically a sector of the first tube structure actually built. A cylindrical cathode is surrounded by a number of parallel wires of smaller diameter, arranged on a cylindrical surface around it. These wires are consecutively numbered 1, 2, 3, 1, 2, 3, etc.; all "1's" are interconnected, as are all 2's and 3's. The three groups are fed from a three-phase network which is driven by a crystal oscillator. Fig. 3 shows conditions at an instant when 1 is on its positive peak while 2 and 3 are negative; the electrons passing around rods labeled "1" are focused together and deflected away from rods 2 and 3. Most of the current, therefore, flows to the first anode. One third of a cycle later, rods 2 will be at their positive peak while 3 and 1 will be negative, and the regions where most electrons strike will have traveled clockwise by a corresponding angle; so, as the potentials travel clockwise from wire to wire, we obtain the equivalent of a rotating system of beams.

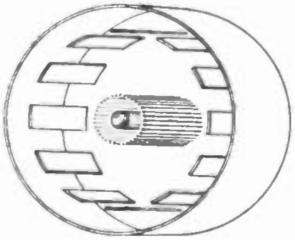

Fig. 4—Electrode structure of first operating model.

The rotating structure of electrostatic fields which produces the electron beams might well be compared to the rotating magnetic field generated by the stator of a slow-running three-phase motor. Such a stator has 3 N poles, N being the number of cycles for one complete revolution of the magnetic field.

Fig. 4 shows, schematically, the first tube actually built. Its first anode had 12 bars, and the three-phase deflector correspondingly consisted of three groups of 12 wires each, or 36 in all. They were operated at a low positive potential with the three-phase excitation superimposed. Fig. 5 shows the first model, with the three-phase connections led out at the top. The tube was first tried out with a carrier frequency of 60 cycles. The push-pull output circuit was connected through a transformer to the oscilloscope and the sweep synchronized from one terminal of the three-phase input. A coil was arranged around the tube to produce the axial magnet field; it was fed from a variable direct-current supply so that the phase shift could be accurately plotted as a function of current through the coil. Plus and minus 720 degrees—or plus and minus two complete cycles—were obtained with this first tube.

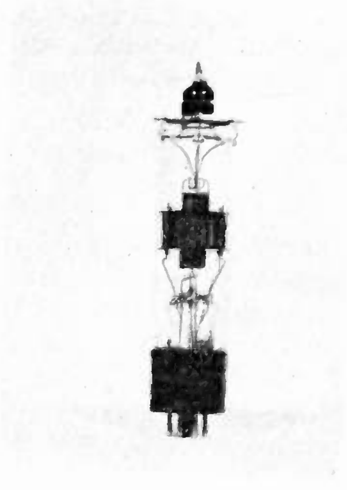

Fig. 5—First operating model: connections to three-phase deflector elements appear on top.

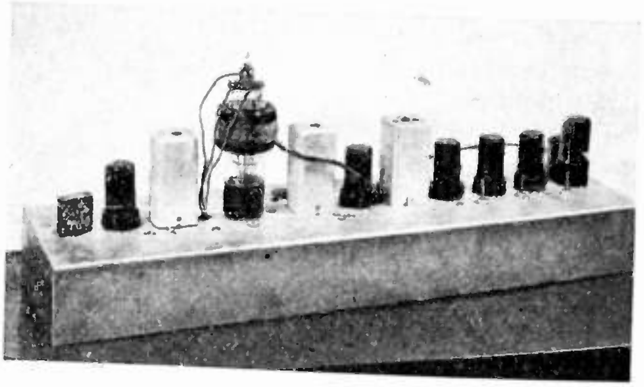

Fig. 6—Frequency-modulation broadcast transmitter, showing (left to right): 235-kilocycle crystal, oscillator tube, three-phase network, modulator tube with audio deflection coil, push-pull output transformer, first doubler, frequency multipliers to 45.1 megacycles.

Next, a complete frequency-modulation transmitter was built (see Fig. 6), which consisted of a crystal oscillator on 235 kilocycles, a simple phase-splitting network, the modulator tube, and a string of multipliers to reach the final frequency of 45.1 megacycles. The total multiplication was only 192 times.

Using the full 720 degrees phase excursion, this transmitter was capable of 75 kilocycles deviation at an audio frequency of 30 cycles. A similar unit, using the same multiplication, was built to operate as exciter for Zenith's 50,000-watt station, WWZR, and several successful test transmissions were made over this station.

Only a fraction of one watt of audio power was needed for the modulating coil. The inductance of this coil is instrumental in obtaining constant frequency deviation over the entire audio band; with a constant voltage applied to the coil, this inductance tends to make the current, and consequently the phase shift, inversely proportional to the frequency of the audio signal. This is exactly the characteristic required to obtain constant deviation.

The first model of the new modulator tube had several failings. Most apparent was the low signal output: with one milliampere direct current on each anode, only about fifty microamperes of signal current was obtained. This low current efficiency caused a considerable amount of hiss; the noise level was 58 decibels below full deviation, instead of 70 or more.

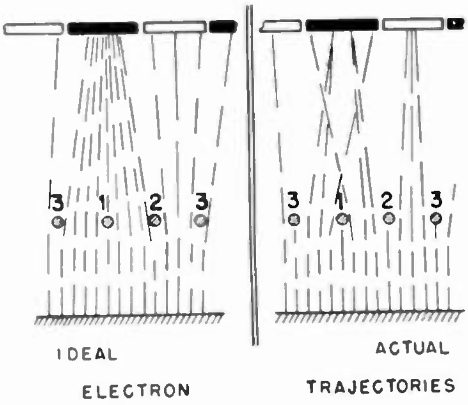

Fig. 7—Comparison of desirable electron trajectories with those probably produced in first model.

The low current efficiency in this first tube was probably caused by the fact that the action of its three-phase deflector electrode was far from ideal, even in theory. This is illustrated in Fig. 7. The left half of this figure shows the electron trajectories in that form which would have been most desirable. The right half of the figure gives a greatly simplified and schematized idea of the actual trajectories. The strong aberrations which occurred close to the deflector wires nearly destroyed the desired focusing action and reduced the current efficiency considerably.

The same strong fields close to the deflector wires which produced defocusing also gave rise to a second undesirable effect: when a gradually increasing magnetic field was applied to the tube, the phase shift did not rise strictly in proportion but showed irregular deviations from the desired linear relation. Correspondingly, a small incremental field produced varying amounts of phase shift at different points of the field-versus-phase characteristic. At low audio frequencies where large phase excursions are required considerable distortion was caused by these irregularities.

The observed type of distortion can be explained if the tangential velocity of the electron beams is assumed to vary from point to point along the periphery of the first anode; for the electrical phase shift produced by a small incremental field is proportional to the time interval in which the beams, through their rotary motion, could cover a portion of the periphery equal to that over which they are deflected by the incremental field. Because it makes no difference which part of a given anode bar an electron strikes, the peripheral velocity matters only in the regions of current transition—at the 24 edges of the 12 first-anode bars.

If we assume an irregular distribution of tangential velocities around the periphery, the average of the velocities measured at the 24 edges may well be different from the average taken over the whole periphery; then, if we rotate the entire system of velocities by applying an axial magnetic field, those velocities appearing at the 24 edges—and consequently their average—are bound to vary. In a position where this average is high, a larger incremental field is required to produce a given increment of phase.

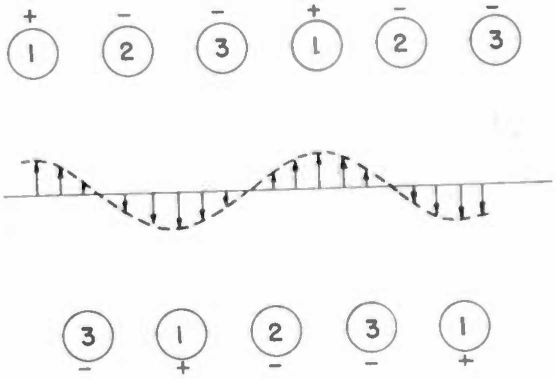

Fig. 8—Schematic view of transverse deflector arrangement.

There can be little doubt regarding the existence of an irregular distribution of tangential velocities in the tube structure described. Cyclic variations (one cycle for each deflector wire) were bound to be present, and because there were 1½ wires for every edge, all even harmonics of such cyclic variations would appear in phase on all edges. More detailed study of the distortion effect tended to verify the above analysis. In the following we shall refer to this effect as "structural distortion."

Two more tubes were built which differed from the first only in detail. Their performance was quite similar.

If the new phase modulator was to meet practical requirements it seemed that the signal output would have to be at least five times higher for the same direct current on the plates, and the structural distortion would have to be substantially suppressed. An entirely different deflector system was needed which would produce a smoothly rotating electrostatic field without noticeable speed variations from point to point.

Fig. 8 illustrates the principle of a new deflector system designed to produce such a field. Two rows of short parallel wires face each other across a center plane. Each row consists of three groups which are interlaced and connected to a three-phase supply as before. The direction in which the voltages travel is the same in both rows; but the wires "1" in one row face the middle between the wires "2" and "3" in the other. A transverse electric field (vertical in Fig. 8) is produced in the center plane between the two rows, and if we plot this field at a given instant from left to right along the center plane we find that it varies quite smoothly, very much like a sine function. Because the individual deflector wires are not in the immediate vicinity of the center plane, their strong local fields can no longer upset the smooth character of the field distribution.

To save time-consuming computation of the field along the center plane, an electrolytic model was built as shown in Fig. 9. On each of two bakelite bars, three bare wires were wound interlaced. Three stationary voltages, corresponding to an instantaneous condition of a three-phase system, were applied to the wires on each bar. The short portions of these wires which faced each other across the gap represented the deflector elements shown schematically in Fig. 8. A two-element probe was moved along the gap between the bars by means of a lead screw. The two elements were positioned symmetrically on a line perpendicular to the center plane, and the potential of each element was plotted with reference to the three-phase neutral. Fig. 10 shows one of these plots, which not only proved that a field distribution closely approaching a sine function could be produced but also yielded data on the intensity of the transverse field as a function of configuration.

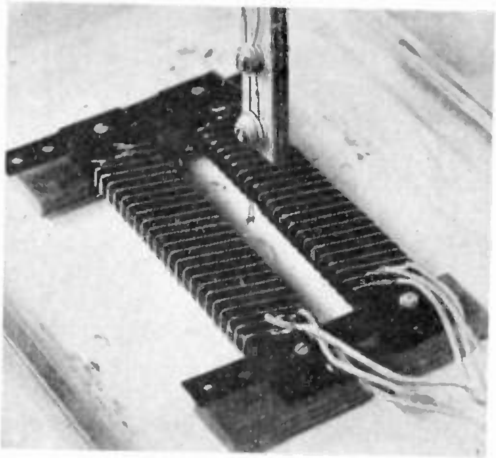

Fig. 9—Electrolytic model for the study of transverse deflectors.

To utilize a deflection system of this new kind in a concentric tube structure, the two parallel rows of short wires had to be bent so as to form two parallel coaxial circles, and the electron flow had to be so arranged that all electrons would travel radially and pass near the center plane between the two deflectors. This led to the concept illustrated in Fig. 11. The electrons are shown here in the shape of a sharp-edged disk from which one quarter has been cut away so that the lower deflector becomes partly visible. The electrons fly along radial lines until they pass between the two deflectors; here they are deflected upward or downward according to the wave-shaped transverse field; and as this field travels around the cathode, so do the waves in the electron disk.

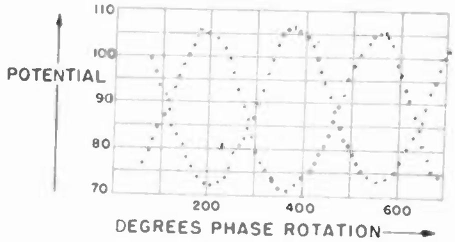

Fig. 10—Potential plot obtained with the electrolytic model shown in Fig. 9.

It is essential for the proper function of this deflector system that most electrons pass near the center plane and that the deflector wires with their strong local fields remain substantially outside the disk-shaped electron stream.

The black and white fields marked *A* and *B* represent portions of the first and second anode, respectively. The white fields correspond to windows in the first anode, and in the position shown in Fig. 11 there is one window visible above each valley in the electron disk, and another window hidden below each peak. As shown more clearly in the schematic sector on the lower right of the figure, the thin edge of the ruffled disk strikes only full portions of the first anode at this moment, and no current flows through the windows to the second anode. If a modulating magnet field of proper

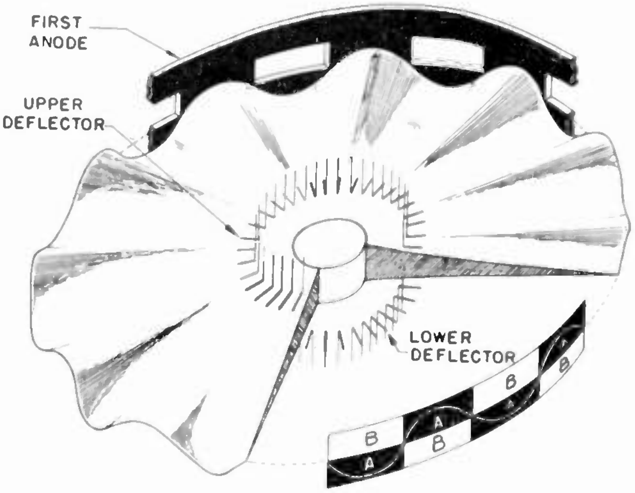

Fig. 11—Transverse deflection produced in a disk formed by electrons traveling radially.

size is applied at this instant, the picture changes as shown in Fig. 12; all current now flows to the second anode and the phase is advanced 180 degrees.

So far, the last two figures merely represent a mental concept. To make the electrons actually travel along such trajectories that they pass near the center plane and come to a line focus on the periphery, we must add an electron-optical system. The problem is not too difficult because accurate focusing as in a cathode-ray tube is not needed. Fundamentally, we need an electron gun developed into a full circle. Again, the electrolytic-model method was used to arrive at a useful arrangement. Because only radial and axial fields exist in a concentric structure, a thin wedge-shaped sector will suffice for a model. This has been well-known for some time.[2]

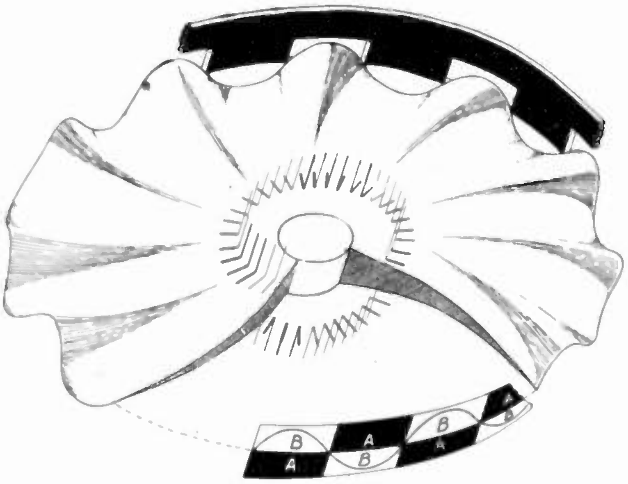

Fig. 12—Electron-beam configuration in the presence of an axial magnet field.

Fig. 13 shows the potential distribution plotted with such a wedge model. The deflectors are here regarded as solid surfaces. The anode (not visible in this figure) would be somewhat farther to the right. Two electron trajectories are drawn in to show that they have the desired character; their shape was found by geometrical construction. A number of such plots were made, varying the ratio of potentials on first focus, deflector electrodes, and anode. It was found that the deflectors had to be longer in the direction of electron travel than had been anticipated; but instead of making the little wires longer, a solid ring was laid around each deflector electrode and the two rings brought out to a separate contact.

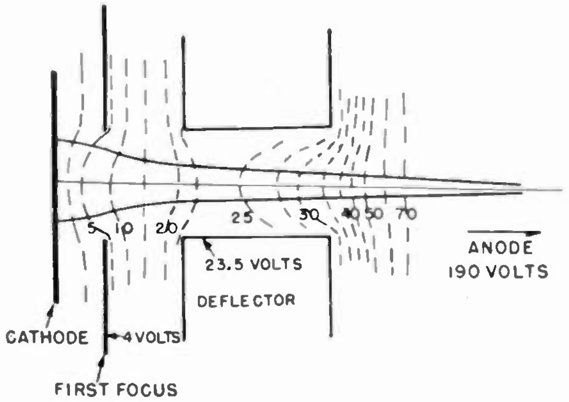

Fig. 13—Potential plot obtained with electrolytic wedge model of a tube structure intended to produce the configuration shown in Fig. 11.

Fig. 14 shows the tube layout which resulted from these plots. Around the cathode there are two cylinders—the first focusing electrode. Next come the two deflectors which carry the short radial wires; they are surrounded by the two rings just mentioned, called the second focusing electrode. These rings are made of soft iron and continued outside the supporting micas by more soft iron; they help to concentrate the modulating magnet field into a narrow region, as indicated by the dotted lines. Finally, the anodes *A* and *B* terminate the concentric structure.

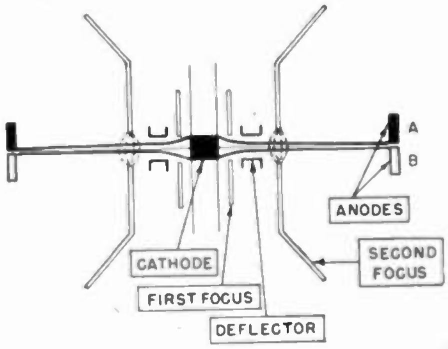

Fig. 14—Section of improved tube structure. Modulating flux is concentrated between edges of second focusing electrodes.

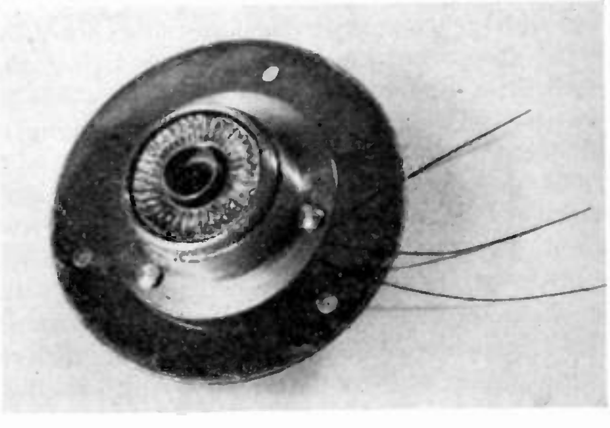

Fig. 15—Three-phase deflector electrode, mounted together with first and second focusing electrodes. Two such assemblies faced each other in the completed tube.

Fig. 15 shows one of the deflectors with the 36 short radial wires and the first and second focusing electrodes mounted on one mica.

In the first tube of this new kind, the checkerboard pattern of anode fields *A* and *B* was made up of 24

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[2] M. Bowman-Manifold and F. H. Nicoll, "Electrolytic-field-plotting trough for circularly symmetric systems," *Nature*, vol. 140, p. 39, July, 1938.

rectangular metal fins, 12 to each anode, all lying in one cylindrical plane and properly interconnected.

Results with the new model were most encouraging. With careful adjustment of the focusing potentials the structural distortion could be almost entirely eliminated; current efficiency was improved by a factor of three or four, and the hiss level was reduced by a much larger factor, probably because the electron stream was no longer intercepted by a positive grid. With a total multiplication of 192 times, the hiss was now more than 70 decibels below full deviation.

The new modulator operated with very low distortion down to about 100 cycles. But at the very large phase excursions required for 50 cycles at this low multiplication, nonlinear distortion appeared. It was mostly of the third-harmonic type, quite similar to that which conventional phase modulators produce when they are used over too wide an angle. Fig. 16 shows the geometrical reasons for this type of distortion; for large deflections, the arc which the electron trajectories cover on the periphery of the tube increases faster than the angle of deflection, so that the graph of phase shift versus magnet field becomes steeper in the regions of large phase shift. The figure also shows that it is better to concentrate the magnet field close to the center, rather than spread it out over the whole tube. It was for this reason that the second focusing electrodes were made of soft iron.

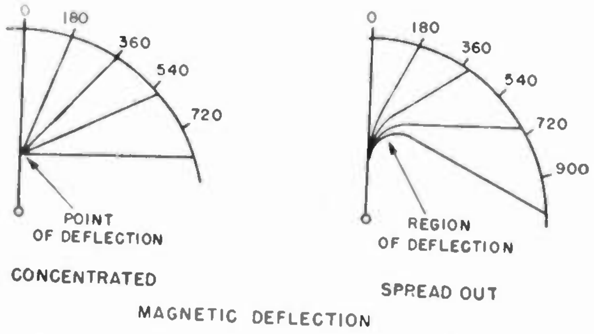

Fig. 16—Nonlinearity at large phase excursions. Distortion is reduced by concentrating the region where deflection occurs.

To suppress the nonlinearity just explained, a number of methods were open. The most straightforward solution would consist in introducing a controlled amount of saturation into the magnetic path so that strong fields would be reduced by just the right amount. Another solution consists in driving the deflection coil from a pair of pentodes in push-pull, so adjusted that their combined transconductance is peaked at the operating point. With a modulator built according to this idea, total distortion at 50 cycles and full deviation could be reduced to 1.35 per cent with a multiplication of only 192 times. The corresponding phase excursion was plus and minus 430 degrees.

In the further course of the development, the hiss level was reduced to such an extent that, at the present time, it is considered more convenient to double the multiplication and dispense entirely with any artificial means of linearization.

A few words about the circuits used to produce the three-phase excitation might be in order at this point. Originally, the simple resistance-capacitance network on the right in Fig. 17 had been used in all experiments. Later it was found that good wave form on the deflector electrodes was important for reducing distortion to a minimum, and the network on the left, derived from the well-known Scott circuit, was substituted.

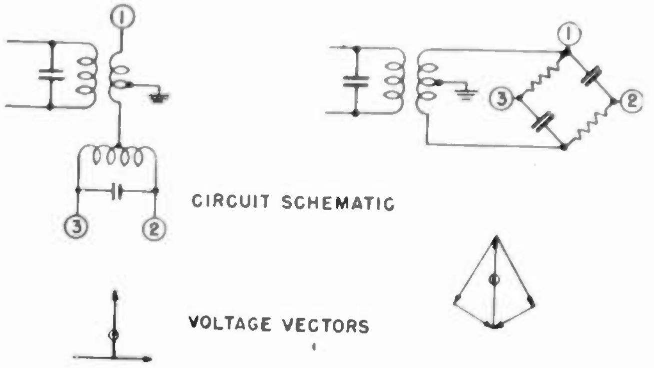

Fig. 17—Networks for generating three-phase potentials.

The tube described above might have been used in frequency-modulation broadcast transmitters, but its output level was still inconveniently low, and certainly too low for use in communication transmitters. Checking into the causes for the low output, it was found that there existed a resistance of only about 10,000 ohms between the two anodes, evidently as a consequence of secondary emission. Such secondary emission would reduce the radio-frequency component of the anode current, even without load, because the anode struck by more electrons would also emit more secondaries. In addition, only low-impedance load circuits could be used. In the first tube of the new kind the anode fields *A* and *B* were arranged side by side, and a direct-current voltage applied between them would only draw all current to the more positive anode. A different anode structure appeared necessary to suppress secondary-emission effects.

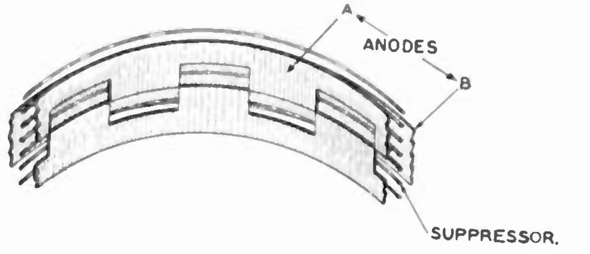

Fig. 18—Arrangement of first anode, suppressor, and second anode in last experimental model.

A tube was then built with an anode structure as shown in Fig. 18. This, in effect, is a return to the original model in which holes in the first anode represented the active parts of the second; but this time, a helical suppressor was set between the two anodes.

Fig. 19 shows an X-ray photograph of this last experimental model. The two deflectors with their little wires, the outer focusing rings with their magnetic extensions, the checkerboard fields of the first anode, and finally the suppressor and the second anode appear in this picture. The current efficiency of this tube was several times that of the previous model; the signal component was equal to more than half the direct current on each plate. The output impedance was several megohms. The hiss level, with a multiplication of 192 times, could not be measured accurately; with twice that multiplication, it was found to be 78 decibels below full deviation.

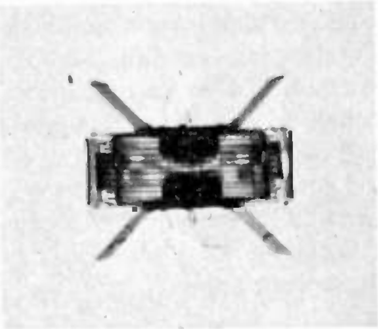

Fig. 19—X-ray view of last experimental model. Electrodes inside the suppressor helix may be identified by comparison with Fig. 14.

The large output from this tube made it possible to use it in a communication transmitter (see Fig. 20). Here, a crystal oscillator was used to drive a three-phase network at 8.37 megacycles. In the output from the phase-modulator tube, a high-inductance permeability-tuned load circuit delivered 30 volts into the grid of a 6AC7 tube operating as quadrupler. More than sufficient power was available in the quadrupler output to drive an 815 dual beam tetrode at 33.5 megacycles. The whole transmitter had only four tubes.

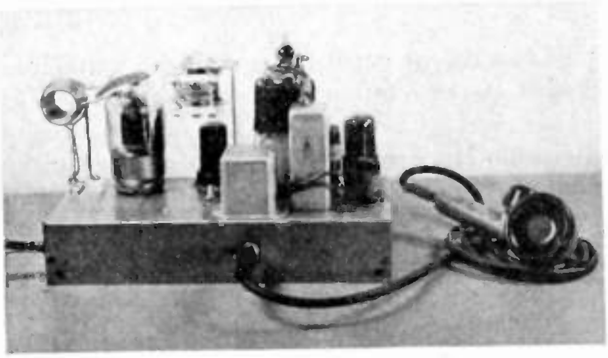

Fig. 20—33-megacycle communication transmitter using experimental phasitron tube.

At this point, it seemed that the project had grown beyond the purely experimental stage. Its further development was then taken over by F. M. Bailey and H. P. Thomas, of the General Electric Company.

Acknowledgment

The author wishes to thank E. C. Ewing, of the Zenith Radio Corporation, for his assistance in making the experimental tubes, and W. E. Phillips, of the Raytheon Manufacturing Co. (formerly of Zenith), whose initiative and co-operation contributed greatly to the success of this project.

A Note on Noise and Conversion-Gain Measurements*

WILLIAM M. BREAZEALE†, SENIOR MEMBER, I.R.E.

Summary—**The development of microwave receivers with a low-gain converter as the first stage has made it desirable that the noise level and the conversion gain be determined independently of the following intermediate-frequency amplifier. Often this converter is a crystal with a conversion gain less than one. This paper discusses some of the general procedures that have been used to measure microwave-converter noise levels and conversion gains.**

Introduction

Many papers on the subject of the limitation of receiver sensitivity by noise have been published in the past twenty years. In an article published in 1942, North[1] proposed that receiver noise be expressed in terms of a certain rating number, namely, the ratio of total noise-power input to the detector of a receiver with a dummy antenna to that portion of the noise power which originated as thermal agitations in the antenna itself. He showed how the number could be measured and emphasized its significance as a just basis for the comparison of receivers for a given service. The number, which is often called the "noise figure," has been widely adopted for this service. Further discussion of receiver noise and a bibliography can be found in an article by Herold.[2]

The papers published before 1944 in general dealt with the over-all noise figure of the receiver. However, the development in the last few years of microwave receivers with a low-gain converter as the initial stage has

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† Department of Engineering, University of Virginia, Charlottesville, Virginia.

[1] D. O. North, "The absolute sensitivity of radio receivers," *RCA Rev.*, vol. 6, pp. 332–343; January, 1942.

[2] E. W. Herold, "An analysis of the signal-to-noise ratio of ultra-high frequency receivers," *R.C.A. Rev.*, vol. 6, pp. 302–331; January, 1942.

made the measurement of the noise figure of the converter alone a matter of considerable importance. Generally, the converter is a crystal with a conversion gain less than unity but with a relatively low inherent noise level.

Friis[3] in a recent paper has supplied a mathematical discussion of the relationship between the noise figure of the receiver as a whole and the noise figures of its components. His discussion is based on a definition of the noise figure F of a network given by the relation

$$F = \frac{N}{kTB} \times \frac{S_g}{S} = \frac{N}{kTB} \times \frac{1}{g} \tag{1}$$

where N is the available noise power at the output terminals of the network, k is Boltzmann's constant $=1.37\times10^{-23}$ joule per degree Kelvin, T is the absolute temperature of the network, B is the effective noise bandwidth, S_g is the available power at the output terminals of the signal generator, S is the available signal power at the output terminals of the network, and g is the power gain. A detailed discussion of these terms is found in his paper. It can be seen from (1) that a "perfect" receiver, i.e., one which has a noise figure of unity or 0 decibels, would have an equivalent input noise level of kTB watts.

Continuing, Friis derives the following equations for the noise output and over-all noise figure of a system with two networks in cascade

$$N_{(1+2)} = F_1 \times kTB \times g_1g_2 + (F_2 - 1) \times kTB \times g_2, \tag{2}$$

$$F_{(1+2)} = F_1 + (F_2 - 1)\frac{1}{g_1}\cdot \tag{3}$$

In the above, $N_{(1+2)}$ is the noise output in watts of the two networks, $F_{(1+2)}$ is the over-all noise figure of the system, and the subscripts 1 and 2 denote the individual quantities of the first and second network. In this paper all powers are expressed in watts and T is always considered as equal to "room temperature," 290 degrees Kelvin.

It can be easily seen that when g_1 is less than unity, as is the case when a crystal converter is used as the initial stage, the over-all noise figure of a receiver depends strongly on the noise figure of the intermediate-frequency amplifier.

From the above relations it is obvious that if two of the three quantities (noise, gain, or noise figure) are known, the third can be calculated. This paper describes some of the procedures which have been used to determine noise and conversion gain, particularly with reference to microwave receivers. A better understanding of the converter's characteristics is given by these two quantities than by the value of the noise figure alone. The converter is usually a crystal, and references to a crystal in this paper generally mean a crystal converter.

I. Noise Measurements

The noise output of a crystal (or other noise source) when excited by a local oscillator only may be fixed in several ways, all of which involve amplifying this noise. For purposes of deriving certain relations, consider that the crystal is connected to an intermediate-frequency amplifier through a variable attenuator of the same iterative impedance. The noise output of the crystal can be divided into thermal-agitation noise and noise in excess of thermal-agitation noise. Denote this excess noise by N_e. Then the noise available to the amplifier with the attenuator on zero is

$$N = N_e + kTB. \tag{4}$$

With attenuation A inserted between the crystal and the amplifier, the noise available to the amplifier is now

$$N' = \frac{N_e}{A} + kTB. \tag{5}$$

That (5) is true is obvious as soon as one remembers that the minimum possible noise output from any network is its thermal-agitation noise. Thus, the attenuator will itself contribute thermal-agitation noise, although it will attenuate any excess noise at the input.

The power outputs of the amplifier will not be proportional to N and N' unless the amplifier is perfectly noiseless. The noise due to the amplifier is introduced with the aid of (2). The first part of the equation covers the noise contributed by the first network. In the case of the crystal, the noise output of the amplifier becomes

$$N_{(1+2)} = (N_e + F_2 \times kTB)g_2 \tag{6}$$

where F_2 is the noise figure of the amplifier. Insertion of attenuation A as described above changes the amplifier output noise to a new value:

$$N'_{(1+2)} = \left(\frac{N_e}{A} + F_2 \times kTB\right)g_2. \tag{7}$$

The largest ratio of amplifier power outputs occurs when A is infinitely large, or equivalently, because N_e is then zero, when the crystal and attenuator are replaced by a resistor with an impedance equal to the output impedance of the attenuator. Equation (7) becomes

$$N'_{(1+2)} = F_2 \times kTB \times g_2 \tag{8}$$

which is (1) rearranged, a not very surprising result.

The ratio of output-power readings in this case becomes

$$\frac{N_{(1+2)}}{N'_{(1+2)}} = \frac{N_e + F_2 \times kTB}{F_2 \times kTB}. \tag{9}$$

From (9) we can determine N_e and consequently the total noise power output of the cystal circuit, which, as mentioned above, is N_e+kTB watts.

The output noise of the crystal is frequently expressed

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[3] H. T. Friis, "Noise figures of radio receivers," Proc. I.R.E., vol. 32, pp. 419–422; July, 1944. (See also the discussion in the February, 1945, issue of Proceedings, pp. 125–127.) At the end of his paper Friis describes briefly a method of measuring noise figures and noise.

as an equivalent temperature, i.e., the temperature necessary for a passive resistance of impedance equal to the crystal impedance to have the same noise power output

$$t_c = \frac{N_e + kTB}{kTB} \cdot \tag{10}$$

With this definition, the noise figure (4) of the crystal becomes

$$F_1 = t_c/g_1. \tag{11}$$

If a square-law detector (i.e., a thermocouple or bolometer) is used, the output-meter readings will be proportional to power. Hence, the ratio of the meter readings will be the ratio required in (9). On the other hand, an intermediate-frequency attenuator can be introduced after all important sources of noise and used to control the amplifier gain. Then the ratio (9) is determined from the intermediate-frequency attenuator settings required to give the same amplifier output from either the crystal or the equivalent resistor at the input. It is not necessary to know the response of the detector, nor does slight overloading of the portion of the amplifier after the attenuator vitiate results. (This statement would not hold if we were considering a mixture of sine signal and noise.) A block diagram of the equipment is shown in Fig. 1.

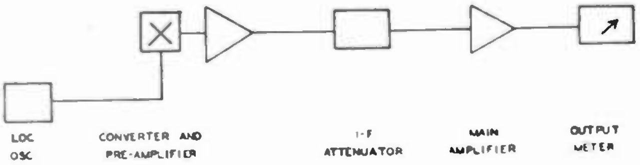

Fig. 1—Noise-measuring equipment.

It will be observed when using a tunable crystal converter that the intermediate-frequency impedance presented to the amplifier is a function of the radio-frequency tuning.[4] The radio-frequency tuning can be fixed with the aid of a standing-wave detector inserted between the signal generator and the mixer.

The above discussion presupposes that the bandwidth is determined after the first stages of amplification. In other words, the over-all bandwidth of the equipment is not a function of the crystal impedance. W. W. Hansen and E. L. Ginzton have described to the writer a scheme of building a circuit in which the bandwidth is a function only of the impedance of the crystal. This is accomplished by using a high-Q input circuit and an amplifier with a pass band sufficiently wide so that the over-all bandpass is determined by the input circuit.

The circuit is shown in Fig. 2. When a resistor is connected across the input, the total integrated mean-square thermal-agitation voltage appearing across the capacitor C (equipartition law) is

$$\overline{E_1^2} = \frac{kT}{C} \cdot \tag{12}$$

If a crystal is substituted for the resistor, the noise increases by the factor t_c, and the new noise voltage is

$$\overline{E_2^2} = t_c \frac{kT}{C} \cdot \tag{13}$$

The amplifier power outputs are not exactly proportional to $\overline{E_1^2}$ and $\overline{E_2^2}$. Following W. H. Harris,[5] let us assume that the amplifier noise is caused by a hypothetical resistance of value R_{eq} in series with the grid of the first tube. The amplifier outputs are then

$$\overline{E_{01}^2} = (\overline{E_1^2} + 4kTB_2R_{eq})g_v^2 \text{ volts,} \tag{14}$$

$$\overline{E_{02}^2} = (\overline{E_1^2}t_c + 4kTB_2R_{eq})g_v^2 \text{ volts,} \tag{15}$$

where B_2 is the amplifier equivalent noise bandwidth and g_v the voltage gain.

When $4kTB_2R_{eq}$ is small compared to $\overline{E_1^2}$, which means that $4B_2R_{eq} \ll 1/C$, it can be neglected and t_c is substantially equal to $\overline{E_{01}^2}/\overline{E_{02}^2}$. Except at high frequencies it is not difficult to build an amplifier to meet this condition.

In practice, of course, equations (14) and (15) break down when the resistance approaches the value of the shunt impedance of the tuned circuit or when its loading is sufficient to make the bandwidth of the input circuit approach that of the rest of the amplifier.

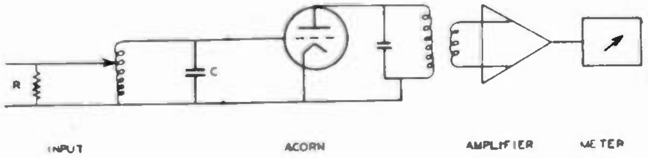

Fig. 2—High-Q input circuit.

The foregoing discussions assume that F_2, the intermediate-frequency noise figure, is known. These noise figures are generally small and sometimes difficult to determine with a signal generator. A substitute method which involves the use of a temperature-limited diode is very satisfactory. The mean-square noise current from such a tube is equal to $2eIB$ amperes^2 where I is the direct-current diode current in amperes and $e = 1.59 \times 10^{-19}$ coulomb is the charge on an electron. When this noise current flows through an impedance Z, the mean-squared voltage produced across the terminals of the impedance is equal to $2eIBZ^2$ volts.[2] If $Z = R$ = impedance across the input of the amplifier, and the diode current is adjusted so that the ratio of the amplifier outputs with and without the diode current flowing is 2, then the noise figure of the amplifier is[6]

$$F_2 = \frac{eIR}{2kT} = 20IR. \tag{16}$$

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[4] L. C. Peterson and F. B. Llewellyn, "The performance and measurement of mixers in terms of linear network theory," Proc. I.R.E., vol. 33, pp. 458–476; July, 1945.

[5] W. A. Harris, "Fluctuations in space-charge-limited currents at moderately high frequencies," Part V, *R.C.A. Rev.*, vol. 5, pp. 505–524; April, 1941. Our use of an equivalent noise resistor in this simple fashion is not as rigorous as Harris' use but quite satisfactory for the present discussion.

[6] For a discussion of the factor 20, see page 507 of footnote reference 5.

The very reasonable assumption is made in (22) and (23) that the impedance of the diode is very large compared to the impedance Z. This general scheme has been discussed by North[7] and used with satisfactory results by E. J. Schremp to measure the noise figures of intermediate-frequency amplifiers.

II. INTERMEDIATE-FREQUENCY IMPEDANCE OF CRYSTALS

All but one of the methods discussed in the preceding pages require a determination of the intermediate-frequency impedance of the crystal while it is being excited by a local oscillator. This impedance can be determined with the aid of an alternating-current bridge operated at the intermediate frequency, or by noting the change in bandwidth of a tuned circuit when it is connected across the crystal.

III. MEASUREMENT OF CONVERSION GAIN

There are two general methods for determining the conversion gain of a crystal. The first is the obvious, straightforward method of measuring the available input and output powers, while the second, in effect, compares the crystal output with the crystal noise.

With one exception[8] the direct methods of determining conversion gain involve no new procedures. The indirect method of comparing the crystal output to the crystal noise is sufficiently novel to be worth describing. A radio-frequency signal generator whose output signal level is calibrated is required.

The noise output of the intermediate-frequency amplifier with the crystal at the input is given by (6). If radio-frequency signal of known magnitude is added until the intermediate-frequency output is doubled (or increased by a known amount) the crystal conversion gain is at once calculated from knowledge of the quantities in the above equation. Specifically, if sufficient signal power S_1 is added to double the output, then

$$S_1 \times g_1 \times g_2 = (N_e + F_2 \times kTB)g_2, \qquad (17)$$

and

$$g_1 = \frac{N_e + F_2 \times kTB}{S_1} \qquad (18)$$

g_1 being the conversion gain in question.

The signal level required to double the amplifier noise is in the neighborhood of 10^{-12} watt, while the smallest radio-frequency power which can be conveniently measured is 10^{-6} or 10^{-7} watt. Hence, the radio-frequency attenuator must be accurate to a fraction of a decibel over some 60 decibels range. This range can be reduced by splitting the intermediate-frequency amplifier into two sections and inserting a variable intermediate-frequency attenuator between the two sections; the reduction in range of the radio-frequency attenuator, of course, being equal to the amount of intermediate-frequency attenuation added. Obviously this intermediate-frequency attenuation must not be so great that the signal is reduced to the level of the noise of the tube following the attenuator.

As this procedure involves the measurement of a mixture of sine signal and noise, the characteristics of the detector circuit must be known. A square-law detector is simplest to use, since in this the noise and the signal powers add directly. However, there may be times when it is desirable to use a linear detector, such as a vacuum-tube diode, because of its mechanical and electrical ruggedness. D. O. North[9] has examined theoretically the properties of various mixtures of sine signal and noise and the effect of introducing these mixtures into a linear detector. Computations based on the assumption that the noise bandwidth is small compared to the mid-frequency, and that a standard diode detector circuit with a time constant equal to or somewhat less than the reciprocal of the bandwidth is used, are not difficult and yield results applicable to this problem. In particular, North calculated that increasing the detector meter reading by adding sine signal to 1.4 times the reading due to noise alone indicates that the output signal power is equal to the output noise power. Approximately the same conclusions can be drawn from the equations derived by Ragazzini[10] and Bennett.[11] Experiments under the condition that the output-meter readings due to the amplifier noise are large compared to the diode current which flows when the amplifier is cut off confirm these calculations. This result is particularly applicable to the indirect method of measuring conversion gain discussed above.

The problem of overloading of amplifiers should also be mentioned. Jansky[12] has shown that noise contains peaks of significant power content that are some four times (in amplitude) the effective value of the noise. As a result, an amplifier will overload considerably more easily on noise than on sine signal. Actually, it has been determined that, if the amplifier does not overload on sine signal some three times the average noise power, then measurements which depend on adding signal to noise will not be appreciably in error.

ACKNOWLEDGMENT

Considerable credit for the development of crystal-noise-measuring procedures should be given to M. C. Waltz, who measured a great many crystals while trying the schemes outlined above and made many helpful suggestions.

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[7] D. O. North, "Fluctuations in space-charge-limited currents," Part II, *R.C.A. Rev.*, vol. 5, pp. 441–472; July, 1940.

[8] S. Roberts, "A simplified analysis of conversion loss of crystal converters," *Rad. Lab. Rep.* no. 53–23, July 3, 1943.

[9] Letter to the writer dated June 3, 1942. Presented orally, 1944 I.R.E. Winter Technical Meeting, New York, N. Y., January, 1944.

[10] J. R. Ragazzini, "The effect of fluctuation voltages on the linear detector," PROC. I.R.E., vol. 30, pp. 277–288; June, 1942.

[11] W. R. Bennett, "Response of a linear rectifier to signal and noise," *Jour. Acoust. Soc. Am.*, vol. 15, pp. 164–172; January, 1944. Bennett does not describe the use of the diode as a noise meter but does present the basic relations.

[12] K. G. Jansky, "Experimental investigations of the characteristics of certain types of noise," PROC. I.R.E., vol. 27, pp. 763–768; December, 1939.

Transit-Time Effects in Ultra-High-Frequency Class-C Operation[*]

W. G. DOW[†], SENIOR MEMBER, I.R.E.

Summary—**Transit-time effects in ultra-high-frequency triodes and tetrodes used in class-C operation are analyzed. Effects discussed are electron-transit reactance; electron-transit phase-delay angle; cathode back-heating; use of a screen grid to improve efficiency; changes in optimum shunt impedance; secondary emission; and anode back-heating by secondary electrons. It is pointed out that by increasing voltage and current density simultaneously the frequency and power can be raised without sacrificing efficiency or bandwidth. An equivalent circuit is described which takes account of certain important transit-time effects.**

I. INTRODUCTION

THIS PAPER will discuss principles important in the design and use of power-output triodes and tetrodes for class-C service at frequencies high enough so that electron transit time is important. The treatment presupposes the use of disk seals and cavity-type circuits, although lumped-constant symbolism is used. The analysis relates primarily to continuous-wave operation in the grid-separation[1–3] type of circuit. With a tetrode, the output resonant circuit is connected between screen and plate, and the input resonant circuit between grid and cathode.[4] The grid is by-passed strongly to the screen, so that the grid and screen operate at a common alternating-current potential. The screen operates at or near anode direct-current potential.

Rationalized meter-kilogram-second units are used. The term "circuit efficiency" will signify the ratio of radio-frequency power delivered to a load by the output resonant ("tank") circuit to the total radio-frequency power appearing in the output resonant circuit. Thus copper losses in the tank circuit reduce the circuit efficiency. "Electron interaction efficiency" will signify the the ratio of the total radio-frequency power appearing in the output resonant circuit to the sum of the alternating- and direct-current power in the electron stream.

II. CLASS-C EFFICIENCY, OPTIMUM SHUNT IMPEDANCE[5,6]

In any class-C circuit good electron interaction efficiency will be obtained only if the electrons have little or no kinetic energy when they return to the metallic circuit by striking an electrode. If transit time is not important, this requirement means that the electrons must strike an electrode that is very little higher in potential than the cathode at the time of impact. At frequencies for which transit time is not important, the following requirements relative to Fig. 1 represent a set of necessary and sufficient conditions for keeping the electron energies small at impact:

1. The "plate swing" must approach as closely to 100 per cent of the direct-current plate voltage as practical considerations permit.[5]
2. The "angle of plate current flow"[5] must not exceed a reasonably small fraction of one-half cycle.
3. The plate-current pulses must be in phase with the downward swings of the plate voltage.

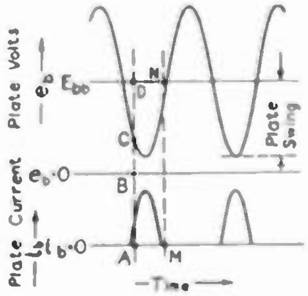

Fig. 1—Plate current and plate voltage variations in a class-C electronic alternating-current power generator.

In Fig. 1, every electron that passes from cathode to anode at the moment *ABCD* receives from the direct-current power source E_{BD} electron volts of energy, delivers as plate-dissipation energy E_{BC} electron volts of energy, and to the tank circuit, as radio-frequency power generation, E_{DC} electron volts of energy.

The electron interaction efficiency for each such electron is the ratio of E_{DC} to E_{DB}. The over-all electron interaction efficiency is the weighted average of such ratios.

If the plate swing is inadequate, none of the electrons will experience a high interaction efficiency. If the angle of flow is large, many of the electrons will experience a very low or even negative interaction efficiency. If the current pulse were shifted in phase relative to the plate swing by a large amount, no electrons would contribute a satisfactory interaction efficiency.

The angle of flow is determined primarily by the extent of the upward swing of the grid voltage in relation to the grid bias.

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[1] M. C. Jones, "Grounded-grid radio-frequency voltage amplifiers," Proc. I.R.E., vol. 32, pp. 423–429; July, 1944.

[2] M. Dishal, "Theoretical gain and signal-to-noise ratio obtained with the grounded-grid amplifier at ultra-high frequencies," Proc. I.R.E., vol. 32, pp. 276–284; May, 1944.

[3] C. E. Strong, "The inverted amplifier," *Electronics*, vol. 13, p. 14; July, 1940.

[4] W. W. Salisbury, "The resnatron," *Electronics*, vol. 19, pp. 92–97; February, 1946.

[5] F. E. Terman, "Radio Engineering," McGraw-Hill Publishing Company, New York, N. Y., 1937, chap. 7.

[6] W. L. Everitt, "Optimum operating conditions for class-C amplifiers," Proc. I.R.E., vol. 22, pp. 152–177; February, 1934.

If the grid drive has a frequency equal to the resonant frequency of the tank circuit, the downward plate-current swing will occur in phase with the plate-current pulses. If the grid-drive frequency is low or high by this standard, the peaks of the plate-current pulses will lag or lead the downward plate swings. The amount of lead or lag, in electrical degrees, caused by a given frequency deviation, will be relatively large for a high-Q tank circuit. The extent of the output-tank-circuit voltage swing is, of course, proportional to the product of the fundamental-frequency component of plate current by the shunt impedance of the tank circuit for the frequency at which the tube is operating.

In class-C operation, for any given combination of tube and supply voltages used, there is an optimum value of output-tank-circuit shunt resistance.[6] When working into a load at resonance with a tank circuit that presents the optimum shunt resistance, the plate-current pulses produce a plate swing desirably close to 100 per cent. If the shunt resistance, often called "shunt impedance," is substantially less than the optimum value, the plate swing will fall far short of 100 per cent and the electron interaction efficiency will be poor. If the shunt impedance is appreciably greater than optimum, the plate swing exceeds 100 per cent. As a result the electron interaction efficiency is less than with optimum shunt impedance, for causes arising from electron rejection by the plate[7]; that is, from the fact that electrons cannot reach a negative plate.

In a grid-separation circuit, "100 per cent plate swing" exists when the maximum instantaneous value of the algebraic sum of input and output radio-frequency voltages equals the direct-current plate-supply voltage. This means, of course, that at the maximum point the instantaneous value of the voltage across the output tank circuit exceeds the direct-current plate-supply voltage by the amount of the instantaneous value of the radio-frequency input voltage at that moment. Thus, optimum shunt impedance is greater with a grid-separation circuit than with a comparable cathode-separation circuit.

The optimum value of shunt impedance is, in general, directly proportional to the direct-current plate-supply voltage, in that if the direct-current voltage is raised the shunt impedance must be increased to maintain 100 per cent plate swing. It is very roughly inversely proportional to plate current in that if the plate current is increased, as for example by increasing peak emission, the shunt impedance must be reduced to prevent the plate swing from exceeding 100 per cent.

Although details and magnitudes differ, the general fact of the existence of an optimum shunt impedance carries over into ultra-high-frequency operation where transit time is important.

In most ultra-high-frequency applications, the need for substantial bandwidth calls for low-Q output tank circuits. For example, to accommodate a 10-megacycle bandwidth at a 1000-megacycle carrier frequency, the Q must not exceed 100, a value on the low side in ultra-high-frequency practice. For the circuit of Fig. 2, using symbols defined therein, and with radio-frequency input frequency such that output radio-frequency voltage and current are in phase, the shunt resistance, or shunt impedance at resonance, is

$$R_{sh} = 1/G_{sh} = \frac{1}{G_0 + G_L}. \tag{1a}$$

The loaded Q is then

$$Q_L = R_{sh}\sqrt{\frac{C}{L}}. \tag{1b}$$

Thus for low Q the shunt resistance must be small and the ratio of L to C large, for a given frequency-determining product LC. However, transit-time considerations usually set a lower limit to practically usable values of C.

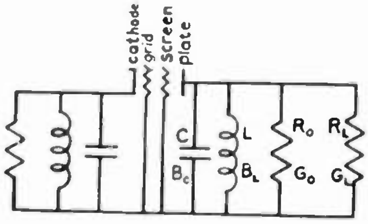

Fig. 2—Alternating-current schematic tetrode circuit diagram, grid-separation circuit. The radio-frequency plate-circuit load is shown as divided between useful output loading, $G_L = 1/R_L$, and internal tank-circuit-loss loading, $G_0 = 1/R_0$. The screen is effectively by-passed to the grid, so that grid and screen are at a common radio-frequency potential.

Thus if Q is to be kept low, the value of R_{sh} must be kept small; for efficient operation this small R_{sh} must be the optimum shunt impedance for the tube and plate supply used. This calls for the use of a tube having a high emission capability, which must be the result of high *emission-current density*. Thus the tube designer must keep C as low as transit-time considerations permit, and increase the emission-current density to a maximum extent. Increasing total emission by enlarging the active electrode-structure area facing the anode does not help, because it results in a compensating increase in C and associated forced decrease in L, both roughly in proportion to the change in cathode area.

III. Limits to Obtainable Power Levels at Ultra-High Frequencies

As far as basic electron-transit-time limitations are concerned, it should be as easy to produce ultra-high-frequency radio-frequency power efficiently at the levels of tens and hundreds of kilowatts as it is to produce it efficiently at levels of a few watts. This statement is made with full recognition of the fact that various effects generally detrimental to electron interaction efficiency begin to appear when the frequency becomes high enough to make any interelectrode transit angle within the tube an important fraction of a quarter of a cycle.

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[7] W. G. Dow, "Fundamentals of Engineering Electronics," John Wiley and Sons, New York, N. Y., 1937, chap. 3.

An obvious approach to minimizing these detrimental effects is to exert design effort toward keeping transit times small, either by making small the distance the electrons must travel between electrodes, or by making large the average velocity during transit.

If, in order to keep transit time to a minimum as the frequency is raised, *the interelectrode spacing is to be made small*, the active electrode area must also be small in order to maintain the interelectrode capacitance at a properly low value in relation to shunt impedance and the requirement of a low Q. Manufacturing tolerances limit progress in this direction, which is limited to low-power-level operation.

If, in order to keep transit time to a minimum as the frequency is raised, *the average transit velocity is to be made very rapid*, a high direct-current plate voltage may be used. The interelectrode spacing need not then be small, so that the active electrode area may be kept substantial without making the interelectrode capacitance excessive. If the low-Q and optimum-shunt-impedance conditions are to be fulfilled, the current must be kept high in proportion to the voltage *by increasing the current density*. Thus as the frequency is raised, good efficiency and low Q can be maintained, *if both voltage and current, and therefore power, are increased sufficiently, at very-high current density.*

This requires high-dissipation anodes, very-high-current-density cathodes, and clearances and vacuum techniques sufficient to prevent flashovers. At very substantial power levels, of the order of tens of kilowatts, these problems are in all probability capable of early and practical solution up to 3000 megacycles.

IV. Electron-Transit Reactance; Electron-Transit Phase-Delay Angle

At ultra-high frequencies it is generally true that *the fundamental-frequency component of radio-frequency current flowing through the interaction space associated with the tank circuit lags in phase behind the fundamental-frequency component of the electron stream injected through the grid or screen into the interaction space.*[8] There exists then an *electron-transit phase-delay angle* which can be used to describe the important transit-time phenomena. It is desirable to analyze this behavior, and to describe an *electron-transit reactance*, used in an equivalent circuit to symbolize certain transit-time effects.

Fig. 3 illustrates various positions of a "block" of electrons, of extent $x_b - x_a$ during passage from left to right through the equipotentials at x_a and x_b with constant space-charge density ρ and constant velocity v. This movement of charge causes a current flow between the equipotentials, generally called the "induced" current, that begins when the forward face of the block passes x_a. The induced current density,[8-10] which includes both the displacement and convection currents caused by the movement of this block of charge, is

$$J = \int_{x_a}^{x_b} \frac{\rho v dx}{x_b - x_a}. \qquad (2)$$

The displacement current flowing as a direct result of any coincident change in potential difference between x_a and x_b is entirely distinct from and not a part of the induced current; symbolically it is thought of as flowing through a separate parallel condenser.

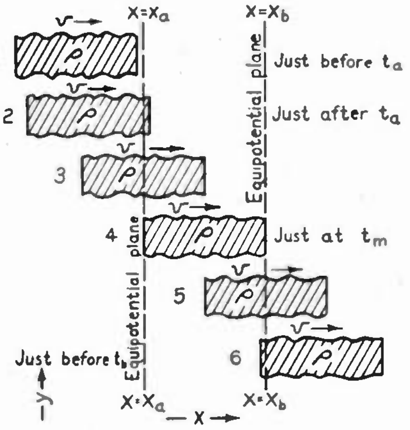

Fig. 3—Diagrammatic portrayal of various positions of a block of space charge passing with velocity v from left to right past the equipotential plane at $x = x_a$, then later past the equipotential plane at $x = x_b$. All block representations are shown with broken-line top and bottom edges to indicate that they extend infinitely in both y directions. The front end of the block passes x_a at moment t_a: the rear end of the block passes x_b at t_b; at t_m the block is centered and just fills the space between x_a and x_b.

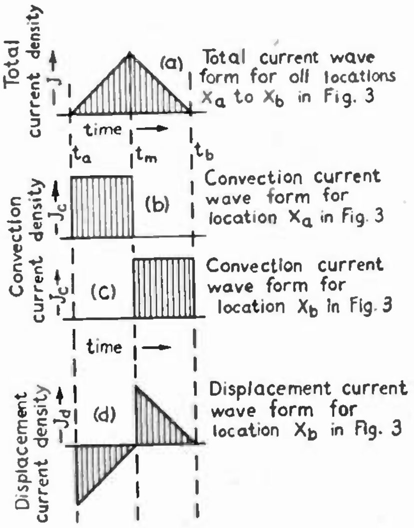

Fig. 4—Current wave forms associated with Fig. 3.
(a) Induced current wave form for all positions x_a to x_b, in Fig. 3.
(b) Convection current wave form for position x_a in Fig. 3.
(c) Convection current wave form for position x_b in Fig. 3.
(d) Wave form of displacement component of induced current for position x_a, Fig. 3.

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[8] F. B. Llewellyn, "Electron Inertia Effects," Cambridge University Press (in New York, The Macmillan Company), 1941. This book contains a complete bibliography relative to small-signal ultra-high-frequency electron behavior.

[9] S. Ramo, "Current induced by electron motion," Proc. I.R.E. vol. 27, pp. 584–586; September, 1939.

[10] C. K. Jen, "On the induced current and energy balance in electronics," Proc. I.R.E., vol. 29, pp. 345–349; June, 1941; also C. K. Jen, "On the energy equation in electronics at ultra-high frequencies," Proc. I.R.E., vol. 29, pp. 464–466; August, 1944.

For constant-velocity constant-ρ charge movement, the ***induced-current*** wave form for the entire region x_a to x_b will be as illustrated in Fig. 4(a). The ***convection*** component of induced current has the wave form shown in Fig. 4(b) at position x_a, and at points infinitesimally displaced to the right and left of x_a. The relative position of the wave forms of Figs. 4(a) and 4(b) is clearly shown by the figure. The wave form and phase position of the convection-current component at and near x_b are shown in Fig. 4(c). At any point between x_a and x_b the convection-current component wave form has the same shape as at the two ends, but its phase position shifts uniformly from left to right.

The wave form of the displacement component of induced current is shown in Fig. 4(d), for position x_a in Fig. 3.

The planes x_a and x_b might correspond to the grid and screen planes of a tetrode. In that case the phase angle between the fundamental-frequency components of Figs. 4(b) and 4(c) represent the electron-transit phase-delay angle between injection past the grid and ejection past the screen.

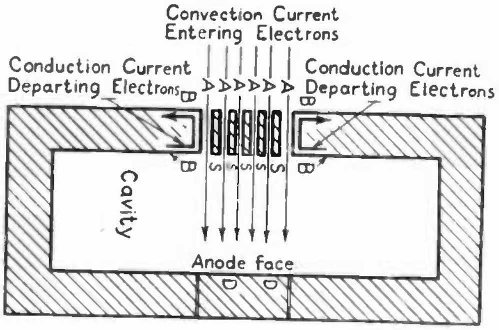

Fig. 5—Electron stream injected into a cavity via an "ideal" grid.

If the plane x_a corresponds to the screen plane, and x_b to the plate plane, the phase angle between the fundamental-frequency components of Figs. 4(b) and 4(a) is the electron-transit phase-delay angle between injection current at the screen and the load current to the tank circuit.

This latter condition is illustrated by Fig. 5, representing an output cavity which electrons enter by means of an "ideal" grid. The grid "wires" are thin enough to offer very little obstruction to electron flow, but they are deep enough so that there is no penetration of the electric field at the left into the region at the right, or vice versa.

Only convection current or conduction current can pass through an "ideal" grid. The zero-field condition at the grid plane prevents any passage of displacement current into or out of the cavity across the grid plane. The entrance of the electron stream constitutes a flow of electrons into the cavity; since no displacement current can enter or leave, there must also be a flow of electrons out of the cavity. Ultra-high-frequency conduction currents pass only on the surfaces of conductors, so that there can be no conduction-current escape of electrons from the cavity except at the edges of the screen openings. Thus, the cavity shown in Fig. 5, looked at from the outside as a high-frequency device, is a two-terminal network. One terminal is at A, A, A, the open spaces between the grid wires, where the electrons enter as convection current; the other terminal is at B, B, the edges around the openings in which the grid is placed, where the electrons leave as surface conduction current.

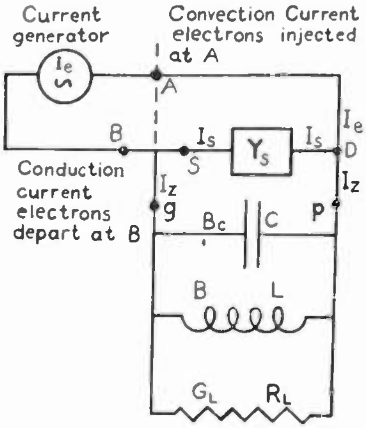

Fig. 6—Convection-current electrons injected at A; conduction-current electrons depart at B. Equivalent circuit diagram for a cavity into which convection current is injected in the form of an electron stream. I_e is generated by an alternating-current generator symbolizing the current injection. Positions A, B, S, D correspond to positions similarly marked in Fig. 5.

The entering electron convection stream and the departing conduction current flow past one another at about the same electrostatic potential, but with no opportunity for mingling.

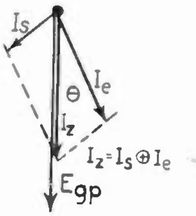

Fig. 7—Vectors showing phase relations between cavity voltage E_{gp} and I_e, I_s, I_z in Figs. 5 and 6. The angle θ is the electron-transit phase-delay angle between injection current and load current.

If a region contains both entrance and exit grids, so that the beam passes through both, just as it is indicated as passing through equipotentials at x_a and x_b in Fig. 3, the cavity is a four-terminal network, all entering and departing currents consisting, however, entirely of convection or conduction currents.

Instead of using the fiction of an "ideal grid," it is possible to entertain the concept of an equivalent grid-plane potential.[8,11,12] The locations A and S then have the potential of the equivalent grid plane, which is connected circuitwise by a capacitance to the points B on the cavity conductor surface and to the grid wires, which

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[11] W. G. Dow, "Equivalent electrostatic circuits for vacuum tubes," Proc. I.R.E., vol. 28, pp. 548–556; December, 1940.

[12] B. J. Thompson, "Space-current flow in vacuum-tube structures," Proc. I.R.E., vol. 31, pp. 485–491; September, 1943.

are at the same potential as B, B. The larger this capacitance is, the closer the situation approaches that of an ideal grid. None of the over-all concepts here discussed are basically altered if the equivalent grid-plane potential concept is used rather than that of an ideal grid.

Figs. 6 and 7 are, respectively, an equivalent circuit and a vector diagram, corresponding to Fig. 5 and to a current flow into the cavity similar qualitatively but not quantitatively to Figs. 3 and 4. The length of the block of charge is not $x_b - x_a$ and the charge density and velocity are nonuniform. However, (2) is valid, v and ρ being functions of x.

The injection current is considered as being ordered into existence by the current generator I_s; it enters the cavity as injected convection current at A, leaves it as surface metallic conduction current at B, and corresponds qualitatively to Fig. 4(b). The load current I_z is the induced current due to the passage of the block of charge; it corresponds to Fig. 4(a), and to the *negative of the plate current* in ordinary electron-tube symbolism. Fig. 7 is drawn on the assumption that the frequency of the injection current is the resonant frequency of the tank circuit, so that I_z is in phase with the output voltage, which is E_{gp} for a grid-separation circuit.

The angle θ in Fig. 7 is the electron-transit phase delay angle between injection current and load current. It will in general differ from the transit angle calculated on the basis of direct-current or maximum or minimum voltages, but it is capable of evaluation by proper measurement techniques.

The electron-transit admittance, Y_e in Fig. 6, symbolizes the fact that the current I_z flows between two points of unlike potentials. Y_e has in general inductive susceptance, and may conceivably have either positive or negative conductance; it is intimately related to the current generator I_s. Llewellyn and Peterson have discussed similar small-signal concepts.[8,13]

In Fig. 5 the currents I_s and I_z have separate, distinguishable identities only at the A, B location. Between the A, B and the D locations there is a continuous point-to-point variation of the displacement and convection components of I_z. The points g and p of Fig. 6 cannot be identified in Fig. 5, because Fig. 5 illustrates a distributed circuit.

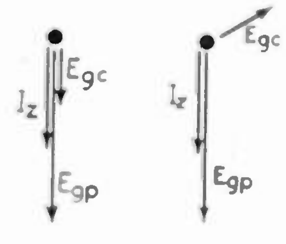
The physical embodiment of the ability of the current generator to handle either large or small voltages on call lies in the fact that a large or small proportion of the direct-current beam energy brought in by the electrons at A will be used, according to what the radio-frequency circuit demands. What is not used is wasted as plate dissipation. Thus the upper limit to the voltage that the generator I_s can handle is set by the direct-current power supply.

The important physical fact is that I_z must lag behind I_s and that the angle of lag depends on the direct-current voltages and currents. Thus the circuit can be "tuned" by varying the direct-current parameters; this operation can be thought of as causing a variation in the "electron-transit susceptance" which is the important component of Y_e.

The terms "electron-transit reactance" and "electron-transit susceptance" are applicable to all situations in which reactance caused by electron transit time appears in a circuit; thus magnetrons possess electron-transit reactance, and entirely new devices may appear whose sole function is to introduce an adjustable electron-transit reactance or susceptance.

V. Effects of Electron-Transit Reactance; "Frequency Pushing"

Obviously there will be electron-transit phase-delay angles for both input and output circuits in a grid-separation triode, with an added grid-to-screen-transit phase-delay angle appearing in a tetrode. Because of the extremely slow velocities of electrons at and near the cathode, the transit phase-delay angle for the cathode-to-grid transit becomes important at a much lower frequency than do the others, and is larger than the others at all frequencies. The various transit delay angles add to produce an over-all electron transit phase delay angle between input and output circuits.



(a) (b)

Fig. 8—Vector diagrams showing phase relationships between input voltage, E_{gc}, output voltage E_{gp}, and load current I_z in a grid-separation-circuit class-C oscillator with and without an over-all input-to-output electron-transit phase-delay angle.
(a) No electron-transit phase-delay angle.
(b) About 120-degree electron-transit phase-delay angle.

Fig. 8 illustrates two contrasting situations, both under conditions where the input frequency is the common resonant frequency of input and output circuits.

Obviously, if an oscillator is to operate most effectively, the feed-back circuit must incorporate a phase-advance angle to compensate for the over-all transit delay angle. Any modulation will change the transit delay angle; the frequency must then shift to retain over-all zero susceptance. Thus amplitude modulation becomes associated with a marked frequency modulation, synchronized with the amplitude modulation.

This effect is the major one contributed to by the cathode-to-grid electron-transit reactance. It appears at a lower frequency than do other transit-time effects, because the very low initial velocities of the electrons cause the cathode-to-grid transit phase-delay angle to be relatively large. It is a very prominent effect, but is usually not serious in relation to efficiency.

It became common practice in ultra-high-frequency

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[13] F. B. Llewellyn and L. C. Peterson, "Vacuum-tube networks." Proc. I.R.E., vol. 32, pp. 144–166; March, 1944.

experimentation during the recent war years to use the term "frequency pushing" to describe the variation of oscillator frequency with plate current or associated parameters. This name was chosen to contrast with the term "frequency pulling" which is used to describe the effect on frequency of changes in load reactance. There is serious doubt as to the desirability of incorporating into accepted engineering terminology the phrase "frequency pushing." The use in this paper of this phrase should not be considered as prejudicial to the future adoption of a better term.

It is reasonable to expect that frequency pushing in an ultra-high-frequency oscillator may cause frequency variations of the order of $\frac{1}{4}$ to 1 per cent of the normal operating frequency, for amplitude modulation of the order of 80 to 90 per cent. The variation in frequency is not a linear function of current.

The effect on an amplifier introduced by electron-transit reactance is the same as that of introducing reactance by any other means, that is, it changes the tank-circuit resonant frequency, thus requiring a change in the drive frequency to achieve optimum shunt impedance. The sharpness of the effect depends, of course, on the *Q* of the circuit, being the most severe for high-*Q* circuits. Also, electron-transit reactance in an amplifier causes amplitude modulation to set up a synchronized phase modulation.

Electron-transit reactance will make difficult the use of ultra-high-frequency power generators in exact accord with previous radio communication practice. However, it should be pointed out that any tube which displays the phenomenon of frequency pushing is itself a reactance tube, and may be used for producing frequency modulation of an associated circuit. It is, of course, reasonable to expect that tubes can be designed to maximize this type of functioning. Reactance tubes, based on the use of electron-transit reactance, can be used to compensate for pushing in an oscillator or for phase modulation in an amplitude-modulated amplifier, or for producing frequency modulation at constant amplitude.

VI. USE OF SCREEN TO PREVENT TRANSIT DEBUNCHING

The sine-wave curve in Fig. 9(b) describes the variation with time of the potential relative to the cathode of the equivalent grid plane in an idealized ultra-high-frequency triode.[8,11,12] At moment *AA* electrons will begin to leave the cathode; subsequent to *BB* no electrons leave the cathode until *AA* is repeated in the next cycle.

The curves in Fig. 9(a) portray the distance-time histories of electrons that leave the cathode at various moments.[14–16] Slopes of these curves are proportional to electron velocities. These curves are shown quantitatively correctly, for a particular charge-free-space condition, for the region between cathode and grid. They are shown as straight lines from grid to plate, which corresponds to the existence of no potential difference and therefore no acceleration between the equivalent grid plane and the plate. The diagram is so drawn partly because it is a reasonably close approach to the truth for efficient class-C operation. The chief reason for drawing these lines straight, however, is that it is the simplest way to illustrate the point about to be made.

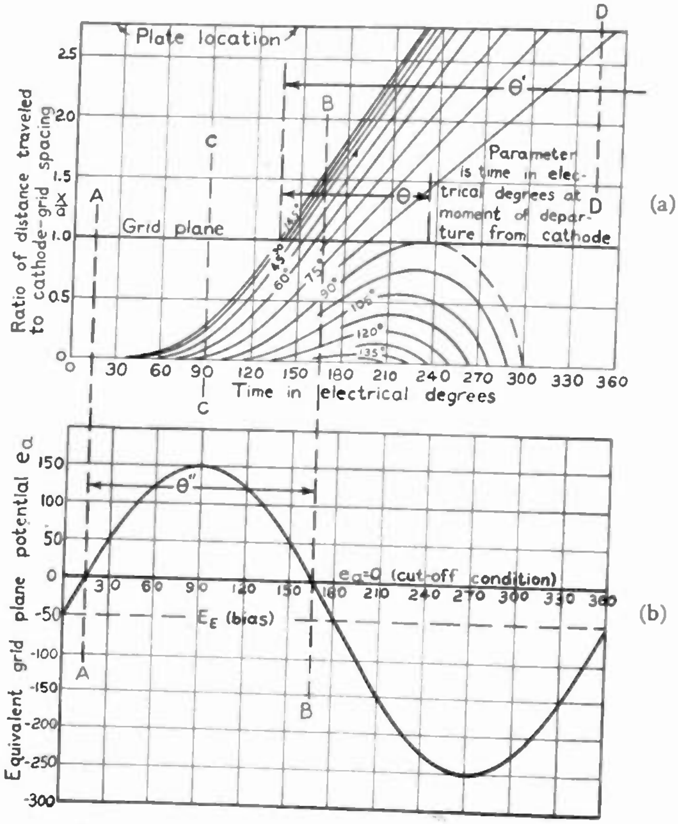

Fig. 9—Illustrative electron distance-time curves, triode, for a cathode angle of flow of 151 degrees. Charge-free space assumed in calculating the curves. The effect of the existence of a space-charge-limited condition would be to increase the cathode-to-grid transit time about 50 per cent, relative to the values here indicated for charge-free space.[11–13]

Fig. 9(a) as drawn indicates that the electrons that leave the cathode before the moment *CC* pass on through the grid to the plate, passing the grid during the angle θ; the later electrons return to the cathode. It might seem proper to call the angle θ the angle of flow of plate current. However, as long as electrons are traversing the space from grid to anode, plate current is still flowing; thus electrons are still arriving at the plate, and therefore current is still flowing, at and later than at moment *DD*. A comparison between the shapes of the plate-current pulses with and without this effect appears in Fig. 10. Since the plate current tails off to zero asymptotically in the condition corresponding to Fig. 9, it is difficult to define at what moment the angle of flow ends.[16]

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[14] Chao-Chen Wang, "Large-signal high-frequency electronics of thermionic vacuum tubes," PROC. I.R.E., vol. 29, pp. 200–213; April, 1941.

[15] H. E. Hollman, "Theoretical and experimental investigation of electron motion in alternating fields with the aid of ballistic models," PROC. I.R.E., vol. 29, pp. 70–79; February, 1941.

[16] B. Salzberg, unpublished thesis, 1941; see footnote 14, p. 200.

Thus the electrons leave the grid plane well "bunched" into a small, initial, angle of flow; they have differing velocities as they pass the grid; a drift-tube action then takes place in the grid-to-plate space which "debunches" the electrons by the time they reach the plate, and causes the actual effective operating angle to be poor.

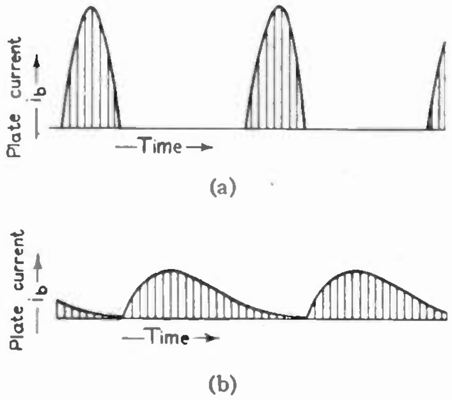

Fig. 10—Qualitative illustration of the effect of transit debunching on plate-current wave form.
(a) Without transit debunching.
(b) With transit debunching.

The introduction of a screen grid operating at control-grid radio-frequency potential but at or near anode direct-current potential causes the average velocities of the electrons during their travel beyond the grid to be large enough so that the effect of the differences between their velocities at grid passage becomes unimportant. This more or less completely eliminates the debunching action. Thus the second grid assists in maintaining a good angle of plate-current flow, if other circuit conditions are right.

VII. CATHODE BACK-HEATING

The distance-time curves in Fig. 9(a) for the electrons that return to the cathode are shown diving into the cathode at a substantial slope; thus these electrons have considerable kinetic energies at impact with the cathode. Important effects of this behavior, called cathode back-heating, are:

1. Back-heating causes an increase in the input conductance. This may prove to be an ultimate limiting factor in ultra-high-frequency amplifier practice.

2. The back-heating varies with radio-frequency input drive, and with frequency in a tunable device. Therefore, if the cathode temperature is to be maintained constant, continuous adjustment of the filament excitation is necessary, if drive or frequency vary.

3. Cathodes must be designed so that the normal filament excitation power is greater than the maximum back-heating power, in watts.

4. During a portion of each cycle the cathode is emitting electrons that play no useful part in the tube operation.

The actual average emission current may be greater, perhaps by half or more, than will be read by any direct-current ammeter, because the back-heating electrons do not pass through the plate circuit. Since the unmeasured emission current occurs *after* the useful electrons have left the cathode, it does not affect the peak emission requirement. Cathode *life* is related to ampere-hours of emission, rather than to peak emission. The period during which the back-heating electrons are departing from the cathode adds to the ampere-hours of emission, and therefore affects cathode life.

VIII. USE OF SCREEN GRID TO SHORTEN INTERACTION TIME

Referring to Fig. 1, if the grid-plate transit angle in a triode is a substantial portion of a half-cycle, an electron that passes the grid plane at the moment $ABCD$ will not arrive at the anode until some such moment as MN. It follows that, even if all the electrons start well bunched and remain so, if they move slowly enough the effect will be similar to that of a large angle of flow, and the efficiency will be poor.

This difficulty can be made less serious by reducing the "electron interaction time," or time of flight through the region in which the electrons by their movements convert direct-current power into alternating-current power.

Calculations and experience both indicate that an interaction time of about one-quarter cycle can be tolerated from an efficiency standpoint, but that a half cycle is too long. Any attempt to reduce the interaction time by reducing the interaction spacing will increase the capacitance in the tank circuit, a change which in general cannot be tolerated because of its effect on the circuit Q.

The alternative is to increase the average electron velocity during interaction. This can be done, *without increasing the plate voltage*, by introducing a second grid between control grid and plate and operating the second grid at control-grid radio-frequency potential and at or near anode direct-current potential. The tank circuit capacitance is then that between screen and anode. This can have the same value as in a comparable triode, yet the tube can operate with a considerably smaller interaction time, because of the high velocity with which the electrons pass though the screen into the interaction space.

There has been ample experimental verification, at power levels of tens of kilowatts and frequencies be tween 350 and 650 megacycles, of the relationships discussed above involving tank-circuit capacitance, interaction distance, and efficiency, when the second grid is employed.[4]

Approximate calculations and experience to date indicate that it is probable that the use of the second grid may raise, by a factor approaching 2.5, the upper frequency limit at which good power-output efficiency is obtainable at usable bandwidths.

Transit debunching and interaction transit time are separate topics, although affected similarly by the second grid. It is possible to describe, in principle at least, a type of operation in which there is transit bunching rather than debunching, and the change from one to the other can occur at a constant *average* electron interaction time.

IX. TRANSIT-TIME EFFECTS ON OPTIMUM SHUNT IMPEDANCE

At frequencies low enough so that transit time is not important, the kinetic energy of an electron in electron volts is equal to the potential difference in volts between the point it has reached and the cathode from which it was emitted. If, however, the frequency is high enough so that the electric fields change substantially during the electron's flight, this is no longer necessarily true.

One result of this difference is that it is quite possible at ultra-high frequencies to have a plate swing well in excess of 100 per cent without introducing electron rejection by the plate. Electrons may conceivably reach the plate at a moment when it is well below cathode potential, or they may reach it with little or no kinetic energy at a moment when its potential is very much greater than the cathode potential.

Since it is thus possible at ultra-high frequencies for the plate swing to be more than 100 per cent without causing electron rejection, and also conceivable for the efficiency to be high with a plate swing of 50 per cent or less, it is no longer necessarily correct to say that the optimum shunt impedance is that which will result in nearly 100 per cent plate swing when accepting a given plate current.

The question as to the manner and extent in which the ultra-high-frequency optimum shunt impedance differs from the optimum under low-frequency conditions has not been answered definitely as yet. It seems likely that for optimum operation the plate swing should be somewhat greater than 100 per cent, and that the electrons should for the most part arrive at the plate, with very little energy, at a moment when the plate potential has risen substantially above zero.

X. SECONDARY EMISSION; ANODE BACK-HEATING

Of course, substantial secondary-emission current due to passage of electrons from plate to screen may flow in a tetrode, and in class-C operation appreciable secondary-emission current cannot usually be tolerated.

The ratio of secondary electrons to primary electrons, for the surface of an anode (usually copper), rises from zero at impact energies of a very few volts to a maximum occurring at an impact energy of a few hundred volts, then tapers off to a very few per cent for primary electrons with energies of several thousand volts.[17]

The simplest and most easily demonstrable way of avoiding secondary-emission trouble in an ultra-high-frequency tetrode is to operate it at voltages high enough to make the secondary-to-primary ratio sufficiently low to avoid trouble. This is, of course, practical only at very high power levels, such that good efficiency corresponds to electron impact energies averaging perhaps 2000 to 3000 volts. Thus high-power operation is in general easily obtained, whereas low-power operation may not be.

There are other approaches to the elimination of secondary-emission troubles; for example: some use of a third grid; space-charge suppression[18] and related schemes; and special treatment of the anode surface to minimize the secondary-to-primary electron ratio. As yet, however, no completely successful method of eliminating the secondary-emission difficulty at low voltages has been reported.

An approach to this problem that has been used with a fair degree of success is an arrangement in which vanes serving the purpose of a suppressor grid are directly connected to and made a part of the anode.[4] The electrons are beamed so as to enter the pockets between the vanes. This does not actually prevent secondaries from leaving the anode, but it can cause a large percentage of them to re-enter the anode or its attached suppressing vanes very early in their flight.

In ordinary tetrodes, secondary emission is made evident by an increase in screen current. This is not always true at ultra-high frequencies, because the action of the electric field may reverse the movement of the secondaries during transit and drive them back into the anode. This effect can be called anode back-heating by secondary electrons. A method has been devised for measuring it, based on the fact that the secondary-electron back-heating represents a radio-frequency loading on the anode tank circuit, whereas normal plate dissipation by primaries has no such effect. The anode back-heating energy can thus be "sorted out" by virtue of the fact that it represents energy extracted from the radio-frequency field, and therefore modifies the operating Q.

ACKNOWLEDGMENTS

This presentation is an outgrowth of experimentation and analysis carried out by many workers during the period 1942 through 1945. The author wishes to emphasize the parts played by very many individuals, far too many to list here. Specific names that must be mentioned are F. E. Terman, W. W. Salisbury, David Sloan, Gunnar Hok, J. J. Livingood, Georges Chevigny, G. J. Lehmann,[19] and B. Salzberg.

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[17] J. M. Hyatt and H. A. Smith, "Secondary electron emission from molybdenum," *Phys. Rev.*, vol. 32, pp. 929–935; December, 1928.

[18] H. O. Schade, "Beam power tubes," PROC. I.R.E., vol. 26, pp. 137–182; February, 1938.

[19] G. J. Lehmann and A. R. Vallerino, "Study of ultra-high-frequency tubes by dimensional analysis," PROC. I.R.E., vol. 33, pp. 712–716; October, 1945.

Discussion on

"Current and Power in Velocity-Modulation Tubes"*

L. J. BLACK AND P. L. MORTON

P. J. Wallis and S. G. Tomlin:[1] We have read the August, 1944, issue of the PROCEEDINGS OF THE I.R.E., which contains the very interesting paper by L. J. Black and P. L. Morton on "Current and Power in Velocity-Modulation tubes." This paper contains results very similar to those we have been using in these laboratories and which have been described in privately circulated reports, but we would like to point out that certain important terms have been omitted in their paper. It is hoped to publish shortly a number of papers on the principles of velocity modulation which give the full details of our method and results and should be consulted for proof of the results quoted in this letter.

The equations for the charge density ((9) and (11)) in their paper are incorrect because they neglect the variation of electron velocity at the catcher, If ρ is the charge density, v the electron velocity, and i the current, then, since charge is conserved,

$$\sum i_0 dt_1 = i_2 dt_2$$

and

$$i = \rho v$$

where the $\sum$ includes all elements dt_1 which correspond to electrons arriving at the catcher in the same interval dt_2. These equations give

$$\rho_2 = \frac{i_0}{v_2}\sum\frac{\partial t_1}{\partial t_2}$$

which has v_2 in the denominator on the right and not v_0 as is given by Black and Morton in (11).

In their section on the catcher current the authors assume in two places that it is good enough to take the electron velocity on entering the catcher as v_0 or, in other words, they neglect the modulation in the first gap except in the first term for the entrance time. In their notation this is neglecting terms of order P_1 but in their expressions for the current they keep terms of order P_2 which are not necessarily greater (in many cases $P_1=P_2$). In other words they have not retained terms of comparable magnitude consistently throughout the calculation. If, in their expression for the induced current Δi and in their integration of the equation of motion of the electron in the second gap, terms of order P_1 are kept, then it is fairly easy to modify the results and get an expression for η which retains all terms of the second order in P_1 and P_2, while their result gives only some of the terms of the second order.

In determining the order of different terms it is necessary to remember that, while the modulations P_1 and P_2 are small, the drift tube may be quite long. Writing $S=ws/v_0$, we can assume that SP_1 is an order of magnitude larger than P_1. In mathematical symbols $SP_1=0\,(1)$ while $P_1=0(1)$.

In their section on transit-time effects at the buncher, the authors show how the finite time spent in the buncher introduces a term in the efficiency due to the absorption of power in the buncher. They do not indicate, however, that because of the different emergent velocity the transit time in the drift tube is altered, and that consequently the argument r of the Bessel functions must be altered by multiplying it by a gap efficiency factor.

Although the authors confine their attention to tubes whose gaps each consist of two parallel grids, the theory can be developed for quite arbitrary gaps and the results given in terms of certain gap factors β, γ, and δ, which depend on the gap geometry and are necessary to describe the velocity of electrons after passing through a gap.

The result then obtained for the efficiency η is

$$\eta = P_2\beta_2 J_1(AP_1)\Big[\cos(C_0-\epsilon_1+\alpha)$$

$$+\frac{SP_1^2}{32}\sqrt{9\beta_1^4+4\delta_1^2}\cos(C_0+\epsilon_1-\epsilon_2+x)\Big]$$

$$+P_2\beta_2 J_3(AP_1)\,\frac{SP_1^2}{32}\sqrt{9\beta_1^4+4\delta_1^2}\cos(C_0-3\epsilon_1+\epsilon_2+\alpha)$$

$$+\frac{P_1P_2\beta_1\gamma_2}{4}\,[J_0(AP_1)\sin(C_0+\alpha)$$

$$-J_2(AP_1)\sin(C_0-2\epsilon_1+\alpha)]$$

$$-\frac{P_1^2}{4}\beta_1\gamma_1-\frac{P_2^2}{4}\beta_2\gamma_2$$

$$-\frac{P_2^2\delta_2}{4}J_2(2AP_1)\sin(2C_0-2\epsilon_1+2\alpha)$$

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* PROC. I.R.E., vol. 32, pp. 477–482; August, 1944.

[1] Standard Telephones and Cables, Ltd., Connaught House, London, England.

where

$$A = \tfrac{1}{2}\sqrt{S^2\beta_1^2 + \gamma_1^2}$$

$$\tan \epsilon_1 = -\frac{\gamma_1}{S\beta_1}, \qquad \tan \epsilon_2 = -\frac{3\beta_1^2}{2\delta_1}$$

and

$$C_0 = S\left[1 + \frac{P_1^2}{16}(3\beta_1^2 - 2\beta_1\gamma_1)\right]$$

and S is measured between the centers of the gaps. We have included in this expression for η the term due to the absorption of power in the first gap $[-(P_1^2/4)\beta_1\gamma_1]$, and written the result only for $n=1$, i.e., both high-frequency fields having the same frequency. The subscripts E and M refer to the first and second gaps respectively.

The above expression for η immediately gives the beam conductance. The expression for the beam susceptance, which is important in connection with electronic frequency control, can be found similarly, and also contains extra terms which the paper omits. It should be noted that the numerator of the coefficient of $P_2/2$ in the second line of (27) should read

$n\theta(i+\cos n\theta) - 2 \sin n\theta$ instead of $n\theta(i-\cos n\theta) - 2 \sin n\theta$.

For a gap between two parallel plane grids, such as Black and Morton discuss, we have

$$\beta = \frac{\sin \dfrac{\theta_0}{2}}{\dfrac{\theta_0}{2}}$$

$$\gamma = \beta - \cos \frac{\theta_0}{2}$$

$$\tfrac{1}{4}\beta\gamma = \frac{\left(\sin \dfrac{\theta_0}{2} - \dfrac{\theta_0}{2} \cos \dfrac{\theta_0}{2}\right) \sin \dfrac{\theta_0}{2}}{4\left(\dfrac{\theta_0}{2}\right)^2}$$

$$= \frac{2(1 - \cos \theta_0) - \theta_0 \sin \theta_0}{4\theta_0^2}$$

$$\delta = \frac{\theta_0 - \sin \theta_0}{\theta_0^2}.$$

It can be shown that for infinitesimal ("perfect") gaps all these constants, except β, vanish and $\beta = 1$. Other types of gaps are discussed more fully in our papers on the principles of velocity modulation.

L. J. Black and P. L. Morton:[2] The discussion submitted by Messrs. Wallis and Tomlin might give an erroneous impression concerning the correctness of our derivations, particularly to one who is not familiar with the subject. The assumptions used in the derivations are clearly stated in the paper and the results are correct within the limits of these assumptions. The validity of the approximations used in the derivations is based upon the percentage velocity modulation being small, i.e., $E_1/E_0 \ll 1$. This assumption places no limitation upon the operation of the tube, since proper bunching can be obtained with a small velocity modulation if the drift-tube is sufficiently long. Under these conditions, it is not necessary to include the variation of electron velocity at the catcher in determining ρ_2. In order to obtain a reasonable efficiency, the requirements on the relative magnitudes of P_1 and P_2 are quite different. While P_1 may be small in comparison with unity, for efficient operation P_2 must be approximately unity. It is evident that the value of P_2 does not necessarily bear any direct relation to the value of P_1.

In the actual operation of klystron tubes, the operating conditions are often within the approximations used in our paper. The approximations, of course, are not valid if the ratio of buncher voltage to acceleration voltage is high. The general theory for this latter type of operation and its relation to the small-velocity-modulation theory are capably discussed by A. E. Harrison.[3]

We acknowledge the typographical error in sign that appears in (27). Fortunately the term involved does not affect any of the results.

In the expression for ρ_2 given in their discussion, the term v_2 should appear after the summation sign if the variation of electron velocity is to be included.

P. J. Wallis and S. G. Tomlin:[1] The first five pages of the paper in question treat the case of an ideal first gap and a nonideal second gap. With this part of the paper we are in agreement, having reread it carefully bearing the authors' remarks in mind, viz., that P_1 is assumed very much less than P_2.

However, we would point out that the only inequalities given in the article are $P_1 \ll 1$ and $P_2 \leqq 1$, and it did not seem to us that this excluded the case of P_1 and P_2 of comparable magnitude, which occurs with tightly coupled oscillators. Perhaps it would have been better if the authors had stated this assumption more explicitly, since other readers may draw the conclusion that we drew.

In the last page, Messrs. Black and Morton deal with transit-time effects in the first gap and appear to have done so a little too cursorily. They have neglected the fact that the change from an ideal to a nonideal gap at the buncher necessitates changes in their equation (28), which correction is not negligible unless our factor β (which in the case of the uniform gap considered in the present paper is $(\sin \theta_{0/2})/\theta_{0/2}$) is very close to 1. When,

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[2] University of California, Berkeley, California.

[3] A. E. Harrison, "Graphical methods for analysis of velocity-modulation bunching," Proc. I.R.E., vol. 33, pp. 20–32; January, 1945.

as is often the case, β is of the order of 0.5, an error which may be large is introduced.

We agree that our formula for the current density should read

$$\rho_2 = i_0 \sum \frac{1}{v_2} \frac{\partial t_1}{\partial t_2}$$

and thank Professor Black for pointing out our slip in putting v_2 outside $\sum$. Fortunately, it leaves our final result for the efficiency unaltered.

L. J. Black and P. L. Morton:[2] We are glad to note that Messrs. Wallis and Tomlin are in agreement with the first five pages of our article. We believe that they will also find themselves in agreement with the remainder of our paper upon a closer examination of (36). Actually, (36) was obtained by treating the buncher as a "nonideal gap" and the equation does contain a "gap factor β," i.e., (36) can be written in the form

$$W = \frac{P_1^2}{4} (E_0 i_0) \left(\frac{\sin \theta_0/2}{\theta_0/2} - \cos \theta_0/2 \right) \left(\frac{\sin \theta_0/2}{\theta_0/2} \right)$$

where

$$P_1 = E_1/E_0.$$

The equation written in this form is identical to "the term due to the absorption of power in the first gap $[-(P_1^2/4)\beta_1\gamma_1]$" included in their first discussion. The first part of (36) of the original paper contains a typographical error that should be obvious to the reader, i.e., in the numerator of the expression, E_1 should be written in place of E_0.

P. J. Wallis and S. H. Tomlin:[1] We would direct attention to (29) as modified by (36). The authors of the paper do not seem to have appreciated that the argument of the Bessel function in (29) must be altered when the first gap is not ideal.

Briefly, we may translate (29) into our notation as follows (taking $n=1$):

$$\eta = -\alpha_2\beta_2 J_1(\tfrac{1}{2}\alpha_1 S) \sin (S + x +) \tfrac{1}{4}\alpha_2^2\beta_2\gamma_2 \quad (29a)$$

where x is the phase difference between the field in the two gaps. Equation (36) gives a term $\frac{1}{4}\alpha_1^2\beta_1\gamma_1$, and the authors assert that the effect of the finite transit time spent in the first gap can be taken into account by subtracting this energy loss from (29), thus giving

$$\eta = -\alpha_2\beta_2 J_1(\tfrac{1}{2}\alpha_1 S) \sin (S + x) - \tfrac{1}{4}\alpha_2^2\beta_2\gamma_2 - \tfrac{1}{4}\alpha_1^2\beta_1\gamma_1.$$

Now this is not enough; the argument of the Bessel function must be altered from $\frac{1}{2}\alpha_1 S$ to $\frac{1}{2}\alpha_1\beta_1 S$ so that

$$\eta = -\alpha_2\beta_2 J_1(\tfrac{1}{2}\alpha_1\beta_1 S) \sin (S + x) - \tfrac{1}{4}\alpha_2^2\beta_2\gamma_2 - \tfrac{1}{4}\alpha_1^2\beta_1\gamma_1.$$

Or, in Black and Morton's notation,

$$\eta = P_2 \frac{\sin \theta_0/2}{\theta_0/2} J_1 \left(r \frac{\sin \theta_0'/2}{\theta_0'/2} \right) \cos \left(\alpha + \frac{\theta_0}{2} + \frac{\theta_0'}{2} \right) - \frac{P_2^2}{4} \left[\frac{2(1 - \cos \theta_0) - \theta_0 \sin \theta_0}{\theta_0^2} \right] - \frac{P_1^2}{4} \left[\frac{2(1 - \cos \theta_0') - \theta_0' \sin \theta_0'}{\theta_0'^2} \right]$$

whereas subtracting (36) from (29) and modifying the drift-tube length as suggested by (35) gives

$$\eta = P_2 \frac{\sin \theta_0/2}{\theta_0/2} J_1(r) \cos \left(\alpha + \frac{\theta_0}{2} + \frac{\theta_0'}{2} \right) - \frac{P_2^2}{4} \left[\frac{2(1 - \cos \theta_0) - \theta_0 \sin \theta_0}{\theta_0^2} \right] - \frac{P_1^2}{4} \left[\frac{2(1 - \cos \theta_0') - \theta_0' \sin \theta_0'}{\theta_0'^2} \right].$$

We have written θ_0' for θ_0 in the first gap since the two gaps are not necessarily equal.

Obviously the modification of the argument of the Bessel function makes a material difference to the result if, as may happen, $(\sin \theta_0/2)/(\theta_0/2)$ departs appreciably from unity, particularly since r can be quite large and thus emphasize any error due to omitting the gap factor. This modification of the argument can only be made by so conducting the whole discussion that account is taken of the transit time in the first gap right from the beginning. This is what we meant when we said that the transit-time effects in the first gap had been treated a little too cursorily. It is impossible to arrive at the correct result by taking an ideal first gap and finite second gap, working this to the end, and then adding a correction for the loss in the first gap. There is more in it than that; and this is what we wished to make clear.

L. J. Black and P. L. Morton:[2] We agree that the argument r of the Bessel function as defined in our paper (after (15)) must be multiplied by the factor $(\sin \theta_0/2)/(\theta_0/2)$ when transit time is taken into consideration. This should be clear from the discussion on page 482 and from (35), in which the velocity modulation is given as

$$\frac{v_1}{v_0} = \frac{E_1}{2E_0} \frac{\sin \theta_0/2}{\theta_0/2}.$$

If this value of v_1/v_2 is used, the expression for r becomes

$$r = \left(\frac{\omega s}{v_0} \right) \frac{v_1}{v_0} = \left(\frac{\omega s}{v_0} \right) \frac{E_1}{2E_0} \frac{\sin \theta_0/2}{\theta_0/2}.$$

We are sorry that we did not point out more explicitly just how the necessary corrections to (29) should be made to give the efficiency under conditions of finite transit time.

Discussion on

"Transmission of Television Sound on the Picture Carrier"*

GORDON L. FREDENDALL, KURT SCHLESINGER, AND A. C. SCHROEDER

E. R. Kretzmer:[1] The authors assume a modulation law whereby the time shift of the pulse edges is determined by the modulating signal at a number of instants equally spaced along the time axis. This law may be termed explicit, since explicit expressions exist for the times of the two edges of the pth pulse.

Among the many possible laws that may govern the shift of the pulse edges, there is one which is probably easiest to conform with in practice and which yields the simplest frequency spectrum.

With this law the time shift of a given pulse edge is proportional to the instantaneous modulating signal at the instant at which the pulse edge actually occurs. Since this time instant can be given only by an implicit relation, pulse-width modulation in accordance with this law may be called implicit modulation.

Implicit modulation is easily physically realized by adding to the signal voltage a saw-tooth type voltage of larger amplitude, clipping, and then limiting the resultant.[2] Evidently, even if a symmetrical saw-tooth voltage is used, purely symmetrical modulation cannot be attained for sinusoidal modulating voltages.

The mathematical expression for implicitly modulated pulses, placing in evidence all components present, can be found by a relatively simple analysis. It is merely necessary to obtain the Fourier series for the pulses in the absence of modulation, replace the constant width parameter by the modulation function, and expand the result.

For "symmetrical" modulation, for example, the result is

$$e = wf_c(1 - \alpha \cos 2\pi f_0 t) + \frac{2}{\pi} \sum_{M=1}^{\infty} \sum_{\nu=-\infty}^{\infty} \frac{1}{M} \left\{ J_{|\nu|}[\pi w\alpha(Mf_c)] \sin\left[\pi w(Mf_c) - \frac{|\nu|\pi}{2}\right] \right\} \cos 2\pi[Mf_c + \nu f_0]t$$

as compared to the author's equation (28) for explicit symmetrical modulation:

$$e = \frac{2}{\pi} \sum_{M=0}^{\infty} \sum_{\nu=-\infty}^{\infty} \frac{1}{M + \nu \frac{f_0}{f_c}} \left\{ J_{|\nu|}[\pi w\alpha(Mf_c + \nu f_0)] \sin\left[\pi w(Mf_c + \nu f_0) - \frac{|\nu|\pi}{2}\right] \right\} \cos 2\pi[Mf_c + \nu f_0]t.$$

The following basic changes result when implicit modulation replaces explicit modulation.

(1) The amplitude of the signal-frequency component is absolutely proportional to the product of duty cycle and modulation index. For explicit modulation this is nearly true for small values of the product.

(2) No harmonics whatever of the signal frequency are produced.

(3) A given component of frequency $Mf_c + \nu f_0$ has an amplitude which depends only on the product of duty cycle and modulation index, and which is independent of the algebraic sign of ν. The corresponding amplitude for explicit modulation (ν negative) is smaller and approaches that for implicit modulation only when the signal-to-pulse frequency ratio f_0/f_c becomes small, in which case the undesired component is likely to fall outside of the audio pass band.

It should be noted that these considerations also apply to pulse-time modulation, if width-modulated pulses are generated for detection in the receiver, as is common practice. As far as the spectrum of the time-modulated pulses is concerned, its characteristics are similarly dependent on the law which governs the time shift of the pulses.

Gordon L. Fredendall:[3] As Mr. Kretzmer states, there are various systems of pulse-width modulation. His analysis of the "implicit" type is a significant addition to the subject.

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* PROC. I.R.E., vol. 34, pp. 49–61; February, 1946.

[1] Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge, Mass.

[2] R. D. Kell, U. S. Patent 2,061,734.

[3] RCA Laboratories, Princeton, New Jersey.

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* * *

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For six years following World War I he was with the Westinghouse Electric and Manufacturing Company. Since 1926 he has been on the faculty of the University of Michigan, where he is professor of electrical engineering in charge of instruction in electronics.

From 1943 to 1945, Mr. Dow was associated with the Radio Research Laboratory at Harvard University where he was responsible for the direction of a research group involved in the Radar Countermeasures Program.

Mr. Dow is the author of a textbook, "Fundamentals of Engineering Electronics," and of several technical papers on electronic subjects. He is a member of Eta Kappa Nu, Tau Beta Pi, Sigma Xi, the American Institute of Electrical Engineers, the Society for the Promotion of Engineering Education, and the American Welding Society.

* * *

Robert Adler (A'42) was born on December 4, 1913, at Vienna, Austria. He received the Ph.D. degree in physics in 1937 from the University of Vienna, and was assistant to a patent attorney in Vienna the following year. From 1939 to 1940, he worked for Scientific Acoustics, Ltd., London, England. After one year with Associated Research, Inc., in Chicago, he joined the Zenith Radio Corporation.

During 1942 and 1943, Dr. Adler was engaged in work on high-frequency magnetostrictive oscillators. More recently, he has been active in the vacuum-tube field through his development of the phasitron system of frequency modulation.

Robert Adler

Correspondence

Cathode-Coupled Triode Amplifiers

The cathode-coupled triode amplifier described by Sziklai and Schroeder in their paper, "Cathode-Coupled Wide-Band Amplifiers," which appeared in the PROCEEDINGS OF THE I.R.E. for October, 1945, may be represented by an equivalent triode having a transconductance, plate resistance, and amplification factor as follows:

$$g' = -g\frac{G}{1+2G} \qquad (1)$$

$$r' = r\frac{1+2G}{1+G} \qquad (2)$$

$$\mu' = -\mu\frac{G}{1+G} \qquad (3)$$

where

$G = z_k g\,(1 + 1/\mu)$

g = transconductance of each of the individual triodes

g' = transconductance of the equivalent triode

r = plate resistance of the individual triodes

r' = plate resistance of the equivalent triode

μ = amplification factor of the individual triodes

μ' = amplification factor of the equivalent triode

z_k = the cathode coupling impedance.

Equations (1), (2), and (3) may be demonstrated as follows:

Referring to Fig. 1, we may write

$$i_1 = ge_g - \frac{1}{\mu}e_k \qquad (4)$$

$$i_2 = -ge_k + \frac{1}{\mu}ge_p \qquad (5)$$

$$(i_1 + i_2)z_k = e_k \qquad (6)$$

$$e_p + e_k = -i_2 z_L. \qquad (7)$$

The currents and voltages are not the total voltages and the current but are incremental values. We may eliminate e_p from this set of equations by combining (5) and (7). Thus

$$i_2 = -g\left(1+\frac{1}{\mu}\right)\frac{1}{1+\frac{gz_L}{\mu}}\,e_k. \qquad (8)$$

Recalling the identity $gr \equiv \mu$,

$$i_2 = -g\,\frac{1+\frac{1}{\mu}}{1+\frac{z_L}{r}}\,e_k. \qquad (9)$$

Substituting (4) and (9) into (6) and solving for

$$e_g = e_k\left[\frac{1+\frac{z_k}{r}}{z_k g} + \frac{1+\frac{1}{\mu}}{1+\frac{z_L}{r}}\right] \qquad (10)$$

we can write

$$e_k = (e_g + e_k)\,\frac{1}{1+\frac{1+\frac{z_k}{r}}{z_k g}+\frac{1+\frac{1}{\mu}}{1+\frac{z_L}{r}}} \qquad (11)$$

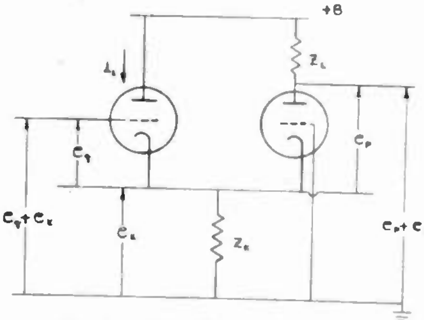

Fig. 1—Cathode-coupled triode amplifier.

Equation (9) can now be written

$$i_2 = (e_g + e_k)\,\frac{-g}{1+\left(1+\frac{z_L}{r}\right)\left(1+\frac{1}{z_k g\left(1+\frac{1}{\mu}\right)}\right)} \qquad (12)$$

The equivalent transconductance g' is i_2 divided by $(e_k + e_g)$ when $z_l = 0$:

$$g' = -g\,\frac{1}{2+\frac{1}{z_k g\left(1+\frac{1}{\mu}\right)}} \qquad (13)$$

or

$$g' = -g\,\frac{G}{1+2G}. \qquad (14)$$

The equivalent amplification factor μ' is $i_2 z_l$ divided by $(e_g + e_k)$ for very large values of z_l. Therefore

$$\mu' = \frac{-gr}{1+\frac{1}{z_k g\left(1+\frac{1}{\mu}\right)}} \qquad (15)$$

$$\mu' = -\mu\,\frac{G}{1+G}. \qquad (16)$$

The equivalent plate resistance r' is μ' divided by g':

$$r' = r\,\frac{1+2G}{1+G}. \qquad (17)$$

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Note on the Expression for Mutual Impedance of Parallel Half-Wave Dipoles

In a recent paper by the writer,[1] the expression[2] for the absolute value of mutual impedance of nonstaggered parallel half-wave dipoles, for the case where the separation between the dipoles is large in comparison with the wave length, was derived from a consideration of an expression for the induced voltage V_{12} given as the product of E_{12} or the component of the field intensity parallel to antenna number 2 and its effective height h_2. The more exact approach to the problem is as follows:

For separations $S \gg \lambda/2$, expression[3] for the component E_{12} of the field intensity parallel to antenna number 2 due to current I_1 in antenna number 1 reduces to

$$E_{12} \simeq j30I_1\left(-\frac{e^{-j\beta D}}{D} - \frac{e^{-j\beta D}}{D}\right)$$

$$\simeq -j\,\frac{60I_1}{D}\,e^{-j\beta D}$$

for when $S \gg \lambda/2$, $r_1 \rightarrow r_2 \rightarrow D$ (see Figs. 2 and 3 of footnote reference[1]).

Substituting the above expression for E_{12} in general expression for the mutual impedance[4]

$$Z_{12} \simeq j\frac{60}{D}e^{-j\beta D}\int_0^l \sin\beta y\,dy$$

$$= -j\,\frac{60}{D\beta}e^{-j\beta D}\left[\cos\beta y\right]_0^{\lambda/2}$$

$$\simeq -j\,\frac{60}{D\beta}e^{-j\beta D}\left[\cos\beta\frac{\lambda}{2} - \cos 0\right].$$

Since $\beta = 2\pi/\lambda$ and letting $S = D/\lambda$

$$Z_{12} \simeq -j\,\frac{60}{2\pi S}e^{-j2\pi S}(\cos\pi - \cos 0)$$

$$Z_{12} \simeq j\frac{60}{\pi S}e^{-j2\pi S} = \frac{60}{\pi S}(\sin 2\pi S + j\cos 2\pi S)$$

$$|Z_{12}| \simeq \frac{60}{\pi S}.$$

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[1] Kosmo J. Affanasiev, "Simplifications in the consideration of mutual effects between half-wave dipoles in collinear and parallel orientations," PROC. I.R.E., vol. 34, pp. 635–638; September, 1946.

[2] Equation (15), p. 637, of footnote reference 1.

[3] Equation (8), p. 636, of footnote reference 1.

[4] Equation (7), p. 636, of footnote reference 1.

Institute News and Radio Notes

1947 Convention News: Exhibits to Be Held at the Grand Central Palace

Plans for the 1947 National I.R.E. Convention, set for March 3 to 6, are moving ahead under the guiding hand of Dr. James E. Shepard, Chairman, and his 1947 Convention committee.

Official U. S. Navy Photograph

VICE ADMIRAL CHARLES A. LOCKWOOD

Most noteworthy development is the moving of the Radio Engineering Show from the 17th Regiment Armory to the Grand Central Palace at 480 Lexington Avenue. Only 182 booths were to be available to exhibitors at the Armory, with over two hundred booths needed to adequately display the latest in radio and electronic devices. In addition, there will be two lecture halls in which some of the technical sessions may be conducted. The move, however, does not change the character of the Convention or Show. It is not planned as a public exhibition but as a technical engineering get-together.

Highlighting the President's Luncheon (noon, March 4) will be Vice Admiral Charles Andrew Lockwood's address to those assembled. Admiral Lockwood was commander of our submarine fleet in the Southwest Pacific from May of 1942 to February of 1943, for which he received his third Distinguished Service Medal. The citation read: ". . . a daring, forceful and inspiring leader, he directed the operations of his forces aggressively in carrying the attack to the enemy with the result that the submarines under his command sank 58 enemy ships and damaged 41 others." Admiral Lockwood also wears the Legion of Merit, Commendation Ribbon, the Victory Medal, and The Order of Orange Nassau with Swords, presented to him by the Government of The Netherlands.

Guest of honor at the luncheon will be the president-elect of the Institute, Dr. W. R. G. Baker. Dr. F. B. Llewellyn, retiring president will act as toastmaster.

Planned for March 5 is the annual banquet, to be held at the Hotel Commodore.

Professor Weber reports plans for twenty-five technical sessions with a total of one hundred and twenty to be read, all hitherto unpublished in any form.

The proposed schedule includes papers on the following subjects: Electroacoustics, Frequency-Modulation Receivers, Television Theory and Practice, Commercial Navigation, the Social and Professional Status of Engineering, Vacuum Tubes, Nucleonics Instrumentation, Broadcasting Components, Recording, Electronic Control, Instrumentation, Antenna and Propagation, Medical Electronics, Basic Measurements Theory, Electronic Computers, and Systems of Communications.

An interesting addition to the agenda is the session on "The Social and Professional Status of Engineering." There will be four speakers to cover the liberal education of the engineering profession, opportunities for the young engineer, relation of the engineering profession to science and industry, and the organization of the engineering profession.

The demand for hotel accommodations will be great. It is suggested that hotel reservations be arranged for through the New York Convention and Visitors' Bureau, 233 Broadway, New York 7, New York, in order that YOU may have a comfortable place in which to spend Convention Week.

NOTICE TO THE LADIES!

The Women's Committee has arranged for a widely varied list of attractions for your entertainment at the Convention, including a reception at the new Headquarters Building at 1 East 79 Street. Other features planned by Mrs. George Gilmen, Chairman of the Women's Trips Committee, include a sightseeing tour of uptown and residential New York, Fred Waring's radio show, a tour of R. H. Macy's, a theater matinee, luncheon and fashion show at the Plaza Hotel, and a visit to the United Nations if sessions are in progress.

With the exception of the tea, our program will operate on a self-sustaining basis. Cost of the various functions will be announced later.

We are looking forward to seeing you at The Convention.

TRIXIE LLEWELLYN
Chairman, Women's Committee

INVITATION TO I.R.E. MEMBERS ACTIVE IN "NUCLEONICS"

The new field of "nucleonics" includes the study of the structure and modification of atomic nuclei, and the industrial or military applications of the information thus gained. It is well known that numerous communications and electronic engineers have been active in projects dealing with "nucleonics." The work of these engineers is understood to have involved the development of measuring instruments, control mechanisms, specialized processes, and many other extensions of the techniques with which they have long been familiar.

The Institute now cordially and urgently invites all of its members who have thus participated in "nucleonic" projects, or who plan to do so, to address a communication to that effect to the Editor, to the extent such information may be released for discussion and publication. It is hoped that they will at the same time express their interest in any Institute activities in this field, and their willingness to take part in the committee work or other contributions of The Institute of Radio Engineers along such lines. Their letters should be addressed to:

THE EDITOR
The Institute of Radio Engineers
Suite 804
597 Fifth Avenue
New York 17, N. Y.

NEW I.R.E. OFFICERS

Dr. W. R. G. Baker (A'19–F'28), vice-president of the General Electric Company's electronics department, has been elected to the Presidency of The Institute of Radio Engineers for 1947. Sir Noel Ashbridge (F'38), chief engineer of the British Broadcasting Corporation, has been elected to the Vice-Presidency of the Institute. Murray G. Crosby (A'25–M'38–SM'43–F'43), consulting engineer for the Paul Godley Company; Raymond F. Guy (A'25–M'31–F'39), radio facilities engineer for the National Broadcasting Company; and Raymond A. Heising (A'20–F'23), patent engineer for the Bell Telephone Laboratories, Inc., have been elected to the Board of Directors.

SPECIAL COMMITTEES

This list of Special Committees, which are established by other than Constitutional or By-Law provision, is an addendum to the lists of Institute Committees and Technical Committees which appeared on pages 766–768 of the October, 1946, issue of the PROCEEDINGS OF THE I.R.E.

Convention Policy Committee

J. E. Shepherd, *Chairman*
Austin Bailey
G. W. Bailey
E. J. Content
B. E. Shackelford

1947 National Convention

James E. Shepherd—General Committee—*Chairman*
Philip F. Siling—General Committee—*Vice Chairman*

Austin Bailey	General Committee
George W. Bailey	General Committee
Stuart L. Bailey	General Committee
Leo L. Beranek	Acoustic Requirements
R. D. Campbell	Hospitality Registration
E. Finley Carter	Hospitality
A. B. Chamberlain	President's Luncheon
R. D. Chipp	Facilities
Edward J. Content	General Committee
William C. Copp	Exhibit Manager
Edmour F. Giguere	Social
Virgil M. Graham	Publicity
Edna Harding	Secretary
D. D. Israel	Exhibit Requirements
F. R. Lack	Finance
E. Lehmann	General Committee
Mrs. F. B. Llewellyn	Women's Activities
George McElrath	Banquet
J. W. McRae	Printed Program
J. R. Poppele	General Committee
John G. Preston	Facilities
Henry F. Scarr	Hotel Arangements
B. E. Shackelford	General Committee
Helen M. Stote	I.R.E. Proceedings
W. O. Swinyard	Section Activities
Ernst Weber	Technical Program

International Liaison Committee

Ralph Bown, *Chairman*
F. S. Barton
E. M. Deloraine
F. B. Llewellyn

Office Quarters Committee

R. A. Heising, *Chairman*
G. W. Bailey
A. N. Goldsmith
F. B. Llewellyn
Haraden Pratt

Professional Recognition Committee

W. C. White, *Chairman*
E. F. Carter
C. C. Chambers
J. V. L. Hogan
H. A. Wheeler
H. R. Zeamans

Committee on Founders

R. F. Guy, *Chairman*

RMA Co-ordinating Committee

W. L. Barrow, *Chairman*
D. D. Israel
F. R. Lack

Piezoelectric Crystals Committee

W. G. Cady, *Chairman*
C. F. Baldwin
F. A. Barrow
W. L. Bond
J. K. Clapp
Clifford Frondel
Paul Gerber
E. W. Johnson
W. P. Mason
P. L. Smith
R. A. Sykes
K. S. Van Dyke
Sidney Warner
J. M. Wolfskill

Editorial Administrative Committee

Alfred N. Goldsmith, Editor—*Chairman*
R. S. Burnap
M. G. Crosby
E. W. Herold
Donald McNicol
Haraden Pratt
L. E. Whittemore

INTER-AMERICAN BROADCASTERS' CONGRESS

Meade Brunet (SM'33), vice-president of the Radio Corporation of America and managing director of its International Division has attended the First Inter-American Broadcasters' Congress, which convened at Mexico City. J. F. Stanbery (M'46) of the Engineering Products Section of the International Division preceded Mr. Brunet to Mexico City. Demonstrations and studies of radio, television, aviation, tube, and electron-microscope equipment were arranged under Mr. Brunet's direction.

NEW ENGINEERING FOUNDATION OFFICERS

Announcement was recently made by the Engineering Foundation of the election of L. W. Chubb (M'21–F'40), director of the Westinghouse Research Laboratories, as vice-chairman, and of the re-election of Edwin H. Colpitts (A'14–F'26), formerly vice-president of the Bell Telephone Laboratories, as director.

The Research Procedure Committee this year will be headed by Dr. Chubb, who was recently awarded the John Fritz medal and certificate.

The Engineering Foundation is one of the departments of the United Engineering Trustees, Inc., a corporation which was set up jointly by the four national engineering Founder Societies which have an aggregate membership of over 88,000. These societies are the American Society of Civil Engineers, the American Institute of Mining and Metallurgical Engineers, the American Society of Mechanical Engineers, and the American Institute of Electrical Engineers. The Engineering Societies' Library, with some 194,000 books and pamphlets, one of the largest collections on technical matters in the world, is also a department of the United Engineering Trustees, Inc.

AWARDS COMMITTEE

Three members of the Institute of Radio Engineers are among the nine persons who comprised the 1946 Awards Committee of the Television Broadcasters Association, Inc. They are: Dr. Alfred N. Goldsmith (M'12–F'15), editor of the PROCEEDINGS OF THE I.R.E.; Frederick R. Lack (A'20–F'37), vice-president of Western Electric Company; and Paul J. Larsen (A'37–M'41–SM'43), member of the Board of Governors and chairman of the Television Committee of the Society of Motion Picture Engineers. Presentation of the annual TBA Awards of Merit was made on October 10, 1946, during the second television conference and exhibition of TBA, which was held in New York.

GEORGE W. BAILEY

On November 13, 1946, George W. Bailey (A'38–SM'46) was elected a director of the Army Signal Association Post number 1, in his capacity as Executive Secretary of The Institute of Radio Engineers and as president of the American Radio Relay League.

Books

Piezoelectricity, by Walter G. Cady

Published (1946) by McGraw-Hill Book Company, Inc., 330 West 42 St., New York 18, N. Y. 787 pages+17-page index+xxiii pages. 168 illustrations. $5\frac{7}{8}\times 9$ inches. Price, $9.00.

This new and authoritative comprehensive treatise and reference volume on piezoelectricity, its applications, and the physical properties of piezoelectric crystals, will be of great value to research workers, students, and others interested in these fields.

Only a brief description of the scope of this monumental work can be attempted here. Of thirty-one chapters, the first is historical, giving an interesting background of the discovery and early development of piezoelectricity.

The next six chapters are on the physical properties of crystals, crystallography, elasticity, rotation of axes (including complete sets of equations both general and specialized), vibrations of crystal rods and plates, elastic constants, and dielectric properties. Five further chapters cover principles of piezoelectricity, special piezoelectric properties of certain crystals, production and measurement of piezoelectric effects, alternative formulations of piezoelectric theory, and secondary piezoelectric effects.

The following three chapters are devoted to piezoelectric resonators and their equivalent electric circuits and the dynamic measurement of piezoelectric and equivalent electric constants. A group of four chapters covers the properties and techniques of quartz, the quartz resonator and oscillator. Resonators of other crystals and composite resonators are briefly considered.

Considerable material, in six succeeding chapters, deals with Rochelle salt primarily, its properties, technique, observed piezoelectric and dielectric effects, and theory. Two related chapters are given to the internal-field theory of Seignette-electric crystals.

The concluding four chapters deal with miscellaneous applications of piezoelectricity, related effects such as pyroelectricity, piezo-optic, electro-optic, and other optical effects. The last chapter outlines the atomic approach to piezoelectricity. In the appendix, the Curie, Weiss, and Langevin Theories of para- and ferromagnetism are summarized for assistance in interpreting the analogous effects in Seignette-electric crystals.

The general bibliography is extensive, including fifty-seven books and 602 periodical references. Numerous additional specialized page and chapter references are given throughout the book.

No errors were noted. An obvious minor error occurs in the set of equations (19) on page 66, where primed quantities for two of the left-hand members should be unprimed.

Voigt has called piezoelectricity the most complicated branch of crystal physics. The author points out that, quite apart from atomic theory, a complete description of the piezoelectric properties of a crystal involves three types of directed quantities, the electric (field and polarization), elastic (stress and strain), and the piezoelectric coefficients by which they are related. To these, in studying the vibration of piezoelectric crystals, must be added the complexities of anisotropic materials, of many modes of vibration with many possible overtones and of boundary conditions.

The author makes it clear that "only the *results* of theoretical considerations" will be presented. Those readers having the necessary background will probably find the mathematical formulations a clear and sufficient statement; others, however, would be assisted greatly by some descriptive and explanatory material. For example, on page 43, it is not clear why differentiating the free energy (thermodynamic potential), expressed in terms of strains, would give the components of stress; or why, on page 47, a stress sytem is represented by a *symmetrical* tensor.

Considering the tremendous scope of this book, it is not surprising that a reader may find topics to his interest which are not treated extensively. A topic of importance in applications is that of the vibrations of crystal bodies. Somewhat more material on modes of vibration, coupled modes, and effects of overtones would have been desirable.

Regardless of any shortcomings in respect to particular topics, the sections on techniques and applications are valuable to those not having extensive mathematical training. General survey material in the opening paragraphs of nearly all chapters is also very readable.

J. K. Clapp
General Radio Company
275 Massachusetts Ave.
Cambridge 39, Mass.

Reference Data for Radio Engineers (Second Edition)

Published (1946) by Federal Telephone and Radio Corporation, 67 Broad Street, New York 4, N. Y. H. T. Kohlhaas, Editor. 322 pages+13-page index. 186 illustrations. $5\frac{3}{4}\times 8\frac{5}{8}$ inches. Price, $2.00; $1.60 in lots of 12 or more shipped to a single address.

As its title suggests, this book is intended to serve as a handbook of formulas, information, and numerical data useful for purposes of design in radio engineering. The material, however, is not restricted to this special field, but formulas and data relating to power engineering and audio-frequency communication problems, as well, are included.

About a quarter of the book is devoted to general information such as conversion factors, systems of units, properties of materials, logarithmic, trigonometric and hyperbolic function tables, and other mathematical data.

In short, the object of providing, in the compass of a single book of moderate size, reference material scattered through many sources is successfully achieved.

The book is, of course, not intended to serve as a text-book, but rather as a tool for the solution of design problems. For the most part, of necessity, the material has to be given in condensed form and some degree of familiarity with the subject and with fundamental electrical engineering theory is presupposed.

The data is presented in the form of tables, graphs, or nomograms. Where numerical results have to be interpolated from the graphs, an accuracy sufficient for purposes of orientation only is possible.

The present publication is a second and enlarged edition. The sections on electrical circuit formulas has been greatly enlarged to include additional material on transients, selective circuits, design formulas and tables for attenuators. The chapter on wave-guides has been revised and includes considerable descriptive material in addition to tables of useful frequency range and attenuation.

A small type for the printing is, of course, necessary, but the font adopted combines legibility and neatness of appearance. The book is light in weight, and this, taken together with its semiflexible covers, makes it conveniently portable.

Frederick W. Grover
Prof. E.E. (retired)
Union College
1036 University Pl.
Schenectady 8, N. Y.

Radio's Conquest of Space, by Donald McNicol

Published (1946) by Murray Hill Books, Inc., 232 Madison Avenue, New York 16, N. Y. 364 pages+10-page index+x pages. 69 illustrations. $5\frac{7}{8}\times 8\frac{3}{8}$ inches. Price, $4.00.

Here, presented attractively on a semi-popular level and in full sweep, is the career of radio as a technique and art from the cradle to the becoming of age in World War II. The human element is featured; the scientists, inventors, and engineers who did the building come and go. Frequently the naming of them in bunches is without discrimination, but perhaps this is as well as can be done in a relatively short book.

The succession of events is described as seen directly or through the literature by the author from the vantage point of the chair of an editor in the allied art of telegraphy. Observers of a more engineering turn of mind would place the emphasis differently, in many cases. Donald McNicol's genuine interest in radio has been attested by the helping hand he gave the I.R.E. at a critical period in its earlier days.

Radio is now older than most of us like to realize, some fifty to sixty years of age, but still not mature. Midway in this period there appeared the vacuum tube, and the book is about evenly divided between the pre- and post-vacuum-tube eras, with the earlier period the better done. Some shortcomings of the book by way of omission are: Braun's cathode-ray-oscilloscope tube; the band-pass filter of Campbell; the high-vacuum tube of Arnold and of Langmuir which, as a development of De Forest's grid-controlled tube, set the future course of events. Leaving something to be desired in respect to engineering values are the chapters on vacuum-tube development, the trend to shorter wavelengths, the problem of atmospheric noise, and several others, including the one on commercial radio, wherein the birth of broadcasting is described.

To the general reader these limitations are outweighed by the brevity and scope of the book and the wholesome human interest of the author's narrative.

Lloyd Espenschied
Bell Telephone Laboratories
463 West Street
New York 14, N. Y.

Der frequenzstabile Schwingtopf-Generator, by Arnold Braun

Published (1946) by Ag. Gebr. Leemann & Co., Stockerstrasse 64, Zürich 2, Switzerland. 79 pages. 39 illustrations. $6\frac{1}{8}\times8\frac{7}{8}$ inches. SFr. 7.50.

Über Frequenzmodulatoren für Ultrahochfrequenz, by Georg Weber

Published (1946) by Ag. Gebr. Leemann & Co., Stockerstrasse 64, Zürich 2, Switzerland. 95 pages. 35 illustrations. $6\frac{1}{8}\times8\frac{7}{8}$ inches. SFr. 9.—.

These publications are summaries of thesis work done at the Swiss Federal Institute of Technology. "Frequency-Stable Oscillator with Resonant Cavity" represents a chapter in research on the use of ultra-high frequencies for multichannel transmission purposes. It deals with single-stage frequency-stable oscillators capable of being tuned within one octave and having an output of a few watts. This subject has been covered in a very clear way, arriving at normalized values of such resonator parameters as inductance, loss factor, and Q's, in the form of graphs. This will help the engineer designing such circuits to obtain optimal values for his resonant cavities and feeding lines within a few minutes and with adequate precision.

The first chapter covers the ultra-high-frequency oscillator with feedback. The influence of the tube constants (cathode-grid impedance, anode-cathode impedance and transconductance) and that of the circuit constants on the stability are given. The influence of nonlinearities on amplitude, as well as on stability, is shown. Frequency shifts due to mechanical motion and changes in temperature, atmospheric pressure, and humidity are discussed. The voltages, relative positions of circuits, Q of cavity, reactive energy, and feedback are discussed in regard to frequency-stable operation.

The second chapter covers the problem of the resonator. First, the inductance is derived from concentric-line theory; then the resonator losses are calculated. The resistive component is readily determined by formula, and a graph of resistive component versus resonant frequency plotted for different materials and over a wide range of frequencies. Then the value of Q is given as a function of two parameters, which are plotted for various resonator dimensions. Several cases are worked out in detail to obtain highest Q for given values of resonator volume, volume and inductance, inductance and outer diameter, and inductance and length of resonator. A succeding part deals with the losses in the capacitive region of the resonator, and the resonance impedance for the whole resonator is developed. A discussion on the practical aspects of resonator construction ends the chapter. It deals with the material used, temperature compensation, and heat conduction inside the resonator.

The third chapter shows the influence of the lines between the resonator and tube. First, the influence on the tuning of resonator by the line is given. Then the Q of a system of both resonator and line is worked out. Again, graphs are plotted to obtain the total Q quickly from the line parameters.

The fourth chapter shows the construction of two oscillators and measurements thereon. Advantage of push-pull operation is shown; then the methods of measurements of the Q's and frequency shifts are discussed. The first oscillator was built with inductive and the second with capacitive feedback, both for wavelengths from 1.2 to 2.5 meters and an output of 5 watts. Detailed measurements of frequency shift as a function of loading and variation of anode and heater voltages are shown. Also, the frequency shift during warm-up and the shift over a length of one week with varying room temperatures are plotted.

"Frequency Modulators for Ultra-High Frequencies" is a contribution to the discussion of multichannel systems, and favors the system where many transmitters are modulated by a few channels, feeding into one antenna, over the system where one transmitter is modulated by many channels. Its subject is direct frequency modulation of oscillators with resonant cavities as circuits and operating between 100 to 600 megacycles.

The first chapter covers electrical modulators. Starting with calculations on reactance tubes at high frequencies, it is shown that the case with zero loss component is obtainable. It is shown that this is not possible at ultra-high frequency, and a practical case of a German ultra-high-frequency pentode type similar to the acorn tube is worked out. The limitation of frequency deviation by distortion is shown. The klystron as a reactance tube is then applied. Here a variation of beam current produces a purely reactive component. Direct frequency modulation of the klystron has the disadvantage of additional amplitude modulation for large frequency deviations. A discussion shows that velocity-modulated tubes are far superior to intensity-modulated tubes in the ultra-high frequency region from the standpoint of oscillation as well as modulation.

The second chapter covers mechanical modulators, including electrodynamic, piezoelectric, electrostatic, and electromagnetic types. Equations for maximum expansion and the equations of motion of piezoelectric rod are given. The possible frequency deviation is worked out and shown to be rather small when practical figures are applied.

A thorough theoretical investigation of the electrostatic modulator is worked out showing the equations of motion of a diaphragm and its equivalent electric circuit. The equation of diaphragm deformation with its solutions in Bessel functions is given. The diaphragm stability and the voltage limit to prevent the diaphragm from sticking to the opposite electrode are worked out. In the following discussion the factors determining frequency deviation are given, and it is shown that, for a 200-megacycle carrier frequency, approximately 50 kilocycles deviation is possible. Solutions are given for the worst possible cases to assure stable operation. Clear mathematical calculations are made and reach partly up into operational calculus for the solutions of differential equations, but the results can very well be understood by anyone with graduate training.

The second part of the book is devoted to the construction and measurement of the electrostatic modulator. It shows in detail the ultra-high-frequency oscillator with resonant cavity having an Anticorodal (aluminum alloy) diaphragm. Measurements were made on frequency deviation, distortion, and stability. A final discussion shows limitation of the modulator for a maximum of two channels having a frequency deviation of ±100 kilocycles.

Hans K. Jenny
RCA Victor Division
Lancaster, Pa.

Capacitors, by M. Brotherton

Published (1946) by D. Van Nostrand Company, Inc., 250 Fourth Ave., New York, N. Y. 104 pages+3-page index+vi pages. 33 illustrations. $6\times9\frac{1}{4}$ inches. Price, $3.00.

The author does not undertake the introduction of original material nor the compilation of a comprehensive treatise on capacitor design. Rather, the book is intended as a guide to the proper application of capacitors to electronic circuits. The stability, complex impedance, alternating-current resistance, frequency and temperature characteristics, leakage, and life of capacitors are discussed from the point of view of the engineer using capacitors as a component. The information given is necessarily general in character and is designed principally to permit the engineer to avoid misapplication of capacitor types.

The preponderance of the information in the book has to do, directly or indirectly, with paper capacitors. Only five pages are devoted to electrolytic capacitors, five pages to mica capacitors, three to variable air capacitors, and two to the ceramic types. Variable types, other than air, are not discussed.

Limited references are made to the literature, none being given with respect to mica, ceramic, or air capacitors.

The engineer using paper capacitors will find the book of interest as a quick and easily readable general guide to the characteristics and limitations of this type of capacitor. The information on other types is so meager as to be of little value.

H. C. Forbes
Colonial Radio Corp.
Buffalo 9, N. Y.

I.R.E. People

W. N. Tuttle

W. N. Tuttle

The Medal of Freedom was recently awarded Dr. W. N. Tuttle (A'26–SM'46) of the General Radio Company, Cambridge, Massachusetts. The citation reads as follows: "for exceptionally meritorious achievement which aided the United States in the prosecution of the war against the enemy in Continental Europe as head, radar subsection, Operational Analysis Section, Headquarters, Eighth Air Force, from 16 October 1942 to 3 October 1944. During this period Dr. Tuttle distinguished himself by adapting blind bombing equipment and technique for introduction into the Eighth Air Force, making a very valuable contribution to the success of numerous close co-operation attacks. His vision and untiring efforts reflect high credit upon him and were of great importance to the operations of the armed forces of the United States."

Dr. Tuttle is a graduate of Harvard University, where he received the S.M. degree in 1926 and the Ph.D. degree in 1929. He is a member of the Acoustical Society of America, the American Physical Society, and of Sigma Xi.

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Davies Laboratories, Inc. Engineering Service

The Davies Laboratories, Inc., College Park, Maryland, was recently formed as an engineering and development service to carry on and expand the work of its president, Gomer L. Davies (M'33–SM'43). C. B. Pear, Jr., (A'36) will also be associated with the new organization.

Mr. Davies received his B.S. degree from the Case School of Applied Science in 1929. Associated with the Bureau of Standards until 1933, he then became one of the original staff members of the Washington Institute of Technology, Inc. In 1945 he resigned as general manager to engage in private practice as a consulting engineer.

Mr. Pear received the B.S. degree from the Massachusetts Institute of Technology in 1939. He was engaged in work for the Blue Hill Observatory of Harvard University from 1936, with brief interruptions, until he joined the staff of the Washington Institute of Technology in 1940. At the time of his resignation there to join the new company, he was director of development.

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Alois W. Graf

Alois W. Graf (A'26–M'44–SM'45) has opened an office in Chicago for the practice of law in patent and trade-mark cases. Mr. Graf, actively associated with various engineering organizations for a number of years, has served on numerous committees of the I.R.E., and is presently chairman of its Chicago Section. He is on the executive committee of the Illinois Engineering Council, which sponsored the enactment of the Illinois professional engineer registration law, and is active in the Chicago Technical Societies Council and the National Electronics Conference.

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L. E. Bessemer

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Louis Gerard Pacent

Louis Gerard Pacent (A'12–M'15–F'27), president of the Pacent Engineering Corporation, was recently appointed a member of the Board of Examiners of the American Institute of Electrical Engineers to represent the radio engineering profession. Mr. Pacent, a well-known consulting engineer, is a Life Member of the Veteran Wireless Operators Association, and was the 1945 general chairman of the Radio Pioneers banquet. He is a Fellow of the American Institute of Electrical Engineers, Fellow and past president of the Radio Club of America, a member of the Engineers Club of New York, and of the Cosmos Club of Washington, D. C., as well as of other technical societies.

L. M. Craft

Collins Promotions

Recent promotions announced by the Collins Radio Company are those of L. Morgan Craft (M'45) from general manager of the manufacturing division to vice-president in charge of engineering and manufacturing, and of L. E. Bessemer (M'45) from chief production engineer to general manager of the manufacturing division.

Mr. Craft received the degrees in electrical engineering from the University of Illinois, and took graduate work in radio communications at Ohio State University. He was affiliated with the Bell Telephone Laboratories for two years, and in 1935 joined the Collins Radio Company, where, within a few years, he became chief engineer. Later he assumed manufacturing responsibilities and was appointed general manager of the manufacturing division. In 1944 Mr. Craft was elected to the board of directors, and in 1945 became chairman of the administrative committee. Since February he has been acting head of the engineering division.

Mr. Bessemer, who received the B.S. degree in electrical engineering from the Georgia School of Technology, was an engineer at WMAZ during his college years. He joined the Collins organization in 1935, where his first job was as wireman on a production line. He was shortly advanced to line foreman, and then to design engineer. In 1942, Mr. Bessemer was given charge of organizing the production control department, and when this was enlarged, he assumed the additional responsibilities of the industrial engineering, tool, planning, material control, and fabrication routing and scheduling departments.

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Jett Addresses Convention

Commissioner Ewell K. Jett (A'29, M'38, F'39), member of the Federal Communications Commission, addressed the 24th Annual Convention of Broadcasters in Chicago on October 22, 1946. He discussed the "Application of War Developments to Postwar Broadcasting."

I.R.E. People

LOUIS McCOMAS YOUNG

Colonel Louis McComas Young (SM'44) has recently returned to KMOX, St. Louis, Missouri, as chief engineer. He was graduated from Johns Hopkins University in 1917 with a B.S. degree in engineering.

In World War I, Colonel Young served as radio officer for the 135th Observation Squadron and the 4th Corps Observation Group. While overseas, he trained in aeronautics and radio at Oxford University and with the Royal Flying Corps in England, and at the United States Signal Corps School, Ecole Superieure D'Electricité, and the University of Paris in France. When discharged in 1919 as a first lieutenant, he was assistant radio officer for the zone of advance. Shortly thereafter, Colonel Young entered the air reserve as captain, continuing in that capacity until 1931 when he became a major. Most of his active-duty periods in the reserve were served in the Aircraft Radio Laboratory at Wright Field, Ohio.

LOUIS McCOMAS YOUNG

Upon returning to the United States, Colonel Young served for nine years as a radio engineer in the Ft. Monmouth, New Jersey, Signal Corps Laboratories, during which time he supervised the research and development division and performed work on radio equipment for use in and with aircraft. He then became associated with Westinghouse Electric Corporation, where he worked at Chicopee Falls, Massachusetts, on radio equipment for naval aviation, and was a member of the engineering staff responsible for the radio equipment on the dirigibles *Los Angeles* and *Akron*. Later, at East Pittsburgh, Pennsylvania, Colonel Young was located in the broadcast-operations department where he engaged in general radio engineering in station design and field surveys. In 1934, he joined the Columbia Broadcasting System in Chicago, and had charge of synchronizing WBBM and KFAB. He served as assistant to the chief engineer of WBBM the following year and became chief engineer of KMOX in 1940.

Called to active duty with the air reserve in 1941, Colonel Young was stationed at Wright Field and assigned to production engineering and procurement activities related to radio and radar equipment for the

SECTIONS

Chairman		Secretary
	ATLANTA January 17	M. S. Alexander 2289 Memorial Dr., S.E. Atlanta, Ga.
H. L. Spencer Associated Consultants 18 E. Lexington Baltimore 2, Md.	BALTIMORE	G. P. Houston, 3rd 3000 Manhattan Ave. Baltimore 15, Md.
Glenn Browning Browning Laboratories 750 Main St. Winchester, Mass	BOSTON January 23	A. G. Bousquet General Radio Co. 275 Massachusetts Ave. Cambridge 39, Mass.
I. C. Grant San Martin 379 Buenos Aires, Argentina	BUENOS AIRES	Raymond Hastings San Martin 379 Buenos Aires, Argentina
H. W. Staderman 264 Loring Ave. Buffalo, N. Y.	BUFFALO-NIAGARA January 15	J. F. Myers Colonial Radio Corp. 1280 Main St. Buffalo 9, N. Y.
T. A. Hunter Collins Radio Co. 855—35 St., N.E. Cedar Rapids, Iowa	CEDAR RAPIDS	R. S. Conrad Collins Radio Co. 855—35 St., N.E. Cedar Rapids, Iowa
A. W. Graf 135 S. La Salle St. Chicago 3, Ill.	CHICAGO January 17	D. G. Haines Hytron Radio and Electronic Corp. 4000 W. North Ave. Chicago 39, Ill.
J. D. Reid Box 67 Cincinnati 31, Ohio	CINCINNATI January 14	P. J. Konkle 5524 Hamilton Ave. Cincinnati 24, Ohio
H. C. Williams 2636 Milton Rd. University Heights Cleveland 21, Ohio	CLEVELAND January 23	A. J. Kres 16911 Valleyview Ave. Cleveland 11, Ohio
E. M. Boone Ohio State University Columbus, Ohio	COLUMBUS February 14	C. J. Emmons 158 E. Como Ave. Columbus 2, Ohio
Dale Pollack 352 Pequot Ave. New London, Conn	CONNECTICUT VALLEY January 16	R. F. Blackburn 62 Salem Rd. Manchester, Conn.
R. M. Flynn KRLD Dallas 1, Texas	DALLAS-FT. WORTH	J. G. Rountree 4333 Southwestern Blvd. Dallas 5, Texas
J. E. Keto Aircraft Radio Laboratory Wright Field Dayton, Ohio	DAYTON January 16	Joseph General 411 E. Bruce Ave. Dayton 5, Ohio
H. E. Kranz International Detrola Corp. 1501 Beard Ave. Detroit 9, Mich.	DETROIT January 17	A. Friedenthal 5396 Oregon Detroit 4, Mich
N. L. Kiser Sylvania Electric Products, Inc. Emporium, Pa.	EMPORIUM	D. J. Knowles Sylvania Electric Products, Inc. Emporium, Pa.
E. M. Dupree 1702 Main Houston, Texas	HOUSTON	L. G. Cowles Box 425 Bellaire, Texas
H. I. Metz Civil Aeronautics Authority Experimental Station Indianapolis, Ind.	INDIANAPOLIS	M. G. Beier 3930 Guilford Ave. Indianapolis 5, Ind.
R. N. White 4800 Jefferson St. Kansas City, Mo.	KANSAS CITY	Mrs. G. L. Curtis 6003 El Monte Mission, Kansas
J. R. Bach Bach-Simpson Ltd. London, Ont., Canada	LONDON, ONTARIO	G. L. Foster Sparton of Canada, Ltd. London, Ont., Canada
Frederick Ireland 950 N. Highland Ave. Hollywood 38, Calif.	LOS ANGELES January 21	Walter Kenworth 1427 Lafayette St. San Gabriel, Calif.

Chairman		Secretary
L. W. Butler 3019 N. 90 St. Milwaukee 13, Wis.	MILWAUKEE	E. T. Sherwood 9157 N. Tennyson Dr. Milwaukee, Wis.
J. C. R. Punchard Northern Electric Co. 1261 Shearer St. Montreal 22, Que., Canada	MONTREAL, QUEBEC February 12	E. S. Watters Canadian Broadcasting Corp. 1440 St. Catherine St., W. Montreal 25, Que., Canada
J. T. Cimorelli RCA Victor Division 415 S. Fifth St. Harrison, N. J.	NEW YORK February 5	J. R. Ragazzini Columbia University New York 27, N. Y.
L. R. Quarles University of Virginia Charlottesville, Va.	NORTH CAROLINA-VIRGINIA	J. T. Orth 4101 Fort Ave. Lynchburg, Va.
D. W. R. McKinley 211 Cobourg St. Ottawa, Canada	OTTAWA, ONTARIO January 16	W. A. Caton 132 Faraday St. Ottawa, Canada
Samuel Gubin 4417 Pine St. Philadelphia 4, Pa.	PHILADELPHIA February 6	A. N. Curtiss RCA Victor Division Bldg. 8-9 Camden, N. J.
W. E. Shoupp 911 S. Braddock Ave. Wilkinsburg, Pa.	PITTSBURGH February 10	C. W. Gilbert 52 Hathaway Ct. Pittsburgh 21, Pa.
C. W. Lund Rt. 4, Box 858 Portland, Ore.	PORTLAND	L. C. White 3236 N.E. 63 Ave. Portland 13, Ore.
A. E. Newlon Stromberg-Carlson Co. Rochester 3, N. Y.	ROCHESTER January 16	K. J. Gardner 111 East Ave. Rochester 4, N. Y.
S. H. Van Wambeck Washington University St. Louis 5, Mo.	ST. LOUIS	N. J. Zehr 1538 Bradford Ave. St. Louis 14, Mo.
David Kalbfell 941 Rosecrans Blvd. San Diego 6, Calif.	SAN DIEGO February 4	Clyde Tirrell U. S. Navy Electronics Laboratory San Diego 52, Calif.
R. V. Howard Mark Hopkins Hotel San Francisco, Calif.	SAN FRANCISCO	Lester Reukema 2319 Oregon St. Berkeley, Calif.
E. H. Smith 823 E. 78 St. Seattle 5, Wash.	SEATTLE February 13	W. R. Hill University of Washington Seattle 5, Wash.
H. S. Dawson Canadian Association of Broadcasters 80 Richmond St., W. Toronto, Ont., Canada	TORONTO, ONTARIO	C. J. Bridgland Canadian National Telegraph 347 Bay St. Toronto, Ont., Canada
M. E. Knox 43—44 Ave., S. Minneapolis, Minn.	TWIN CITIES	Paul Thompson 4602 S. Nicollet Minneapolis, Minn.
F. W. Albertson Room 1111, Munsey Bldg. Washington 4, D. C.	WASHINGTON February 10	G. P. Adair Federal Communications Commission Washington 4, D. C.
W. C. Freeman, Jr. 2018 Reed St. Williamsport 39, Pa.	WILLIAMSPORT February 5	S. R. Bennett Sylvania Electric Products, Inc. Plant No. 1 Williamsport, Pa.

SUBSECTIONS

K. G. Jansky Bell Telephone Laboratories, Inc. Box 107 Red Bank, N. J.	MONMOUTH (New York Subsection)	Lloyd Hunt Bell Telephone Laboratories, Inc. Box 107 Deal, N. J.
C. W. Mueller RCA Laboratories Princeton, N. J.	PRINCETON (Philadelphia Subsection)	A. V. Bedford RCA Laboratories Princeton, N. J.
A. R. Kahn Electro-Voice, Inc. Buchanan, Mich.	SOUTH BEND (Chicago Subsection) January 16	A. M. Wiggins Electro-Voice, Inc. Buchanan, Mich.
W. A. Cole 323 Broadway Ave. Winnipeg, Manit., Canada	WINNIPEG (Toronto Subsection)	C. E. Trembley Canadian Marconi Co. Main Street Winnipeg, Manit., Canada

Army Air Forces. In 1942, he became a lieutenant colonel, and early in 1946, advanced to the rank of colonel. When separated from the service, Colonel Young was in charge of the procurement information branch of the electronic subdivision, where he was responsible for drawings, specifications, and other technical data for the purchase of AAF developmental and production electronic equipment.

LEON PODOLSKY

LEON PODOLSKY

Leon Podolsky (A'30–M'46) has been named manager of a new field-engineering department of the Sprague Electric Company, North Adams, Massachusetts. Recognized for his outstanding wartime developmental work on Sprague Koolohm and other resistor types, he will be responsible for engineering contacts and will provide technical assistance required by the sales department. Mr. Podolsky will also have charge of the field engineering work of the Sprague Products Company, a subsidiary organization with which he was previously associated.

THEODORE A. SMITH

Theodore A. Smith (J'25–A'26–SM'45) recently was made general sales manager of the engineering-products department of the Radio Corporation of America.

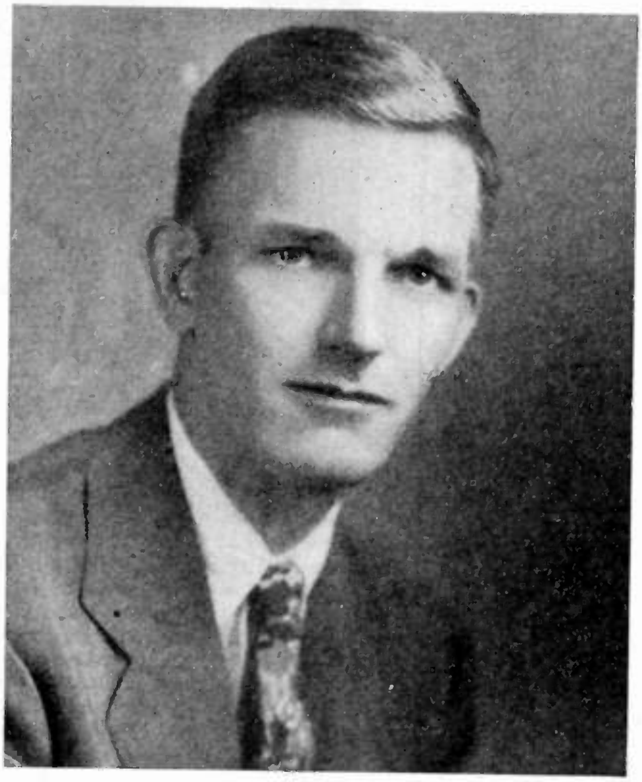
Upon receiving a mechanical engineering degree from Stevens Institute of Technology in 1925, Mr. Smith joined RCA's technical and test laboratories staff in New York. Placed in charge of television engineering in 1928, he was instrumental in building their pioneer television station, W2XBS, now WNBT, New York. In 1930 he entered commercial engineering work as RCA district sales manager for broadcast equipment, covering the eastern United States. Assigned to RCA Victor's Camden staff in 1938, Mr. Smith held key sales engineering posts. Since 1943 he has been sales manager of communications and electronic equipment of the RCA engineering-products department.

He is a member of the executive committee of the transmitter division of the Radio Manufacturers Association, and former chairman of the Philadelphia Section of The Institute of Radio Engineers.

Waves and Electrons Section

St. Louis Section Officers

Officers elected for the 1946–47 term of the I.R.E. St. Louis Section are S. H. Van Wambeck (A'40–SM'46), chairman, and R. L. Coe (A'32–M'45), vice-chairman.



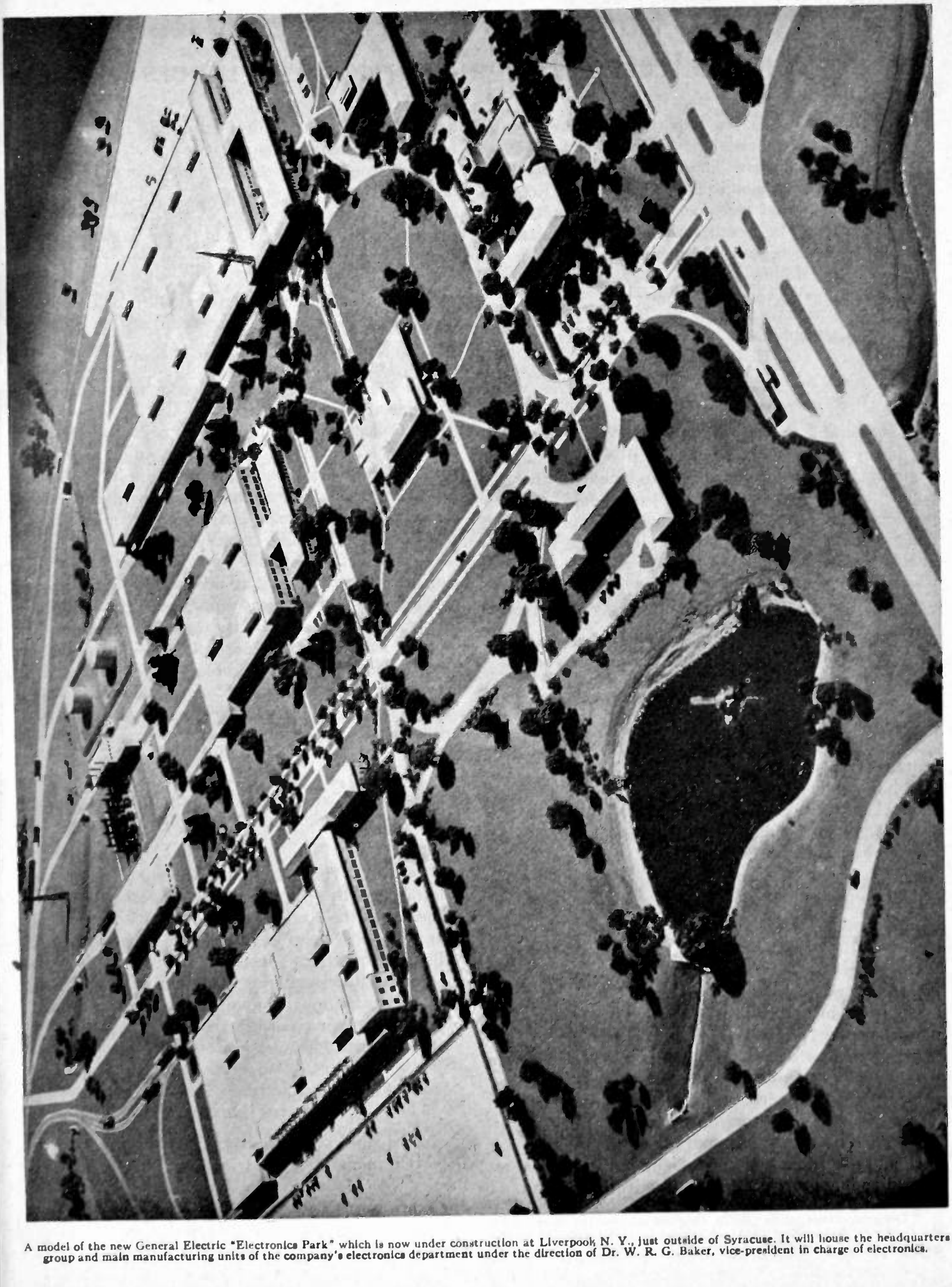
S. H. Van Wambeck
Chairman, St. Louis Section

Mr. Van Wambeck was born in Minnesota in 1910. He received the B.S. degree in mechanical engineering in 1931, and the M.S. degree in electrical engineering in 1936, from Washington University, St. Louis, Missouri. From 1933 to 1935 he was engaged in research and development work on initiating explosives and ballistic test equipment at the Western Cartridge Company, East Alton, Illinois. During 1935–36 he was sales correspondent in the special motor division of the Emerson Electric Manufacturing Company, St. Louis, Missouri. He served as instructor in electrical engineering at Oklahoma Agricultural and Mining College in 1936–37, and in the same capacity at Rice Institute, Houston, Texas, from 1937 to 1942. Here he was also acting chairman of that department in 1941–42. From 1942 to 1946 he was assistant professor in Washington University's electrical engineering department, where he is presently associate professor. During this period, Mr. Van Wambeck was also engaged as consulting engineer by the following organizations: C. H. Gurnsey and Company, Oklahoma City, Oklahoma, on rural electrification, from 1937 to 1940; Seismic Explorations, Inc., Houston, Texas, on the development and design of geophysical equipment, 1938 to 1942; Knapp-Monarch Company, St. Louis, Missouri, on problems related to the proximity fuse, among other assignments, from 1943 to present; and he was consultant on problems of instrumentation related to wind tunnel and flight test, 1942–44, for the McDonnell Aircraft Corporation in St. Louis.

Mr. Van Wambeck is a member of the American Institute of Electrical Engineers, the Society for the Promotion of Engineering Education, and of Sigma Xi and Tau Beta Pi fraternities. He was 1945–46 vice-chairman of the St. Louis Section of the I.R.E.

R. L. Coe
Vice-Chairman, St. Louis Section

Mr. Coe, a licensed radio amateur during 1914–17 and 1919–25, entered the broadcasting field in 1922. In 1924 he joined the engineering staff of station KSD, of which he became chief engineer in 1933. Joining the Army Air Forces in 1941, he served as deputy chief of staff, I Troop Carrier Command, from 1942 to 1943, and was in charge of Army Air Forces radio communications for the China-Burma-India theater in 1944. When separated from the service in 1945, he held the rank of lieutenant colonel.



A model of the new General Electric "Electronics Park" which is now under construction at Liverpool, N. Y., just outside of Syracuse. It will house the headquarters group and main manufacturing units of the company's electronics department under the direction of Dr. W. R. G. Baker, vice-president in charge of electronics.

Reports of the Committee on Professional Recognition on Collective Bargaining for Engineers*

RECOGNIZING the importance of the maintenance of professional standards and practices among engineers, The Institute of Radio Engineers appointed a Committee on Professional Recognition in 1945. The first two reports of this Committee are here presented. The membership of the Committee at the time of the acceptance of these reports was:

W. C. White, *Chairman*
E. F. Carter J. V. L. Hogan
C. C. Chambers H. A. Wheeler
H. R. Zeamans

PART I

It is desirable that the Institute have a definite present policy on the problem of collective bargaining for engineers. Such a policy at best can be only a "present policy" as conditions may be subject to rapid and unpredictable change. The most important item of such a policy is to what extent the Institute should be active in this problem.

Factors Favoring Active I.R.E. Participation

From time to time several reasons have been suggested favoring Institute action. Among these are:

(1) The attitude of "Why doesn't the Institute do something about this matter besides talk; we want action."

(2) "Employee engineers" represent a real percentage of our membership; for them collective bargaining may be of importance; therefore, the Institute should be active in this field.

(3) The Institute might lose membership and prestige if it is not active in this problem.

(4) The American Society of Civil Engineers is doing something; therefore, the Institute should or must.

Factors Against I.R.E. Participation

On the other hand, the following reasons against greater activity have been presented:

(1) It might lower our prestige and is considered by many as not a "professional" sort of activity.

(2) It divides our membership into distinct classes with different or conflicting interests.

(3) Many of the members of the Institute are officers or employees of industrial organizations. The relationship between the Institute and these organizations has been and is a friendly one. It might accordingly be regarded as inappropriate for the Institute to set up an activity that might be deemed to add to the problems of industrial management.

(4) It would be difficult to carry on in proper balance on the one hand the publication of engineering knowledge for the broad benefits of the profession, the industry, and the employer, and on the other hand projects directly seeking more compensation for the individual employee.

(5) The Institute membership and its officers include men classed as "employers" and thus at best could only sponsor or advise on such a project; otherwise the work would be "employer-dominated" and, therefore, its efforts subject to disqualification or questioning by a government agency.

(6) Collective bargaining under present government regulations must be a local activity. In many locations it may be found advisable for all engineers to band together for action. Therefore, looking to the future and for many cases, a broad engineering group must function and not just radio engineers alone.

(7) The activity of the American Society of Civil Engineers in this field has involved difficulties, complications, and some disadvantageous results, although it is generally conceded that they have approached the problem actively and intelligently and in the early and formative period of the problem they have rendered many of their members a real service.

(8) There is still a wide divergence of opinion concerning the benefits that collective bargaining might bring to professional engineers

A study of these factors leads the Committee to recommend a confirmation of the Board action of December 1, 1943, on this phase of the subject, which stated:

> RESOLVED: That, the Board does not favor the formation by the Institute, or otherwise encouraging the formation, of collective bargaining agencies for radio engineers and, therefore, the Board requests the Institute Committee on Professional Representation to report on the practicability of national action, preferably in co-operation with other technical societies, to protect and enhance the professional interests of engineers.

Courses of Action that Might be Followed by the Institute

Most engineers who have studied the subject now believe it is undesirable and impracticable to work for modification of the Wagner Act to exclude professional engineers from collective bargaining.

It is generally conceded that decisions, rulings, and events of the past year or so now greatly lessen the

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* Decimal classification: R070. Original manuscript received by the Institute, November, 1945; additional material received by the Institute, July 3, 1946.

possibility of engineers being practically forced to join a labor union against their own desires, which was the situation in several cases only a few years ago.

It is also generally agreed that the organization and operation of collective-bargaining groups for engineers is an expensive and specialized undertaking and in the near future will have to be carried on by a national organization including many types of engineers.

The Institute should take the attitude that when, in the judgment of a group of engineering or technician employees, it is to their best interest to form a collective-bargaining group, the Institute will not view such action as inappropriate. The engineers or technicians forming such a group may properly represent themselves as qualified members of the Institute but may not indicate that they directly or indirectly represent the Institute itself in collective-bargaining activities nor that the Institute represents them.

There appear to remain, therefore, two general courses of action for the Institute, which are alternative and not mutually supplemental:

(1) Limit its activities to telling its members about available information and continuing study by a committee so that, if conditions change and the engineering societies can or should play a part, the Institute has the necessary background for intelligent and expeditious action.

(2) Proceed along the lines followed by the American Chemical Society. It retained specialized counsel to advise and aid its members or groups of members in detail when confronted with the problem.

Although there is much to be said in favor of the latter procedure, several factors should be kept in mind. The subject is a constantly changing one with new rulings, decisions, and directives. No booklet or manual can long be authoritative or correct. Therefore, the work of a retained legal counsel for aid to members is of a continuing nature and expensive. It seems illogical for each of several engineering societies to parallel this work. It is almost certain that in the near future some one group will be set up to function for all engineers who desire to participate in collective bargaining.

RECOMMENDATIONS

Our committee recomends that for the present the Institute confine its activities on collective bargaining to further study and the advising of its membership of the best available information on the subject, either sent in response to requests or through publication of certain items in the PROCEEDINGS OF THE I.R.E. or both. As an appendix to this report, there is a selected list of references that engineers may consult to form a basis for a study of the problem. It also recommends that a committee continue actively to study the subject so that the Board may be informed or advised of any new developments that might make it desirable for the Institute to participate.

It is further recommended that the Institute accept or seek representation on any committee studying the problem jointly in behalf of a group of recognized engineering societies.

CONCLUSION

This report could not be better concluded than by quoting the final paragraph of a recently published article[1] on the subject:

"The important fact to recognize is that collective bargaining is here to stay and that engineers, like everyone else, must harmonize their attitude and action to that fact. The action taken by the American Society of Civil Engineers and the American Chemical Society is but one step in this adjustment. Neither society has any desire or any intent to change its basic character and function. Each stepped in because there was no organization better fitted or able to act. What happens from here on will be more and more the responsibility of the men for whom collective bargaining is a real issue. No society run by men not directly concerned with collective bargaining can long carry the ball. A way has been opened for independent action on the part of professionally minded engineers, architects, and chemists. If enough of them want such independence in collective-bargaining matters and are prepared to support a bargaining agency, they can have it. The choice is up to them, not the technical or professional societies."

The Committee wishes to acknowledge the aid of Dr. Alfred N. Goldsmith in collecting material on the subject and his many helpful suggestions.

October 31, 1945

PART II

This is a list of printed material and references selected as being of interest to members of the Institute desirous of studying this subject. None of the material listed below is available from The Institute of Radio Engineers, but where indicated may be obtained from the various organizations in question.

The following three manuals are recommended.

(1) TECHNOLOGISTS' STAKE IN THE WAGNER ACT. The National Labor Relations Act in Operation, as it affects engineers, chemists, and architects. Published by the American Association of Engineers (1944), 8 South Michigan Avenue, Chicago 3, Illinois. Price $2.00 postpaid.

This group also publishes the *Professional Engineer*, a quarterly, to help keep its manual up to date. Subscription rate, $1.00 per year.

This group offers the following for a comprehensive study of the subject:

(a)	TECHNOLOGISTS' STAKE IN THE WAGNER ACT	$2.00
(b)	The five back issues of *Professional Engineer* (a quarterly) which contain the series,	$1.25

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[1] "Where We Stand on Collective Bargaining for Engineers," by V. T. Boughton. (See reference list forming part of this report.)

"Exploded Views of the Wagner Act," supplementing "Technologists' Stake in the Wagner Act" by reporting significant decisions of the National Labor Relations Board, as handed down; 25 cents a copy.

(c) The March, 1946, (Special Issue) *Professional Engineer*, which contains our Statement to the Senate Committee on Education and Labor in re our petition for amendment of the Wagner Act; a review of the American Society of Civil Engineers plan for autonomous labor organizations of technologists, and of two test cases; Chart for Clinical Analysis of the Problems of Technologists affected by the Wagner Act. $ .50

(d) The *Professional Engineer* (two-year subscription), beginning with the June, 1946, issue which will contain a survey of the market for engineering services during the six months preceding and six months immediately following VJ Day; also a study of the problems of technologists released from the armed forces and seeking civilian employment. $2.00

Total value $5.75

Items may be ordered separately. If all the material is ordered they will send a bonus copy of "Technologists' Stake in the Wagner Act," which may be donated to a technical society, industrial school, public library, an engineering club, or a group of technologists affected by the Act.

(2) The Engineer and Collective Bargaining. An assembly of factual information regarding certain aspects of employment conditions. By Howard F. Peckworth, assistant to the secretary. Prepared July, 1943, for and sponsored by the Board of Direction, American Society of Civil Engineers, in conjunction with the activities of the Committee on Employment Conditions. This is listed as Manual of Engineering Practice No. 26 of the American Society of Civil Engineers, 33 West 39 Street, New York City, and is sold for 50 cents per copy to anyone outside its organization.

(3) Collective Bargaining for Professional Employees. January 10, 1944, page 10. This is a report issued by the American Chemical Society on the subject. Reprints are available at 17 cents each (12 cents for quantities over five) when cash is sent with order to the Mack Printing Company, Easton, Pa.

Selected Recent References

Unions of White Collar Employees. Studies in Personnel Policy No. 51 (page 6). National Industrial Conference Board, Inc., 247 Park Ave., New York City.

Illuminating Engineering. Collective Bargaining and the Engineer. By L. A. Hawkins, April, 1944, page 239.

Collective Bargaining—Does it Conflict with Engineering Ethics. By Z. G. Deutsch. *Chemical and Metallurgical Engineering*, August, 1944. Editorial on same subject in same issue.

Bargaining Position of Engineers Clarified. By W. H. Davis. *Engineering News*, November 23, 1944, page 652.

Employment Conditions for Engineers. A report of Engineers' Council for Professional Development. *Mechanical Engineering*, December, 1944, pages 809 and 812.

NLRB Forbids Status Change for Men Who Join Collective Bargaining Group. *Engineering News*, December, 28, 1944, page 807.

Engineer and His Future. By C. A. Powell. *Electrical Engineering*, January, 1945, pages 14–16.

Where We Stand on Collective Bargaining for Engineers, By V. T. Boughton. *Engineering News Record*, February 8, 1945; volume 134, page 214.

Raising Engineers' Salaries. By E. H. Robie. *Mining and Metallurgy*, March, 1945, pages 176–177.

What is Ahead in Manpower and Labor Relations. *Chemical and Metallurgical Engineering*, April, 1945, pages 120–121.

A Letter to the Editor. By Philip Hill. *Review of Scientific Instruments*, July, 1945, page 193. (Reviews progress of several independent groups of professional employees in obtaining certification by the National Labor Relations Board.)

AIEE Report on Collective Bargaining. *Electrical Engineering*, July, 1945, page 239.

American Association of Engineers Seeks Change in Labor Laws. *Engineering News Record*, October 11, 1945, page 465.

Collective Bargaining and Tapee. By E. Hanson. *Civil Engineering*, November, 1945, pages 520–521.

Personal and Personnel Problems of the Engineer. *The American Engineer*, December, 1945, pages 4–7.

Societies Jointly Study Engineers' Economic Status. *Electrical Engineering*, January, 1946, page 47.

Collective Bargaining by Engineers in Los Angeles, Calif. By W. N. Carey. *Civil Engineering*, March, 1946, pages 130–132.

Bargaining Under Existing Laws Advised. Report of American Society of Civil Engineers on Employment Conditions. *Civil Engineering*, March, 1946, pages 133–134.

Professional Engineers in Southern California Form Bargaining Units. By S. S. Green. *Civil Engineering*, May, 1946, pages 213–215.

Report on Professional Standing to the Canadian Council of The Institute of Radio Engineers*

On various occasions, the task of defining the terms "radio engineer," "communications engineer," and "electronics engineer" has been undertaken. It has remained for active and analytical groups, namely the Committee on Professional Standing of the Canadian Council of The Institute of Radio Engineers, and the Education Committee of the Montreal, Canada, Section of the Institute, to prepare the significant and stimulating report which follows. It is believed that this report will be useful to the membership of the Institute and to many other Institute Committees considering radio engineering training and qualifications. The Institute is indebted to the Canadian Council and the named Committees for their contributions.—*The Editor*

The Committee on Professional Standing has been able to collect information on and give consideration to some of the points required to fulfill its duties as outlined in Minute 20 of the April 7, 1945, meeting, but many other points, mentioned in that minute, remain unconsidered. The complete task, which was probably much too ambitious at any time, was made impossible by the general unrest attending the end of the war and by the demands of the reconversion period on the time of the committee members. The chairman, on recognizing these limitations on himself and on the committee, decided to economize time by confining the activities of the committee to those matters which somewhat paralleled day-to-day activities of the members, and in fact, because some of the committee found their time to be too completely occupied to permit any co-operation at all, the committee has given its attention only to those topics which were of interest to and could be conveniently carried on by the two of us who are active in professional training.

The information contained in this report was collected by the University of Western Ontario and is presented here with the University's permission; it is not a coincidence that this information is of importance both to The Institute of Radio Engineers and to the University but rather it is the result of a common interest in the professional education and qualification of radio engineers and radio physicists which has been awakened in both at the same time by the same set of conditions. In the case of the University, the reasons for this interest were about as follows: During the war years all Canadian universities expanded their courses in radio subjects, and as the war drew to its close, evidence accumulated that not all of the new courses should be abandoned when peace-time conditions were re-established; graduates in physics and electrical engineering who had entered some part of the electronic and communications field were critical of the background that they had been given during their undergraduate and graduate years as preparation for the work in which they had become engaged; discussions with the physics and the electrical engineering staffs of both Canadian and American universities and research laboratories led to the tentative conclusion that some revision was required in the training of radio professionals. The Institute of Radio Engineers, acting through the Education Committee of the Montreal Section, was particularly vigorous in its demands for a revised syllabus for the education of professional people engaged in the fields of electronics and communications and the report of that committee served to crystallize certain ideas which were already under consideration by the University authorities.[1] These and other evidences prompted the University of Western Ontario to make a survey to discover the nature of the training required by a radio professional, and to determine his duties; a later survey was made to extend the existing information and to add to it information on the demand by industry for people with the indicated type of training.

The fact-finding activities can be called a survey by courtesy only, since the information was collected largely by means of a series of more or less personal and usually informal letters. A tentative syllabus for a new course, which leaned heavily on the suggestions of the Montreal section, modified by many comments and discussions, was prepared and mailed, along with a letter of explanation and a request for advice, to a group of more than fifty people (mostly Canadians) who were known to have expert knowledge in some or all of the fields that the option proposed to cover. In the course of a year these discussions, by mail, had produced information which led to a complete revision of the option and this revision was again circulated for comment. The most recent information which is reported here was collected in personal interviews with the senior executives and senior technical men of seven leading manufacturers of radio equipment and of three wire companies in Eastern Canada, with the senior technical officers of the armed services, and with the executives of the government agencies interested in radio, communications, and electronics.

The task of defining the qualifications of a radio, electronics and communications engineer was given to this committee, (section (a) of Minute 20) and of using this definition to plan an educational campaign (see sections (c) and (d) of this same Minute). The committee has discussed the duties of a hypothetical radio engineer with a number of employers and teachers but has not been able to arrive at any simple definition. The problem is much like that of trying to seize on some typical feature of a new and varied landscape when seen from a high and distant place; as the attention shifts it is possible to fill the whole field of vision with one feature or another but true perspective shows each to be only a detail of the whole. In this situation the committee cannot offer a set of co-ordinates by which the typical radio engineer can be located but offers instead a map of the whole country. The committee believes that the key to the educational campaigns is to be found in this analogy: just as the intelligent selection and successful farming of a new homestead must hinge on a wide knowledge of the country and its capabilities, so must the radio engineer possess a wide knowledge of the fundamentals of science; a contrast between the pioneer homesteader and the settled farmer is here implied.

The radio, communications, and electronics industries in Canada have hired and are hiring many graduates of university engineering and physics departments, but whether these should be called radio engineers, or radio physicists, or for that matter engineers or physicists, is problematical. The positions which these graduates fill are for the most part associated with accounting, sales, and with production, positions which require a background of technical knowledge but which do not demand active use of the techniques of science. Only a relatively few find positions in design, development, and in research where they continue to use and to extend the scientific techniques which they were taught as undergraduates. If the first group are to be called radio engineers (because they practice the profession of engineering in the radio industries) then the radio engineer and radio engineering can be defined: radio engineering is a triangular domain, the corners of which have been occupied by the electrical engineer, the business man, and the physicist, and the definition follows that the radio engineer is one who combines in himself in some proportion each of the three corner elements. It is not necessary for the Canadian Council of The Institute of Radio Engineers to set up standards of technical excellence for such employees, for, in the main, they are non-technical and are selling executive ability, common sense, and a superficial technical background to a purchaser who is well qualified to judge for himself. These arguments have led this committee to interpret its duties as relating only to the second and more technical group who actively practice the profession of the radio scientist.

This committee estimates that there are about two hundred people in Canada who are engaged in design, development, and research in the fields of communications and electronics; perhaps fifty of these are employed by manufacturers and the rest are scattered among the wire companies, the Civil Services, the Armed Services and the universities. Because of new developments

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* Decimal classification: R070. Original manuscript received by the Institute, August 5, 1946. The report is dated June 25, 1946.

[1] The Report of the Montreal Section appears at the end of this report. The Montreal section acted on its own initiative in compiling its report.

there seems to be some possibility that this group may grow at an accelerated rate for some years to come; estimates collected from a diversity of sources suggest that the increase in this very specialized group may amount to as much as twenty per year, apart from the armed forces, and that a much larger figure should be used if the armed forces are included. (These figures should be regarded as no better than considered guesses.) The formal university training of the group is variegated and ranges through electrical engineering and physics, to either no training, or training in some field removed from the sciences; the number of competent men in this group who have no formal training is higher than might be expected. Many of this group have found undergraduate university training to be inadequate, and it is supposed that all, except a few, who have attended American universities, have trained themselves in radio techniques.

In order to map fully the region in which the radio engineer is to be discovered, it would be necessary to detail each type of task to which he has applied himself, but because of the magnitude of such a project only a few of the landmarks can be mentioned. If some of the landmarks are very near the boundary of the realm, it is only to call attention in a marked way to the important fact that the true radio engineer is often forced into researches in that part of pure science which until just now was the undisputed field of the physicist.

The field of radio engineering in the radio factory and in the wire companies needs little comment because it is so well known; it is mentioned to call attention to the necessity for the radio engineer to combine the functions of scientist and economist as he interprets design formula into electronic devices and services, that the public will like and buy at a profit.

The new fields of ultra-high-frequency techniques swarm with illustrations of the radio engineer who must possess much of the training of the physicist: consider the engineer who studies the work of Huygens, Fraunhofer, and Kirchhoff on the diffraction of light in order to design a parabolic reflector, or in order to predict the propagation of radio waves over a spherical earth; consider the telephone engineer who must read Lord Rayleigh on the permitted aberrations of lenses in order to produce a better radio-link system; consider the use of Fresnel's work on reflection coefficients in the choice of polarization of radar antennas for use over the sea.

Circuit analysis, as used by the radio engineer in the analysis of problems in electricity, acoustics, and mechanics, might pass unrecognized by Steinmetz; Fourier might not be able to use the table of transforms which have been published by our most prominent wire company; certainly a mathematical technician is required to apply the methods of LaPlace, Mellin, and Bromwich to the analysis of transients.

The fundamentals of physics and mathematics have reappeared in the transmission of electric power for the purposes of communication, so that the radio engineer is again closer to Kelvin than Edison; as an example, note that only a few of the applications of the transmission-line calculator are apparent, if the user has no knowledge of mapping as developed in the theory of functions of complex variables, while many of the ramifications of wave guides and lines cannot be appreciated without a grasp of Maxwell's work as well. These illustrations can be multiplied, but a few are enough to suggest that the radio engineer should be a generalized, rather than specialized, version of the electrical engineer; in fact, it is doubtful that the radio engineer is an engineer at all as the term is commonly understood, nor a mathematician, nor a physicist, but rather a harmonious combination of all three from each of which has been omitted, of necessity, that part of the superstructure which is least essential in the new field. This fact has been recognized by many other than this committee; the need for generality in background comes from the newness of the field in which the radio engineer works and from the rapidity with which new developments are being piled up; when the field has become old and stabilized, engineering in the traditional sense of application can appear.

Two of us have given a very great deal of thought to the training of radio professionals, over a number of years, and have concluded that if a course takes for its objective the general training, suggested in the previous paragraph, some radio engineers (in the narrow sense) will be produced and the rest of each class still will be acceptable in many of the positions open to the much wider group of production, sales, and executive people. The survey of industry showed that this conclusion was fairly accurate; the radio communications and electronics industry distinguishes between electrical engineers and all others trained in electricity and for the most part is willing to accept both classifications on equal terms in a broad middle ground which requires neither the electrical engineer's knowledge of construction and power engineering nor the physicist's knowledge of fundamental phenomena. The majority opinion, of the correspondents who offered advice, concerning the structure of a course to accomplish the general training of the radio engineer, agreed very closely with the suggestions of Mr. J. A. Ouimet's Montreal group. The officials of the industrial organizations who were consulted were almost unanimous in the suggestion of several further points that required consideration: radio engineers who found employment with them, they said, were often dull dogs, with few of the social graces that might well have been acquired in extracurricular undergraduate activities. (In all fairness this state of affairs probably is the result of too-heavy courses.) The writing of clear, understandable English, they claim, is an art that has been very nearly lost by technical people; business sense, in university graduates, is so rare as to be almost non-existent, and industry recommends that summer employment in industry be considered almost as part of the university course to counteract "ivory-tower" influences. The committee submits the current syllabus of the course,[2] now being taught at the University of Western Ontario, as its suggestion of the manner in which this type of training is to be accomplished; this syllabus has been compiled over some years by the two of us and other staff members, after long and careful thought and with the help and advice of many people with expert knowledge of the subject, scattered over many parts of North America; it certainly would not be improved had our thinking on this subject started at the request of the Canadian Council at the time of the formation of this committee. We can only suggest to possible detractors that the University of Western Ontario does not need the additional students which this advertising might bring, cannot accept all of those who have already applied for registration, and on the basis of the recent survey of the demand for radio professionals is now engaged in discouraging all but the most likely students from entering the profession of radio physics.

The radio designer, whose ambition for perfection is constantly thwarted by economics and tradition, will understand in what sense the syllabus is the best effort of the committee. Just as the designer must balance many things, so the pedagogue may consider ideal training, but must plan training which is economically sound; the course must be kept to four years, by eliminating fundamentals that can be safely left to after-graduation study; even tradition plays its part in determining the consumer's acceptance of the pedagogue's product. The committee considers the fact that this course is taught in a department of physics, to be an accident; and the trappings of physics that go with it to have little to do with its essential nature; provided that the fundamentals and intention remain, it will fill the existing need whether taught by physicists or engineers.

This committee, as did the one before it, suggests that the radio engineer's training rest on the fundamentals of mathematics and physics, with some of those subjects added which are peculiar to the engineering school, and which are useful for maintaining the bridge between discovery and application. Above all others, mathematics is the tool of the radio engineer; it permits him to understand both physics and engineering, and in this sense mathematics is also important to the man who on leaving the university may never again practice the techniques of his science. Only a few of the present executives of the radio industry, who were once engineers, have advised the elimination of mathematics on the grounds of waste time for one who is to practice the industrial arts; most have said that, though the mathematics is long since gone, the end results, which it taught, are still working. There has been much to criticize in the teaching of mathematics to science students in the past: the mathematician is a man apart, with his own ideas and objectives, and by his nature is out of sympathy with the objectives of the scientist; the gap between the mathematician and scientist has resulted in the rather nonmathematical state of many scientific people, especially engineers, and such a gap must be avoided in the training of the radio professional. This committee believes that the cure lies in a division of the mathematical courses between the mathematician and the scientist; let the mathematician teach the fundamentals and some of the advanced

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[2] Copies of this syllabus are available through Prof. G. A. Woonton, Chairman of the Committee on Professional Standing, 103 Huron Street, London, Ontario, Canada.

courses from his own point of view, so that the student may discover the philosophy of the mathematician and the excellent discipline of his rigor, and let the scientist teach an earthy kind of mathematics applied directly to problems in the sciences, so that the student can gain a knowledge of the manner in which the mathematician's thoughts are transformed into scientific tools. In this syllabus, physics 277b, 351a, 470a, and 470b are such courses. In example: it is the belief of the committee that, after calculus, the theory of functions of a complex variable is most necessary to a radio physicist, and for that reason physics 277b, while ostensibly an introduction to alternating-current circuits, is in fact the students' introduction to the application of complex numbers and their relation to differential equations.

It is axiomatic that a new course cannot contain all the subjects of its predecessor, and, so in planning this revised syllabus, traditional topics of physics were eliminated, not because they were not useful, but because there was no room for them; thermodynamics, properties of matter, nuclear physics, and optics are some examples of traditional courses that have been omitted. If it be recognized that very nearly the whole field of classical physics can be presented in a unified manner, in terms of energy interchanges in two classifications only, namely, those systems in which the constants are lumped and those in which the constants are distributed, then it is possible to eliminate whole courses but not to lose much in content. In example: there is no reason why the whole subject matter of physical optics cannot be taught in a course on electromagnetic theory, while discussing the reflection, refraction, and diffraction of radio waves, or during the discussions of differential equations, or while considering the ramification of Fourier analyses. Since it must be assumed that this same process of unification can be applied to engineering topics, the committee suggests that the unification of existing material in physics and engineering, as well as the introduction of new material, is essential to the planning of a new course intended to train radio physicists and radio engineers.

Even though this report is incomplete, and is intended to serve mainly as the starting point for a new committee, yet some of the suggestions that follow may be considered to be of value, and if they are, this committee urges the Canadian Council to take immediate and forcible action, for this matter is of great importance to the members for whom the Council exists.

(1) We suggest that the training of the radio engineer should be different in content and character from that of an electrical engineer. Some publicity should be given to this fact, for the similarity in designation leads to the unwarranted assumption of similarity in function.

We are convinced that a need has arisen for a type of training that is not supplied by either engineering or physics, except after long graduate training. It can be argued that modifications in traditional engineering, and in traditional physics courses, will be adequate to produce radio engineers and radio physicists who can then each fill a part of the need according to his training; this argument must wait the test of experiment, for both systems are being tried, but it should be observed that Canadian factories and laboratories hardly can afford the services of two men, if one can be taught to combine both functions in himself.

(2) We suggest that, because employers have been educated to demand a warrant of professional status from employees, and because the radio professional has not time to qualify in his own field and as a physicist, or an engineer, as well, the Canadian Council of the I.R.E. should make every effort to secure professional recognition for the radio engineers and radiophysicists.

We recommend that the Canadian Council:

(a) Obtain permission from The Institute of Radio Engineers to organize a Canadian Examining Body to pass on the qualifications of applicants for membership in the Institute, in much the same manner as the British Institution of Radio Engineers. It is possible that, though this procedure would have no legal standing, employers can be educated to accept membership in the Institute as a mark of standing in the profession.

(b) Open negotiations with the Canadian Association of Professional Physicists, for the purpose of exploring their attitude toward membership for radio professionals who have pursued a course of study according to a syllabus containing the essentials suggested by the Montreal committee. It must be made clear that the course content and not the university department (engineering, physics, or other) is the determining factor, for it can be argued that a graduate of such a course, with little extra specialization, can enter either the fields of physics or engineering at will and so is eligible for membership in this association.

(c) Open negotiations with the Provincial Associations of Professional Engineers for the purpose of exploring their attitude toward a revision of their syllabus so as to include a category for graduates of the type which is here discussed. It is suggested that, for radio engineers, the standards as laid down for some topics in their syllabus might be raised, that some new topics might be introduced and some existing topics eliminated. Of the two practice examinations of the Ontario Association, Examination II is unsuitable for radio professionals in that the training required is neither fundamental nor pertinent to the practice of radio engineering.

The members of the Committee of Professional Standing of the Canadian Council of The Institute of Radio Engineers are as follows:

G. A. Woonton, Chairman, Research Professor, Department of Physics at the University of Western Ontario, London, Ontario.

R. C. Dearle, Professor of Physics and head of the Department at the University of Western Ontario, London, Ontario.

K. R. Patrick, formerly Wing Commander in the Royal Canadian Air Force and Commanding Officer of the R.C.A.F. radar school at Clinton, Ontario; now manager of the Engineering Products division, RCA-Victor Co., Ltd., 1001 Lenoir Street, Montreal, P. Q.

R. C. Poulter, Director of Education, Radio College of Canada, and president, Poulter Publications, Ltd., Toronto, Ontario.

REPORT OF EDUCATION COMMITTEE (1943–1944)
MONTREAL SECTION
THE INSTITUTE OF RADIO ENGINEERS

This is to report on the preliminary study made by your Committee on the adequacy of present Electical Engineering courses in terms of the specific needs of engineers entering the communications and applied electronics fields.

Although the investigation undertaken by your committee is by no means complete, it has already revealed certain facts which are considered of such importance as to warrant immediate consideration.

1. The electrical engineering industry can be divided into two broad groups:

(a) *The "Power" group*, which comprises industries and operating companies concerned with the manufacture and operation of the equipment for the generation, transmission, distribution, and utilization of electric power. This group may be said to have reached maturity in its development.

(b) *The "Electronics" group*. The term electronics is used here to designate this group only for the sake of simplicity of reference and for lack of a more adequate term. This group comprises the electrical communication industries and operating companies—telephone; telegraph; radio, including ultra-high frequency, frequency modulation, television, telephoto; sound pictures and public address; radiotherapy, and other industries engaged in the field of applied electronics such as the production of industrial-control and radio-frequency-heating equipment.

Although relatively young, this group of industries has had such an amazing development in the past decade that it now occupies in magnitude and importance one of the leading positions amongst the various fields of engineering endeavor. The extremely vital role played by these industries in the present conflict is an indication of the possibilities for future expansion under peacetime conditions.

2. Although these two groups both derive directly from general electrical engineering, there is a radical difference between them in terms of the fundamental phenomena involved.

(a) In terms of frequencies used: The "Power" group is concerned mainly with one or two extremely low frequencies—25 and 60 cycles per second. Even considering the harmonics of these frequencies, the limit is perhaps 300 cycles per second. The "electronics" group on the other hand deals with a practically unlimited spectrum of frequencies ranging from the lowest to the infrared regions.

(b) In terms of circuits: The "Power" group is mainly concerned with lumped and bilateral circuit elements, while the "electronics" group deals also with circuits having distributed constants and unilateral devices and in which electron tubes and resonance phenomena play a major role.

(c) Finally, the "Power" group is concerned mainly with the conduction phenomena in physical circuits while the "Electronics" group has also to deal with the motion of electrons in gases and vacuum and the propagation of electromagnetic waves.

3. Because of the vastly different range of phenomena involved in the two groups, it is only normal to expect that the educational requirements of the two groups would also be widely different. However, it is found that electrical engineering courses now given in our universities have been designed primarily for the needs of the "Power" group. Few engineering schools give more than a rudimentary training in the "Electronics" field. Even the "Communications" options provided in some curricula are essentially modified "Power" courses which maintain emphasis on "Power" subjects and approach.

4. Because of the facts given above, it is the unanimous opinion of engineers of the "Electronics" group whom we have consulted that the present electrical engineering courses are entirely inadequate for their specific requirements.

It is also their unanimous opinion that their requirement can only be met by the establishment of a distinct and separate syllabus of courses designed specifically for "Electronics" students in the third and fourth years.

5. Although it is fully realized that the development of an appropriate syllabus for an adequate course in "Electronics" can be undertaken only by experienced engineering educators, the experience of the industry sets down certain broad principles which should be followed.

For convenience, these principles have been shown below in the form of a table which indicates the natural differences between the educational requirements of the "Power" and the "Electronics" groups.

Although this report touches only the highlights of the problem, your committee feels that it could be brought with advantage to the attention of University authorities.

The 1943–1944 Education Committee of the Montreal Section of The Institute of Radio Engineers was composed of the following members:

J. A. Ouimet, Chairman, associated with the Canadian Broadcasting Corporation.
J. E. Hayes, associated with the Canadian Broadcasting Corporation.
G. E. Sarault, associated with Laval University.
H. H. Schwartz, associated with Dee Electronics Ltd.
A. M. Patience, associated with RCA-Victor.
S. Knights, associated with the Canadian Marconi Company.
J. G. Bernier, associated with Ecole Polytechnique.
K. R. Patrick, associated with RCA-Victor.
R. A. Chipman, associated with McGill University.

Subject	*Present "Power" course*	*Recommended "Electronics" course*
1. Mathematics	Relatively elementary. Includes differential equations.	Much more mathematics needed to permit proper study of high-frequency phenomena and wave propagation. Should include training in the theory and application of advanced differential equations, determinant matrices, vector analysis, functions of a complex variable, Bessel functions, etc.
2. Physics	There exists in the "Power" courses an apparent lack of co-ordination between physics and engineering courses.	In "Electronics" (and this should also be the case in "Power" courses) physics must be an integral part of the engineering curriculum, for the basic physical laws are the foundation on which the whole electronics structure is built.
3. Electrical Engineering Theory and Laboratory	The "Power" approach is to teach simplified theory and to study phenomena which are important at "Power" frequencies. Apart from the study of direct-current and low-frequency alternating-current flow, most of the emphasis is placed on the study of large rotating machines, large transformers, and electrical short transmission lines.	In "Electronics" simplified "power" concepts do not generally apply and often (because of the difference in frequency) are actually erroneous. Modern network and electromagnetic theory are an essential part of the equipment of the "Electronics" engineer. In "Electronics" the main emphasis must be placed on circuits, thermionics, audio- and radio-frequency equipment. Small motors, small transformers, etc., form a part of such equipment and are thus of secondary interest.
4. Electrical Measurements and Laboratory	In "Power" courses, the emphasis is naturally largely on measurements at "Power" frequencies.	In "Electronics" a very special laboratory technique must be taught and this technique differs in the different frequency bands—audio, radio, and ultra-high frequencies. Measurements, therefore, becomes a major subject.
5. Mechanical Engineering and Laboratory, Machine Design Thermodynamics	These auxiliary courses are designed in the "Power" courses to supply to the "Power" engineer the fundamentals of mechanical and heat engineering needed for their particular applications in power industries. Thus, emphasis is placed on large diesel and gasoline engines, turbines, boilers, etc.	In "Electronics" these "Power" applications are of little importance and hence these auxiliary courses should be modified, with emphasis on small mechanisms and machines and on industrial processes used in the "electronics" industry such as welding, metal finishing, plating, etc.
6. Characteristics of Materials.	Very little information is given in the average "Power" course as to the materials used in industry.	In an "Electronics" course, the study of electrical and mechanical properties of common materials is essential. There should be included a study of metals, alloys, plastics, ceramics, rubbers, lacquers, abrasives, textiles, and fibers, etc., commonly used in "Electronics" industries.
7. Mechanical Drawing	Although the courses in mechanical drawing are given in the early years, emphasis appears to be entirely upon mechanical design.	In "Electronics" courses (also desirable in "Power" courses), teaching of drawing should be given with emphasis on electrical plans and schematics. It is also felt by many that the time actually given to drawing is much more than is necessary and that such training in drawing as is given to the engineering student should be largely free-hand sketching. Professional draftsmen, trained in technical schools, are more economically employed in industry for this work of making detailed drawings.
8. Machine Shop	The present curricula of "Power" courses gives little or no opportunity to the student to become familiar with the use of modern hand and machine tools. Frequently the engineering graduate is not able to use even a hand drill or a soldering iron with any facility.	Although it is not the function of universities to teach "trades," a model shop should be available to students under the guidance of experienced craftsmen. It is very important that the "Electronics" engineer be able to use ordinary hand tools and understand the use of modern machine tools.
9. Engineering Acoustics	This subject being of no particular interest to "Power" engineers, acoustics is not included in the syllabus.	Because of the importance of sound in communications industries, a course in "Engineering Acoustics" is essential for the "Electronics" group.

Two Multichannel Microwave Relay Equipments for the United States Army Communication Network[*]

RAYMOND E. LACY[†], SENIOR MEMBER, I.R.E.

Summary—Radio Set AN/TRC-5 and Radio Set AN/TRC-6 are described. An explanation is given of radar-type pulse-time-division methods of modulation which eliminate the relatively cumbersome carrier-frequency terminal equipment usually associated with a multiplex system. Audio performance and radiation characteristics are shown which presage a microwave epoch for long-lines communication.

I. INTRODUCTION

THE DEVELOPMENT of the equipment described in this paper was initiated by the United States Army Signal Corps Engineering Laboratories early in 1943 to provide transportable multichannel radio-relay sets for interconnection with high-grade voice circuits. The program was classified as secret for military reasons based upon the use of relatively unexplored frequency spectrums and newly applied circuit techniques. Declassification was accomplished after the equipment had been utilized for a period of time in the European Theater of Operations.

Since the introduction of wire communications equipment in the field of operations, a need has been evident for means of bridging streams and other obstacles, restoring service on communication lines broken due to enemy action, and providing communication traffic-handling facilities to rapidly moving armies. The first application of radio-relay equipment to this military contingency was demonstrated in the North Carolina maneuvers in 1941.[1] With the advent of multichannel wire equipment such as the carrier-telephone equipment CF-1, this need became more urgent because failure of wire equipment interrupted more than one communication channel. Early in 1942 investigations were started on means of transmitting the complete CF-1 frequency band of approximately 12 kilocycles over existing army radio equipment. Some of the schemes were successful, but at best they were only makeshifts. Highly directional, easily transportable antennas were desirable, but the existing state of the art of microwave techniques necessitated a stopgap radio design which could be used with existing spiral-four equipment and operate at a frequency for which components were immediately available. This design became the 70- to 100-megacycle frequency-modulated Radio Set AN/TRC-1,[2,3] later supplemented by an improved Radio Set AN/TRC-8 working at the higher frequency of 230 to 250 megacycles.

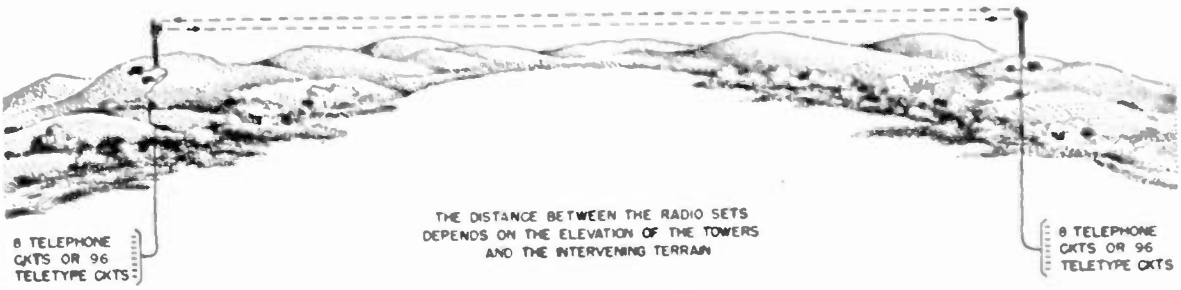

Fig. 1—Perspective of one radio link or jump.

Information was received in 1942 concerning the development of a British communication equipment, an eight-channel, 6-centimeter radio-relay set which incorporated pulse modulation and obtained eight voice-frequency channels with a time-division-multiplex method. Representatives from Signal Corps Engineering Laboratories and commercial laboratories were sent to England to investigate this development. This equipment was basically radar-modulation techniques applied to communication to provide multichannel facilities in the microwave region. It was decided to continue a research project on similar equipment in this country. It was further decided to continue research in the neighborhood of 25 centimeters, in addition to the 6 centimeters utilized by the British, to insure, if the propagation characteristics of the shorter wavelengths proved to be unsuitable for military applications, that equipment would be available on longer wavelengths. The choice of eight audio channels was made on the premise that this equipment would probably be used with the British set and/or with the CF-1 carrier-current equipments, each of which provided four audio channels. Two carrier equipments could therefore continue the facilities of one of the proposed radio-relay systems, or vice versa. Radio

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[*] Decimal classification: R480×R561×R310. Original manuscript received by the Institute, January 18, 1946; revised manuscript received, April 22, 1946; second revision received, August 9, 1946. Presented, 1946 Winter Technical Meeting, New York, N. Y., January 26, 1946.

[†] Signal Corps Engineering Laboratories, Coles Signal Laboratory, Red Bank, N. J.

[1] Roger B. Colton, "Army ground communication equipment," *Elec. Eng.*, vol. 64, pp. 173–179; May, 1945.

[2] William S. Marks, Jr., Oliver D. Perkins, and Willard R. Clark, "Radio-relay communications systems in the United States Army," PROC. I.R.E., vol. 33, pp. 502–503; June, 1945.

[3] "Multi-channel army communications set," *Elec. Ind.*, vol. 4, pp. 90–93; June, 1945.

Set AN/TRC-5, operating in the frequency range of 1350 to 1500 megacycles and designed by RCA Laboratories, and Radio Set AN/TRC-6, operating in the frequency range of 4350 to 4800 megacycles and designed and produced by the Western Electric Company, both under contract to Coles Signal Laboratory (the Signal Corps Communication Laboratory), are the equipments resulting from this program.

Each Radio Set AN/TRC-5 and Radio Set AN/TRC-6 consists of a combined transmitter and receiver with multiplex arrangements for providing eight two-way radiotelephone channels between two points. They may be used as part of a field communication network which may include other facilities, such as sections of wire lines or carrier-telephone systems. The sets are used in pairs, one set being used at each end of an unobstructed line-of-sight radio-transmission path. Two sets so used are shown in perspective in Fig. 1. The average working range of such a pair of sets is about 25 to 50 miles, depending on the elevations available at the sites of the radio sets. Longer circuits can be set up by placing radio links end-to-end in tandem, as shown in Fig. 2. The length limitation of such a tandem arrangement is dependent primarily on maintenance rather than on electrical characteristics. Electrically, upwards of 4000 miles has been experienced with no errors in teletypes or appreciable impairment of voice intelligence. The eight telephone channels can be made an integral part of the nation's regular telephone networks. As far as the user is concerned, the operations involved are identical with normal telephone use. The equipments provide complete communication systems, as they include ringing and level-setting facilities as well as the multiplexing-terminal circuits integral with the basic units. The audio characteristics of the individual channels are such that, with the use of auxiliary teletype terminal equipment, 4 to 9 two-way or 8 to 18 one-way teletypes, dependent upon the particular terminal apparatus utilized, can be communicated over each of the eight audio channels. Facsimile can be sent with equal case. If each of the eight audio channels is thus used with teletypes, a relatively tremendous volume of communication is possible over such a system which can provide 144 one-way or 72 two-way teletypes.

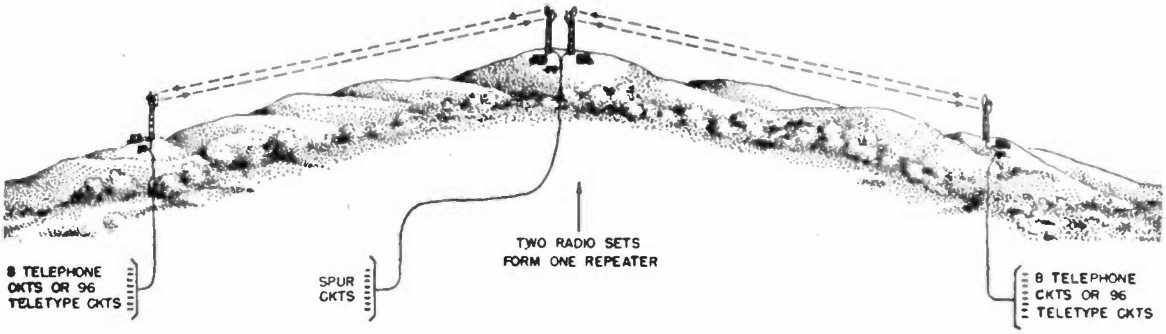

Fig. 2—Perspective of two radio links in tandem.

Radio Sets AN/TRC-6 were introduced into the European Theater in December, 1944, and the response to their performance was most enthusiastic. The first installation, in January, 1945, provided telephone facilities between General Bradley's 12th Army Group and 15th Army Group. When more equipment arrived in April, circuits were installed between General Bradley's Headquarters at Weisbaden and General Patton's Army, and between 6th Army Group and 7th Army. As General Patton advanced into Germany, close communications between his army and the 12th Army Group were maintained by Radio Sets AN/TRC-6, which were installed and operated by soldier operators under combat conditions at each new location to which General Patton moved. The last installation was made in the vicinity of Munich. The total distance by radio from 12th Army near Frankfurt, Germany, to General Patton's Headquarters at Bad Tolz in the Bavarian Alps was nearly 300 miles, using two terminal and three relay stations. A circuit of approximately the same length was installed in 12th Army Group through the 6th Army Group in Heidelberg to the 7th Army, located approximately 50 miles northwest of General Patton's 3rd Army. During the winter months, supplies were dropped from airplanes to snow-bound relay stations in the Hartz Mountains. The first installation of the AN/TRC-6 on the east side of the Rhine was made shortly after the taking of the Remagen bridgehead. After the surrender, the AN/TRC-6 was used to link General Eisenhower's Headquarters in Frankfurt to the Allied Control Commission in Berlin. In the Pacific, Radio Set AN/TRC-6 has seen only limited use in Hawaii, but it is expected that a limited number will be used in the occupational zones. Production of Radio Sets AN/TRC-5 had just started at the Rauland Corporation when the war ended, so none of this equipment has been used in the theaters. It was anticipated, however, that it would also be used by the occupation forces.

II. Mechanical Description

Radio Set AN/TRC-6 is illustrated in Fig. 3 and Radio Set AN/TRC-5 is illustrated in Figs. 4 and 5. When installed for operation, all the operating units of the AN/TRC-5 equipment, shown in Fig. 4, are mounted on the ground. The AN/TRC-6 is installed similarly except that the transmitter and the receiver converter are mounted on the antenna tower directly behind the paraboloidal reflectors (see Fig. 3). This difference is occasioned by the much higher frequency of the AN/TRC-6, but even with the lower frequency of the AN/TRC-5 there is approximately a 6-decibel loss in radio-frequency power in the 65 feet of coaxial cable utilized between the equipment and the antenna radiators.

Fig. 3—Radio Set AN/TRC-6 including stand-by (spare multiplex and radio) units. Note that the radio-frequency units are mounted on the antenna reflectors.

Both the AN/TRC-5 and the AN/TRC-6 are designed to provide 24-hour service similar to that supplied by existing commercial communication systems. For this purpose, various units are in duplicate. This provides stand-by spares which can be put into operation to replace any defective units. Whole major units are so utilized in each set, except for the multiplex unit of radio set AN/TRC-5 wherein a finer breakdown is provided. Here, instead of providing a complete spare multiplex unit, it is merely necessary, because of the type of circuit design, to supply spare transmitting and receiving common and channel plug-in units. This method of replacing defective units has the advantage, beside spare-unit weight reduction, of allowing one defective channel unit to be replaced without communications through the other seven channels being interrupted even momentarily.

III. Electrical Description

Principles of Pulse Transmission

Both radio sets make use of what is commonly called "pulse transmission"; both have a multiplex unit which translates each of the voice-frequency signals of the eight channels into a pulse-position-modulated, time-division signal for transmission over the radio links. In the case of the AN/TRC-6, the multiplex unit generates a 4-

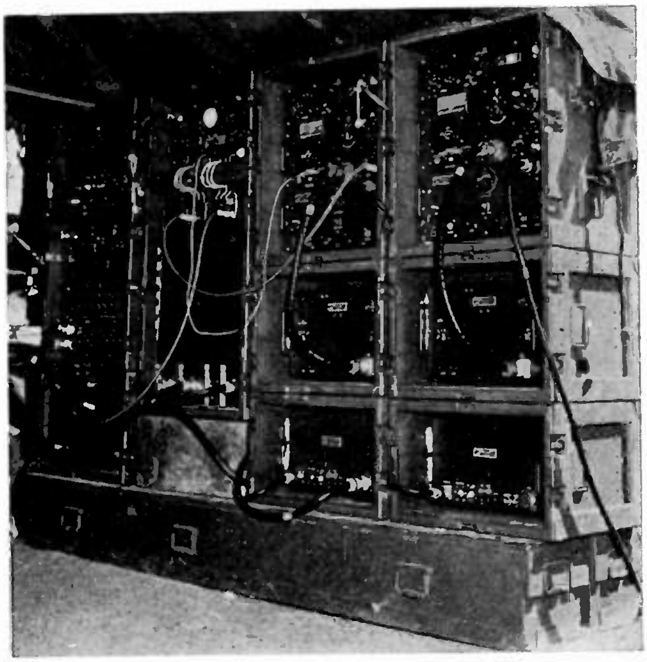

Fig. 4—Radio Set AN/TRC-5 including stand-by units. The antenna and support are shown in Fig. 5.

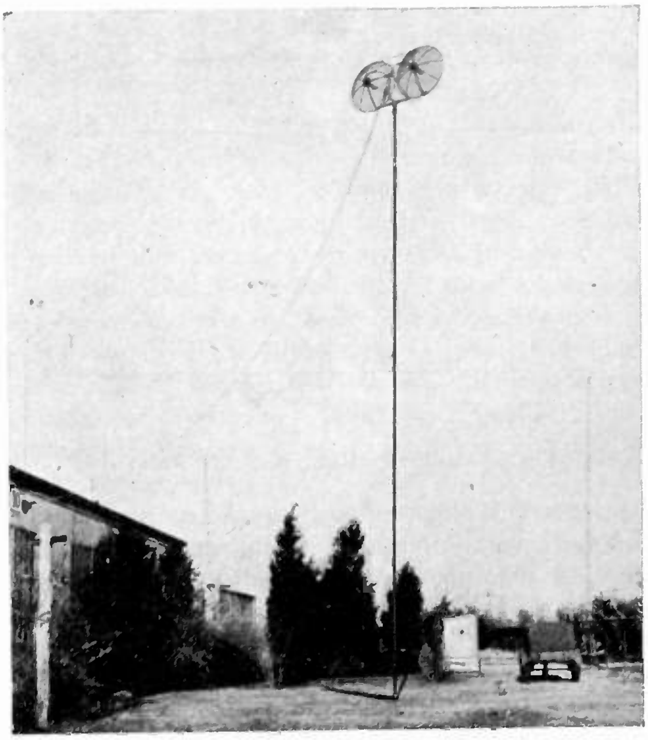

Fig. 5—Antenna for Radio Set AN/TRC-5. Note that the radio-frequency energy is fed to and received from the transmitting and receiving dipoles directly by coaxial cables.

microsecond synchronizing or marker pulse (see Fig. 6) at a recurrence frequency of 8000 cycles per second, providing a space of time of 125 microseconds between the successive marker pulses. Evenly spaced between the marker pulses are eight 1-microsecond channel pulses. The eight channel pulses and their associated marker pulses are referred to as a frame; thus each frame is 125 microseconds in length. The position of each channel pulse with respect to the marker pulse can be varied during succesive frames at a rate determined by the frequency of the modulating signal; the amount that the channel pulse is displaced from its normal position is dependent on the amplitude of the modulating signal. The modulation of channel 2 is illustrated in Fig. 7 for a constant-amplitude input signal of 1000 cycles. With no input signal, or when the input-signal voltage passes through 0 (points *a*, *e*, and *i* of Fig. 7(a)), the channel pulse will occur at the middle of the position allocated to the channel (see Fig. 7(b)). If the pulse occurs while the 1000-cycle signal is positive, the channel pulse will occur at a slightly later time with respect to the marker pulse, as represented by the two dotted pulses to the right of the mid-position. Likewise, if the pulse occurs while the input voltage is negative, the channel pulse will occur at a slightly earlier time with respect to the marker pulse, as represented by the two dotted pulses to the left of the mid-position. The distance from the mid-position at which the pulse occurs is proportional to the amplitude of the input signal; that is, maximum amplitude of input voltage produces maximum displacement of the channel pulse. Since the space allocated to a channel is fixed, the maximum amplitudes of the input signal must be limited. Accordingly, each channel is provided with a voltage-limiting circuit arrangement to prevent channel overlapping which would produce cross talk between channels. Each channel occupies only a small fraction of the total time. From this feature the term "time-division" signal is derived; that is, the total time is divided among the several channels.

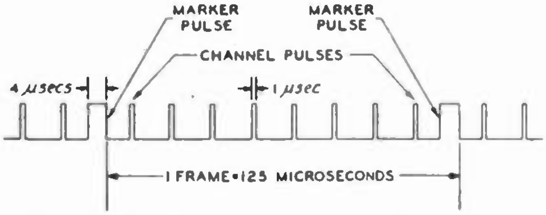

Fig. 6—Pulse-time-division signals with markings pertaining to radio set AN/TRC-6.

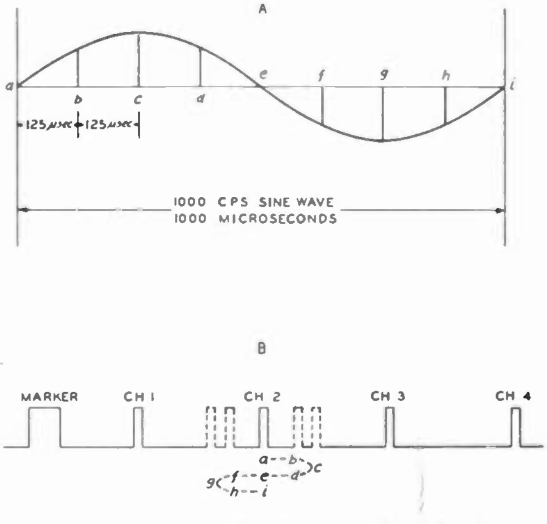

Fig. 7—Position modulation of a channel pulse.

The 1-microsecond unmodulated channel pulses are spaced about 15 microseconds apart but when modulated may only occur at any time in a 12-microsecond period per pulse, depending on the amplitude of the voice-frequency signal. The difference between 15 and 12 microseconds gives the separation between pulses allowed for preventing cross talk between channels. In any frame of the 125 microseconds, the total time during which all pulses are transmitted is only 12 microseconds (8 microseconds for the eight channels and 4 microseconds for the marker). The ratio of 125/12, or about 10, is referred to as the "duty cycle," and is defined as the ratio of the total time per frame to the length of time that the transmitter is operating during that frame.

The highest frequency of the input signal in each channel is limited by a low-pass filter to about 3000 cycles, thus eliminating the recurrence frequency of 8000 cycles.

Technical Characteristics and Performance Curves

The technical characteristics of Radio Sets AN/TRC-5 and AN/TRC-6 are as follows:

TABLE I

	AN/TRC-5	AN/TRC-6
Transmitter power, average	30 watts	200 milliwatts
Transmitter power, peak (duty cycle times average)	600 watts	2 watts
Type of antenna	dipole radiator, paraboloidal reflector	horn radiator, paraboloidal reflector
Net gain in beam direction (referred to a dipole)	14 decibels	30 decibels
Antenna beam width	15 degrees at half-power points	3½ degrees at half-power points
Average range of single radio link	25 to 50 miles	25 to 50 miles
Power source	115 or 230 volts ±15 per cent, 50 to 60 cycles	115 or 230 volts ±15 per cent, 50 to 60 cycles
Power drain (without spares)	1500 watts	1500 watts
Frequency range	1350 to 1500 megacycles	4350 to 4800 megacycles
Receiver intermediate frequency	16 megacycles	60 megacycles
Channel pulse length	0.4 microsecond	1 microsecond
Synchronizing pulse length	2 microseconds	4 microseconds
Repetition rate (frame)	10 kilocycles	8 kilocycles
Net duty factor	5.2 per cent	10 per cent
Antenna-support height	50 feet	50 feet
Voice-frequency range	300 to 3000 cycles	300 to 3000 cycles
Voice-frequency input impedance	600 ohms	600 ohms
Voice-frequency input connections	2-wire or 4-wire	2-wire or 4-wire
Ringing input	20 to 60 cycles	20 to 60 cycles
Ringing output	20 cycles	20 cycles
Radio-frequency channel spacing	30 to 70 megacycles	150 megacycles
Receiver noise figure	12 decibels	20 decibels
Intermediate-frequency band width (3-decibel-down points)	3 megacycles	10 megacycles
Wavemeter accuracy	0.07 per cent	0.02 per cent
Ringing sensitivity	13 to 18 volts	25 volts
Image rejection	28 decibels	negligible

From the communication-systems viewpoint, various other characteristics of this equipment, in addition to the noise level, the effect of which is explained later, are relatively important. Fig. 8 shows a load or "peak-limiting" characteristic curve which is of especial interest when applying tone teletypes, with their necessary filter requirements, to a relay system. Radio Set AN/TRC-6 has no loss due to peak limiting up to an input of about 4 decibels based on a zero level of 1 milliwatt, but from then onward it has approximately 4 decibels loss for 10 decibels input, 8 decibels loss for 15 decibels input, etc. Radio Set AN/TRC-5 has a similar requirement whereby 0 decibels loss is introduced by limiting for −5 volume-units speech input, 1½ decibels loss for 0 volume-units speech input, 3½ decibels loss for +5 volume-units input, 6½ decibels loss for +10 volume-units input, etc.

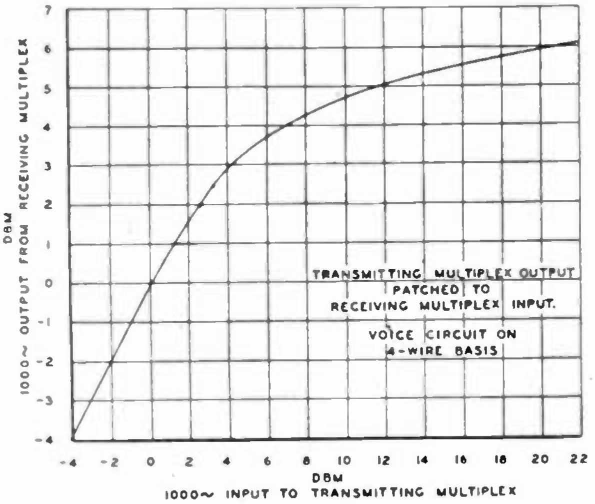

Fig. 8—Load characteristic of a typical channel of Radio Set AN/TRC-6.

Figs. 9 and 10 illustrate channel audio-frequency-response curves. They are relatively flat for one link over a frequency range of approximately 300 to 3000 cycles. Systems involving these equipments have operated in the field for continuous periods of time ranging from 24 hours to several days, and a check of their audio-channel levels has shown no measurable drift or, at worst, no more than a ½-decibel change per link. Of course, any deviations of this characteristic curve from being exactly flat will contribute to or may even tend to cancel system deviations. One such curve resulting from ten links in tandem is illustrated in Fig. 10, wherein a frequency distortion of approximately 3 decibels is obtained. Any teletype auxiliary-carrier equipment utilized over this channel would obviously have to take this total distortion or change in loss into account.

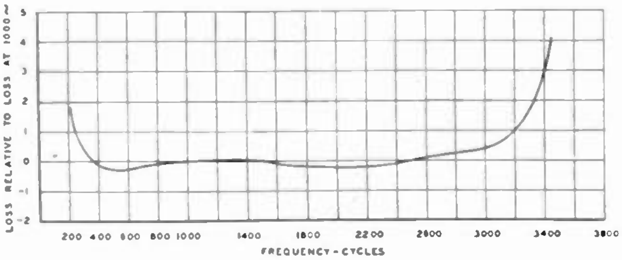

Fig. 9—Typical loss-frequency characteristic over one radio link or jump.

The audio-channel-frequency harmonic distortion present in one 50-mile link involving two equipments usually varies from 2 to 7 per cent, depending on which of the eight voice channels is tested. The production limit for such distortion has been set at 8 to 10 per cent, dependent upon the individual channel under test.

The output capabilities (audio-volume-control limits), based on a 0 decibel above 1 milliwatt, 1000-cycle sine-wave signal applied to any two-wire channel-input terminal, are such that the receiver audio (1000-cycle) outputs at the end of a 50-mile link are capable of being adjusted from +3 to −6 decibels.

The criterion for noise is based upon two equipments operating together over a 50-mile uninterrupted transmission path while seven of the audio channels are modulated with audio intelligence at a 0-volume-unit level. The noise measurements are made on the eighth, or unmodulated channel, with a Western Electric 2B Noise Measuring Set, utilizing "line weighting" (a frequency-weighting network which in effect compensates for the characteristics of telephone instruments and the human ear because they do not have the same efficiency at all frequencies). This test is made on a two-wire basis with a net circuit loss of 6 decibels. Under these conditions, which include cross talk, the total measured-noise-level average is about 28 decibels above a reference noise of −90 decibels as measured upon the 2B sets. Without cross talk, the noise levels vary from 16 to 25 decibels above reference noise level.

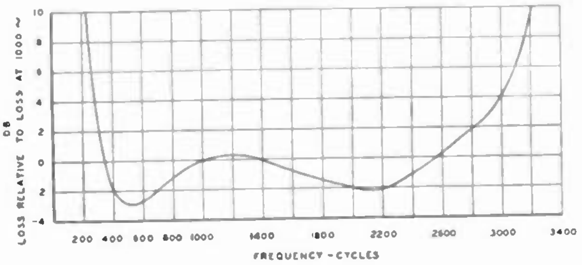

Fig. 10—Loss-frequency characteristic of a typical channel over ten radio links in tandem using voice-patch repeaters.

These noise levels indicate the extended range or distance of the communication circuit which can be set up if these equipments are tied in tandem to the limit whereat the system can still be used as a trunk line. With AN/TRC-6 equipments, wherein the video-repeater facility can be utilized at relay points rather than coming down to voice, harmonic distortion as a limitation will be negligible; and for voice intelligence, the limiting characteristic in all probability will be noise.

Tactical wire circuits are designed for a 35-decibels-above-reference maximum noise level which includes an allowance of approximately 5 decibels margin for below-average talkers.

For the purpose of this discussion, assume that a series of uninterrupted transmission paths of 25 miles each is available (25 miles is understood to be the average length of line-of-sight-paths in this country). For this distance, a noise level of approximately 20 decibels above reference noise will be experienced per jump, involving two AN/TRC-5 radio sets or two AN-TRC-6 radio sets; 35 decibels minus 20 decibels will therefore allow 15 decibels of noise to be added and still have a radio relay circuit which can be tied into a high-grade four-wire tactical wire circuit. If the noise levels in all the radio links tied in tandem to expend the 15 decibels are the same, i.e., 20 decibels above reference noise, the antilogarithm of 1.5, or 31 jumps are feasible. This will require 62 radio sets for a 775-mile circuit. If good intelligibility or no teletype errors due to noise is all that is required over such a circuit, a value of noise as high as 50 decibels above reference can be tolerated; 50 decibels minus 20 decibels, or at least 30 decibels, will then be available. This 30-decibel figure will then allow a theoretically possible figure of 1000 links or 25,000 miles. However, sufficient radio sets and properly trained personnel have never been accumulated in one locality to explore thoroughly the communication-distance limitation of these equipments.

Indications are that equipment of this type will be used to the utmost in commercial communication systems because of its relatively low cost and ease of maintenance as compared to coaxial cables or other wire lines. Of course, these factors will vary dependent upon weather conditions, the type of application, terrain, etc.; but in any event, the microwave experience gained by industry during the war while working in conjunction with the services on such equipments as the AN/TRC-5 and AN/TRC-6 has shown the outstanding feasibility of the microwave spectrum for specialized communications.

The main troubles encountered to date at these microwave frequencies are those which were expected, or rather, feared, at the initiation of the developments, namely their propagation characteristics. Premonition of such difficulties was the reason for seemingly to duplicate equipments, i.e., the AN/TRC-5 and AN/TRC-6, since no information was available on point-to-point propagation in the microwave range and military necessity demanded equipments in a hurry. Wave interference has been experienced between the direct beam and that reflected from the intervening terrain or water. It has been possible to practically eliminate this multipath reception effect by a means known as "diversity reception." This essentially consists of two receiving antennas (as shown in Fig. 11) so spaced vertically that a choice within the range between interference or reinforcement of the received beams is possible. The receiver is so connected as to utilize automatically the antenna receiving the strongest signal. Anomalous propagation, whereby a wave-guide duct effect is present in the atmosphere due to a temperature-gradient inversion from normal, may capture a microwave beam and play tricks with it. If it bends the beam around the earth, the "line-of-sight" transmission-path length is thus extended, but generally this effect is intermittent and may be equally harmful to the operable range. In any event, a complete study of the propagation characteristics at these frequencies is much closer to being accomplished now that the necessary fundamental microwave components are available. It is expected that much of this fundamental and invaluable data will emanate from future field tests of the AN/TRC-5 and AN/TRC-6 equipments or modifications thereof.

Fig. 11—Vertical diversity antenna for Radio Set AN/TRC-6.

IV. ACKNOWLEDGMENT

Particular credit for the basic design is due B. A. Trevor, project engineer for RCA Laboratories on Radio Set AN/TRC-5, H. S. Black, project engineer for Bell Telephone Laboratories (Western Electric Company) on Radio Set AN/TRC-6, and their very capable technical staffs.

Capacitance-Coupled Intermediate-Frequency Amplifiers*

MERWIN J. LARSEN†, SENIOR MEMBER, I.R.E., AND LYNN L. MERRILL†

Summary—**This paper discusses the performance characteristics of capacitance-coupled intermediate-frequency-amplifier circuits. Double-ended loading is preferred to single-ended loading in order to satisfy practical tolerance requirements.**

An approximate method is treated for the design of capacitance-coupled attenuating traps for a television intermediate-frequency amplifier. Performance specifications for such an amplifier are considered. Performance characteristics of an amplifier built to conform to the specifications as outlined are shown.

INTRODUCTION

THIS PAPER deals exclusively with the design and performance of the capacitance-coupled system, with particular reference to the application at frequencies above 20 megacycles. No special effort is made to compare the capacitance-coupled circuit with mutual-inductance-coupled circuits[1] or others which may function equally well. Even though a coupling capacitor is used, the capacitance-coupled circuit has constructional simplicity. The complete isolation of the coils made possible by capacitance coupling permits the use of slug tuning.

Although capacitance coupling has been used, it has not had as extensive application to wide-band amplifiers as inductive coupling. Consequently there is little literature specifically pertaining to it.[2-4] Fig. 1 shows the circuit diagram with its alternating-current equivalent for a typical television intermediate-frequency-amplifier stage. The capacitance-coupled trap attenuates either the adjacent or the desired-channel sound signal. A desired band-pass response may be obtained by loading either the grid or the plate (single-ended loading), or by loading both the grid and plate (double-ended loading). The results of an experiment made to compare the two methods of loading are discussed.

SINGLE-ENDED LOADING

For the same bandwidth, single-ended loading yields more gain than double-ended loading. With single-ended loading, however, the response in the pass region is badly disturbed by a small change in either coil inductance or tube capacitance. Experiment (see Appendix A), representative of a typical television intermediate-frequency-amplifier application indicates that a change of only 5 per cent in the inductance or capacitance associated with either coil produces an unbalance in the peaks of approximately 15 per cent, considering one stage only. A change in tubes alone might often cause more than a 5 per cent capacitance change.

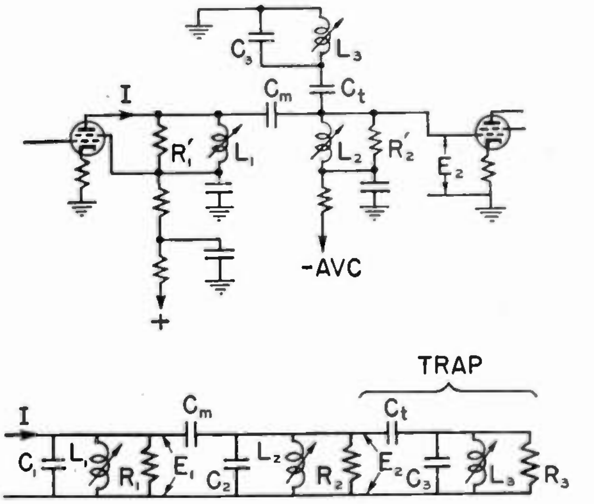

Fig. 1—Circuit diagram of a typical capacitance-coupled intermediate-frequency stage with its alternating-current equivalent.

It is concluded, therefore, that single-ended tuning is poor practice because component tolerance requirements are strained.

DOUBLE-ENDED LOADING

With double-ended loading, the shape of the response is only moderately vulnerable to alteration by minor component change. Damping resistors are selected which give a flat response, or a response with a slight center dip, when both coils are tuned to the same frequency. (In the case of capacitance coupling, the tuning frequency is at the extreme of the high-frequency end of the pass region. The coupling capacitor spreads the band toward the low-frequency end, leaving the other end relatively fixed.) When the damping resistors are selected in this manner, a moderate change in either inductance or capacitance shifts the band and amplitude somewhat, but does not unbalance the peaks appreciably. Quantitative data are tabulated in Appendix A. The experimental results given on loading are consistent with the trends indicated by Aiken.[5]

It appears evident that. for practical application where it is desirable to maintain a uniformity of response, double-ended loading is essential.

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† Stromberg-Carlson Company, Rochester, New York.

[1] G. Mountjoy, "Simplified television i-f systems," *RCA Rev.*, vol. 4, pp. 299–309; January, 1940.

[2] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Company, New York, N. Y., 1943; sec. 3, pp. 164–172.

[3] K. R. Sturley, "Radio Receiver Design, Part I," John Wiley and Sons, New York, N. Y., 1939; pp. 289–295.

[4] N. Marchand, M. Dishal, S. Frankel, "Wide Amplifier Design Nomograph," Federal Telephone and Radio Corporation, Newark, N. J., 1945. (Distributed at the 1945 I.R.E. Winter Technical Meeting.)

[5] C. B. Aiken, "Two-Mesh tuned coupled circuit filters," PROC. I.R.E., vol. 25, pp. 230–272; February, 1937.

TRAP DESIGN

In addition to the trap frequency, three other factors enter to determine the choice of trap-circuit parameters: first, the degree of attenuation desired (or obtainable); second, the effect of the trap on the response throughout the pass region; third, the shape of the response curve in the immediate vicinity of the trapping frequency.

Considering the capacitance-coupled circuit as shown in Fig. 2, the ratio of output voltage at trapping frequency to the output voltage in the pass band may be expressed approximately

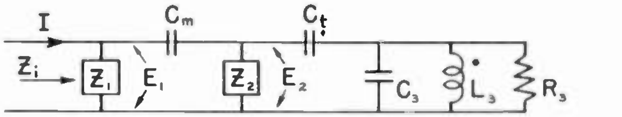

Fig. 2—Schematic diagram of capacitance-coupled circuit with trap.

$$\frac{E_{2t}}{E_{2p}} \doteq \frac{Z_{ip}C_m}{Z_{it}}\left[\left(\frac{C_3}{C_t}\right)\left(\frac{1}{Q_3C_t}\right)\right] \tag{1}$$

where

Z_{ip} = input impedance at a frequency within the pass band
Z_{it} = input impedance at the trap frequency
C_m = main coupling capacitance
C_t = trap coupling capacitance
C_3 = trap tuning capacitance
Q_3 = Q of trap coil.

While (1) gives the approximate attenuation for a given set of circuit parameters, it does not dictate automatically the optimum values to use. More information is needed. It has been found by experiment that, with the trapping frequency on the low side of the band-pass region, C_3 and L_3 should resonate at some frequency f_2 close to the trapping frequency f_t. The position indicated in Fig. 3 is representative for trapping the television sound signal of the desired channel. As the ratio C_3/C_t is related to the ratio f_2/f_t in the following manner

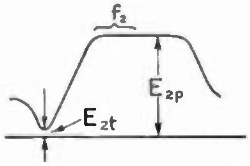

Fig. 3—Response showing approximate location of trap parallel-resonance frequency f_2.

$$C_3/C_t = 1/[(f_2/f_t)^2 - 1], \tag{2}$$

a choice of f_2 determines the ratio C_3/C_t.

Assuming that the ratio C_3/C_t is established from (2), it is clear from (1) that the attenuation increases with C_t. Increasing the trap coupling capacitance indefinitely places excessive capacitive loading across Z_2 and reduces the gain in the pass region. Thus C_t will be a compromise value allowing sufficient attenuation without seriously reducing the pass-band gain.

The greater the Q of the trap coil Q_3, the greater the attenuation and the sharper the dip. If the shape of the trap response is too narrow, a slight amount of oscillator drift throws the signal to one of the steep sides. On the steep slope the discriminator action on the frequency-modulated sound carrier causes spurious amplitude modulation in the picture.

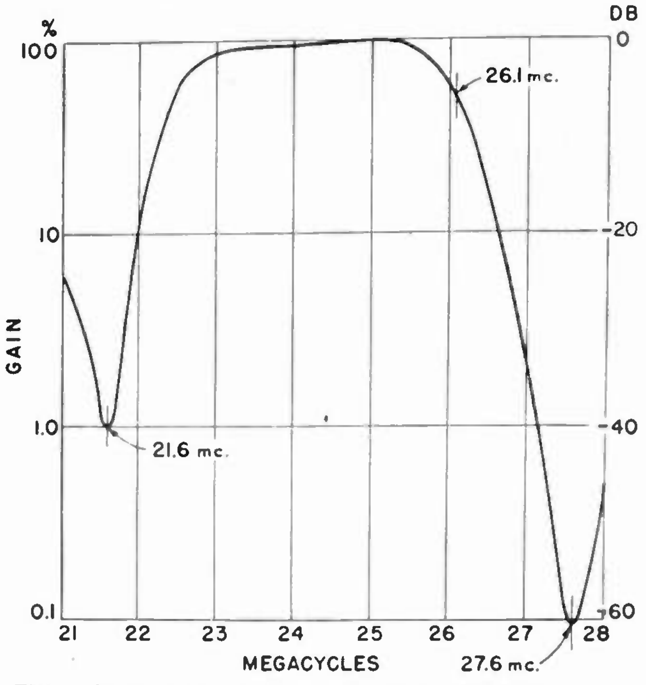

Fig. 4—Response curve of four-stage intermediate-frequency amplifier.

PERFORMANCE SPECIFICATIONS FOR TELEVISION PICTURE INTERMEDIATE-FREQUENCY AMPLIFIER

A frequency of 21.6 megacycles is used for the sound carrier of the amplifier to be discussed. This frequency was advocated by Newlon[6] in a report presented to the Radio Manufacturers Association committee on Television Receivers. The sound carrier selected lies midway in the 21.25- to 21.9-megacycle region formulated as a standards proposal by the executive committee of the receiver section of the RMA engineering department. With the sound carrier at 21.6 megacycles, the picture carrier is 26.1 megacycles.

The desired-channel sound should be attenuated at least 35 decibels, provided the bottom of the trap response is flat enough to limit the envelope caused by discriminator action to −50 decibels.[7] The bottom of the trap response should be broad enough to allow for oscillator drift.

The adjacent-channel sound signal will be on the threshold of discernibility at approximately −50 decibels from the level of the received television signal.

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[6] A. E. Newlon, "Selection of television sound and picture intermediate frequencies," Report to Radio Manufacturers Association Committee on Television Receivers, August, 1945.

[7] M. J. Larsen, "Minimum Acceptable Trappage Performance," Report to Subcommittee on Standards of Good Engineering Practice of Radio Manufacturers Association Committee on Television Receivers, October, 1945.

How much attenuation is needed, however, depends on ultimate station allocations as well as on how much protection the manufacturer wishes to give to those using receivers in unfavorable locations.

The response curve of Fig. 4 fulfills the specifications outlined above. The desired-channel sound trap attenuates 40 decibels at 21.6 megacycles. The adjacent-channel sound attenuates 60 decibels at 27.6 megacycles. The picture carrier is 6 decibels down at 26.1 megacycles to meet the vestigial-sideband-transmission requirements. The response shown permits a video response of over 3.5 megacycles before the high-frequency end tapers rapidly.

SELECTION OF CIRCUIT PARAMETERS

The picture intermediate-frequency amplifier constructed to produce the response of Fig. 4 has four stages of amplification. Including the mixer output, there are five sets of coupled circuits with traps of the type shown in Fig. 1. The only capacitances involved are the tube output and input capacitances, including associated socket, wiring, and coil strays. The inductors are copper-slug tuned.

The first three sets of coils have the adjacent-channel 27.6-megacycle traps. The desired sound carrier is taken off the plate of the third intermediate-frequency tube via a small (½-micromicrofarad) capacitor. The third and fourth stages have the desired-channel sound traps.

Automatic gain control is applied to the first three stages. Unby-passed cathode resistors are necessary in order to stabilize the tube input capacitances as the bias is varied. Although this reduces the gain, it permits over 60 decibels variation in gain without perceptible change in the shape of the response curve.

The loading resistors and coupling capacitors for the band-pass circuit can be found readily by experiment with the aid of a sweep generator, oscilloscope, etc. In the alignment procedure the coils L_1 and L_2 are tuned with their respective capacitances to the same frequency near the high end of the band, say 26 megacycles. The resistors and capacitor C_m are adjusted to give the desired response, allowing a margin for "cutting in" of the traps. The traps are designed with the aid of (1) and (2) and tuned to attenuate at the appropriate frequency.

Typical approximate values for the various circuit parameters are listed below. For the band-pass circuit:

C_1—5 micromicrofarad
C_2—11 micromicrofarad
L_1—7.5 microhenry
L_2—3.4 microhenry
R_1'—10,000 ohms, ½-watt direct-current nominal
R_2'—3700 ohms, ½-watt direct-current nominal
C_m—2.0 micromicrofarad.

For the desired-channel sound trap, $f_t = 21.6$ megacycles:

C_3—23 micromicrofarad
C_t—3 micromicrofarad
L_3—2.1 microhenry
Q_3—70.

For the adjacent-channel sound trap, $f_t = 27.6$ megacycles:

C_3—12 micromicrofarad
C_t—2 micromicrofarad
L_3—2.3 microhenry.

The gain averaged over the four stages, using type 6AG5 tubes, measured 8 per stage.

APPENDIX A

EXPERIMENTAL MEASUREMENT OF COUPLED-CIRCUIT PERFORMANCE

For the coupled circuit of the type shown in the schematic of Fig. 1, but minus the trap, the equation of the output voltage E_2 may be expressed in the form

$$E_2 = \frac{I}{2\pi f_0\sqrt{C_1C_2}}\left[\frac{f/f_0}{\sqrt{\frac{C_2}{C_1}}\left[\frac{1}{Q_2} - j\alpha\right] + \sqrt{\frac{C_1}{C_2}}\left[\frac{1}{Q_1} - j\alpha\right] - j\left(\frac{f_0}{f}\right)^2\frac{\sqrt{C_1C_2}}{C_n}\left[\frac{1}{Q_1} - j\alpha\right]\left[\frac{1}{Q_2} - j\alpha\right]}\right]. \tag{3}$$

In (3) both coils are tuned to the same frequency f_0. Constant input current I is assumed. $Q_1 = R_1/wL_1$, $Q_2 = R_2/wL_2$, $\alpha = 1 - (f/f_0)^2$.

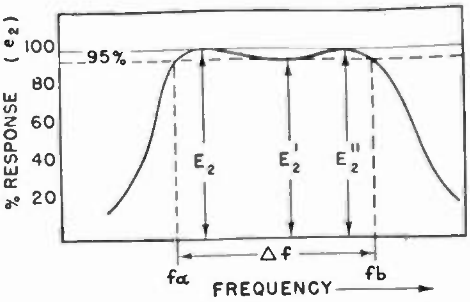

Fig. 5—Response curve used as reference in determining response deviations with changes in circuit inductances.

It is noted that the portion of (3) in the brackets is a function of frequency ratio, capacitance ratio, and Q. These three are simultaneously related to produce a given response shape. The relative shape will be the same, therefore, at any frequency, provided the frequency ratio, capacitance ratio, and Q's remain the same. The amplitude may vary, however, depending on the value $I/2\pi f_0\sqrt{C_1C_2}$.

It is convenient, then, to write (3) as

$$E_2 = \frac{I}{2\pi f_0\sqrt{C_1C_2}} B_{a,b} \tag{4}$$

where $B_{a,b}$ indicates that the values in the brackets of (3) are selected to give a curve of some reference characteristic (see Fig. 5) having a frequency ratio of f_a/f_b, where f_a and f_b are the band-pass limits, and having specified ratios of E_2, E_2', and E_2''.

The above analysis shows the manner in which it is valid to extend from one frequency to another. Regard must, of course, be taken for any change with frequency of any of the circuit components.

An audio-frequency coupled circuit was used to measure the effect of component changes on the shape of the response curve. Audio frequencies were used to facilitate taking quantitative data. The circuit employed was made approximately equivalent to the television intermediate-frequency-amplifier circuit, without trap, as discussed herein, so that the results would have practical significance. In the tests only the inductance was varied, it being apparent that similar changes in capacitance would have comparable results.

Table I shows the case with double-ended loading. The loading was adjusted to yield symmetrical response, after both coils were tuned to the same frequency f_0. The inductances were varied 5 per cent from their reference values. The symbols used in the table are indicated on the reference curve shown in Fig. 5.

TABLE I

Double-Ended Loading, Showing Effect on Response Shape of 5 Per Cent Changes in Inductance

Condition	Per cent of f_0		Per cent $\Delta f/f_0$	Per cent of L_2		Per cent of E_2		
	f_a	f_b		L_1	L_2	E_2	E_2'	E_2''
Response balanced $L_1C_1=L_2C_2$	86	100	14	100	50	100	95	100
L_1 decreased 5 per cent	87	100	13	95	50	105	102	106
L_1 increased 5 per cent	84	99	15	105	50	95	90	95
L_2 decreased 5 per cent	86	101	15	100	47.5	94	88	94
L_2 increased 5 per cent	85	99	14	100	52.5	105	101	106

Under balanced-response conditions of operation as shown in the first row: $Q_1=8.8$, $Q_2=7.2$, $f_0=5000$ cycles, $L_2=0.615h$, $C_2/C_1=2$ (L_2 and C_1 resonate at f_0), $B_{a,b}=3.3$. (See (4) for computing gain at any other frequency, using $\Delta f/f_0=14$ per cent, 5 per cent dip, $C_2/C_1=2$).

Table II shows the results with the loading pre dominantly on the secondary side Values of L_1, L_2, and the loading resistor on the secondary were adjusted to yield the same reference curve as under the double-loaded, balanced condition of Table I. It was necessary to increase the coupling capacitance 20 per cent over that used for the results of Table I. Experiments show that, where both coils are tuned to the same frequency, one set of Q's only will give symmetrical response for a specified per cent center dip. If the loading is predominantly on one side, then the coils must be detuned relative to each other in order to produce the symmetrical response. Measurements for this particular case showed a gain increase of 30 per cent over the case of double-ended loading.

TABLE II

Single-Ended Loading, Showing Effect on Response Shape of 5 Per Cent Changes in Inductance

Condition	Per cent of f_0		Per cent $\Delta f/f_0$	Per cent of L_2		Per cent of E_2		
	f_a	f_b		L_1	L_2	E_2	E_2'	E_2''
Balanced response but $L_1C_1=0.89L_2C_2$	86	100	14	92	51.5	100	95	100
L_1 decreased 5 per cent	90	103	13	87	51.5	94	93	109
L_1 increased 5 per cent	83.5	97	13.5	97	51.5	109	93	94
L_2 decreased 5 per cent	86	100	14	92	49	105	90	91
L_2 increased 5 per cent	86	100	14	92	54	94	93	109

Under balanced response conditions: $Q_1=27$, $Q_2=5.1$, $f_0=5000$ cycles, $L_2=0.615h$, $C_2/C_1=2$ (L_2 and C_1 resonate at f_0), $B_{a,b}=4.3$. (See equation (4) for computing gain at other frequencies, using $\Delta f/f_0=14$ per cent, 5 per cent dip, $C_2/C_1=2$).

APPENDIX B

DERIVATION OF TRAP DESIGN FORMULA

The following is an approximate method for designing capacitance-coupled traps yielding results close enough for practical use at the higher frequencies.

Considering the capacitance-coupled circuit shown in Fig. 2, the approximate expression for the output voltage across Z_2, neglecting voltage drop across C_m, is

$$E_{2p} \doteq IZ_{ip} \tag{5}$$

where subscript p denotes calculation at a frequency in the pass region. Also

$$E_{1t} = IZ_{it} \tag{6}$$

where subscript t denotes calculation at the trapping frequency.

The circuit of Fig. 6 is an approximation of the circuit of Fig. 2 operating at the trapping frequency at which frequency Z_2 is effectively shunted out by the lower trap impedance. Resistance R_3 (see Fig. 2) is chargeable to the losses in coil L_3. As Q_3 of coil L_3 is moderately high, say over 50, R_{3t}', representing the series resistance of the trap when the capacitive reactance of C_t is equal in magnitude to the inductive reactance of the combination C_3 and L_3, may be written

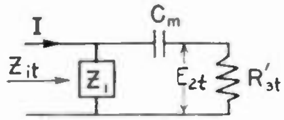

Fig. 6—Approximate representation of capacitance-coupled circuit with trap when considered at the trapping frequency.

$$R_{3t}' = X_{C_t}{}^2/Q_3X_3. \tag{7}$$

Also, at the trap frequency, from Fig. 6,

$$E_{2t} = E_{1t}R_{3t}'/X_{C_m}. \tag{8}$$

From the above four equations, the attenuation ratio follows:

$$E_{2t}/E_{2p} \doteq (Z_{ip}/Z_{it})(C_mC_3/Q_3C_t^2). \tag{9}$$

Fixing empirically the frequency f_2 at which L_3 and C_3 are in resonance, and knowing that f_t is the frequency of series resonance between C_m and L_3, C_3, it follows that

$$C_3/C_t = 1/[(f_2/f_t)^2 - 1]. \tag{10}$$

In equation (9) the ratio (Z_{ip}/Z_{it}) and C_m remain reasonably constant over a considerable range of trap parameters. The trapping is adjusted then by the quantity $C_3/Q_3C_t^2$ after giving consideration to selecting the value for f_2, as discussed in the body of the paper.

As an example of the application of (9) and (10), using the circuit components listed for the television intermediate frequency, the attenuation of the 21.6-megacycle desired-channel sound trap is approximately 18 decibels. The two traps in the complete amplifier, with the help of some "sloping off" of the response by the remaining three sets of coupled circuits, give a total attenuation of 40 decibels by test measurement.

Factors Affecting the Accuracy of Radio Noise Meters*

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Summary—**The factors which influence the accuracy of radio noise measurements can be divided into two classes: namely, those which affect the accuracy of steady-state sine-wave radio-frequency measurements, and those which are peculiar to the nonsinusoidal wave shapes of radio noise. Some of the factors can be controlled by suitable standardization of design specifications and methods of measurement, while other factors are dependent on the characteristics of the wave form and on the point at which it is to be measured. Probably the greatest measurement discrepancies that would be encountered under well-controlled conditions can be attributed to differences in the wave collectors employed and to the variations in charge time "constant" of the metering circuit with differences in noise levels. An extensive analysis of the latter effect is presented.**

INTRODUCTION

THE PURPOSE of this paper is to discuss various factors that contribute to the difficulty of accurately measuring radio-frequency noise voltages and field strengths. Continuous-wave field intensities can ordinarily be measured, with modern equipment, to accuracies within 5 to 10 per cent up to about 5 megacycles, and within 20 per cent from 5 to 400 megacycles, if experienced personnel use normal care in making the measurements. The measurement of radio-frequency noise fields, however, is considered to be as good as can be expected if accuracies within 30 to 50 per cent are obtained with similar noise meters. When different types of meters are used, the results may disagree as much as 1000 to 1.

Part of the difficulty is due to the erratic nature of most noise voltages, but much of the inaccuracy encountered can be attributed to limitations in the noise meter proper and to the different types of wave collectors used with the meter. The latter factor is especially important when working in induction fields wherein the field along the antenna is not uniform.

With sine wave form, such as an unmodulated radio carrier, the effective and average values are simple functions of the peak value. With the irregular wave forms of radio noise, the ratios of the effective and average values to the peak value will vary over extremely wide ranges. Inasmuch as this large variation exists, it becomes necessary to specify whether the peak, effective, average, or some intermediate value is to be used. It has been generally agreed[1] that the interfering effect of frequently occurring impulses, for most applications, is more nearly proportional to the peak value than to the effective or average values. For this reason it is common practice to design radio noise meters to read a value that is as near the peak amplitude as possible. Since this usually falls short of the true peak, it is sometimes referred to as a quasi-peak value. It should be pointed out here that a single unit for designating the degree of interference is probably inadequate and that one or more additional quantities capable of designating the characteristic type of the noise may eventually be used. At present there is no suitable method of accomplishing this, at least in field equipment.

Inaccuracies can be attributed to several or all of the following factors: (a) antenna differences; (b) differences in bandwidth and image response; (c) differences in automatic-volume-control and weighting-circuit constants; (d) differences in indicating-instrument characteristics; (e) overloading and other non-linear effects; (f) inadequate shielding; (g) detuning of amplifier circuits by Miller effect; (h) mistracking of radio-frequency circuits; (i) power-supply voltage variation; (j) variation in tube characteristics; (k) inaccurate attenuator steps; (l) inaccurate calibration technique, and (m) careless or improper operation.

In an attempt to provide uniformity in results, certain specifications have been recommended by the Joint Co-ordination Committee on Radio Reception of the Edison Electric Institute, the National Electric Manufacturer's Association, and the Radio Manufacturer's Association.[2] Extensive measurements made with different meters built in accordance with these specifications, however, indicate that additional requirements must be met before good correlation is to be obtained.

REQUIREMENTS

Certain requirements are imposed on radio noise-measuring instruments that are not necessary and frequently not desirable in ordinary field-intensity meters. Since the amount of interference passing through the receiver will depend, in some manner, on the selectivity of the receiver circuits, the bandwidth must be known, and if the results are to be correlated with other instruments, it should be fixed at a suitable standardized value. The image response ratio must be large so that noise outside of the regular pass band does not contribute to the measured value. The automatic-volume-control and weighting circuits must have charge and discharge time constants that fall within specified limits. The indicating instrument must have a damping factor

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† Naval Research Laboratory, Washington, D. C.

[1] Collected International Radio Consulting Committee (C.C.I.R.) documents on questions 2, 5, 14, 15, 16, 17, and particularly, documents 5, 9, 13, 14, 17, 18, and 52.

[2] Report of Joint Co-ordination Committee on Radio Reception, "Method of measuring radio noise," Publication No. C 9, 1935 and No. G 9, 1940; Edison Electric Institute. Publication No. 102, 1935, and No. 107, 1940; National Electric Manufacturers Association Engineering Bulletin No. 13, 1935, and No. 32, 1940; Radio Manufacturers Association.

and time constant that fall within specified limits. The antenna system should be of a type that allows flexibility in its use about the installations or equipments to be measured and should be standardized as to type, form, and size, since the field to be measured ordinarily will not be of the plane-wave variety. The necessity of keeping the size and weight of the complete noise meter low enough to meet portable requirements has frequently been responsible for a large sacrifice in accuracy.

ANTENNA CONSIDERATIONS

In general, there are three fundamental types of antennas in use with present-day radio noise meters: (a) short single-rod antennas; (b) half-wave dipoles; and (c) loops.

For plane-wave fields, all of the above types have been successfully adapted. When measuring fields near the source, however, different results are obtained with different types. In fields having a high gradient the term "effective height" becomes meaningless; the size and shape factors must be closely alike if similar results are to be expected; and near the source either the magnetic or the electric field may predominate, and thus the results obtained using a loop will differ from those using a rod antenna. Inasmuch as either the magnetic or the electric field can be a source of interference, it is desirable to be able to measure both.

The single-rod type has high-impedance output, fixed length, and nearly constant effective height for *plane* waves. The half-wave type has low-impedance output. Its length must be adjusted for frequency, and therefore its effective height varies with frequency. These linear types give greatest response for high-impedance noise sources, that is, for fields predominantly electric. The loop type is fixed in size but its effective height varies with frequency. Its impedance varies with design and frequency. It gives greatest response on low-impedance noise sources, that is, on fields predominantly magnetic. For unbalanced loops, and also for single-rod antennas, still another complication is added by stray capacitances between the case of the meter and the body of the operator or ground.

BANDWIDTH CONSIDERATIONS

The effect of receiver bandwidth on various types of noise has been investigated by Landon[3] and Jansky.[4] It is common practice to divide noise into two categories which have been defined as follows: (1) *random noise* is noise due to the aggregate of a large number of elementary disturbances with random relative phases and with its energy not confined to a narrow band in the frequency spectrum. Thermal noise, shot noise, and corona discharge belong to this group; (2) *impulse noise* is noise due to a single elementary disturbance or to an aggregate of elementary disturbances with systematic relative phases and with its energy not confined to a narrow band in the frequency spectrum. Ignition noise and radar interference are excellent examples of this type. It is doubtful whether these two simple classifications are sufficiently specific to permit adequate accuracy in measurement, since various combinations of the two are frequently encountered in practice. It may be desirable to make use of an additional quantity which will, in some manner, indicate the statistical distribution of the noise. For random noise the peak, effective, and average values are proportional to the square root of the bandwidth. For impulse noise the peak value is proportional to the bandwidth, and the average value is independent of bandwidth.

RESPONSE TIME

It has been standard practice to use a resistance-capacitance network in the second-detector output circuit of a sensitive superheterodyne receiver in order to obtain a metering voltage that rises rapidly to a value corresponding to the peak radio-frequency input voltage and remains near that value between successive peaks. Three different sets of values for the charge and discharge time constants of this network have been standardized by different authorities. The International Special Committee on Radio Telephone Disturbances (CISPR) recommends 1 millisecond charge and 160 milliseconds discharge.[1] The Joint Coordination Committee (U.S.A.) specifies 10 milliseconds charge and 600 milliseconds discharge.[2] The British Electrical and Allied Industries Research Association (ERA) requirement is now 1 millisecond charge and 500 discharge.[5]

Since automatic volume control is normally used to obtain logarithmic response on the indicating instrument, the time constants of the automatic-volume-control circuit must be taken into consideration along with those of the weighting circuit proper. In some cases the weighting circuit is entirely separate from the automatic-volume-control circuit, but more frequently certain circuit elements are common to both. In either case, the over-all time constant will determine the over-all response characteristic.

The charge time of a radio noise meter, exclusive of the indicating instrument, may be defined as the time required for the voltage applied to the indicating instrument to reach 63 per cent of its final value upon the application of a *specified value* of sine-wave radio-frequency voltage to the input terminals of the noise meter.

In order to measure accurately the over-all charge time of several equipments, several electronic-photographic arrangements were devised providing both instantaneous viewing and permanent records. Good correlation between the methods was obtained.

Numerous measurements made on several commercial noise meters indicated that the over-all charge time differed considerably from calculated values based on

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[3] V. D. Landon, "A study of the characteristics of noise," PROC. I.R.E., vol. 24, pp. 1514–1521; November, 1936.

[4] K. G. Jansky, "An experimental investigation of the characteristics of certain types of noise," PROC. I.R.E., vol. 27, pp. 763–768: December, 1939.

[5] F. M. Colebrook, "An interim report on preliminary work by the British Institution committee on the measurement of radio interference," Radio Division, National Physical Laboratory, Report, September 4, 1944.

the resistance-capacitance circuit constants. The charge time was found to depend on the design of the weighting circuit, the design of the automatic-volume-control circuit, *and the amplitude of the applied radio-frequency voltage.* With impulse type of interference, it also depends, in some manner, on the pulse length and repetition rate, although it is not a direct function of the duty cycle. Since it is a variable, rather than a constant, the term *charge time* is used here in place of the more common term, charge time constant.

The measured values of charge time for a typical commercial instrument were plotted against input voltage on log-log paper (see Fig. 1). The points fell on a straight line which when projected to 1 microvolt (the low end of the meter scale) passed through a value equal to the computed charge time constant of the weighting circuit alone (based on an effective diode resistance of 10,000 ohms). This variation in charge time with signal input level is a result of the wide variation in receiver gain as the automatic-volume-control voltage builds up to a steady-state condition.

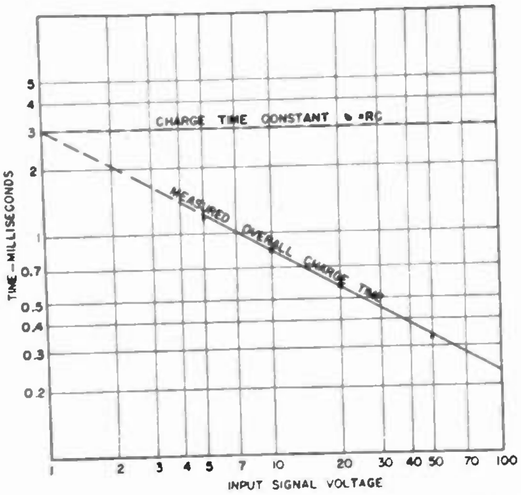

Fig. 1—Variation in over-all charge time with signal level.

The manner in which the automatic volume control influences the charge time can be demonstrated by means of a graphical solution of the over-all charge-time relationship. Fig. 2 shows a typical equivalent weighting circuit with the associated voltages. The effect of the variation in diode resistance is neglected and a constant equivalent value is included in the value of R. This effect is believed to be small in comparison with the effects under consideration, and omitting it greatly simplifies the analysis.

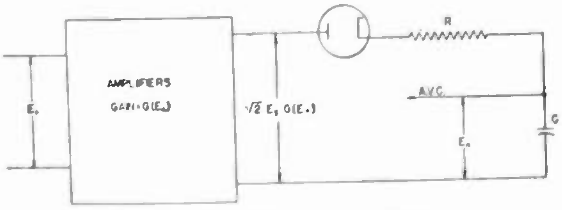

Fig. 2—Simplified charge circuit.

The differential equation for the voltage appearing across R and C is

$$RC\,dE_a/dt + E_a = \sqrt{2}\,E_sG(E_a)$$

or

$$\frac{t}{RC} = \int \frac{dE_a}{\sqrt{2}\,E_sG(E_a) - E_a} \tag{1}$$

where E_a is the voltage across C, E_s is the root-mean-square value of the input signal, and $G(E_a)$ is an exponential gain function of E_a.

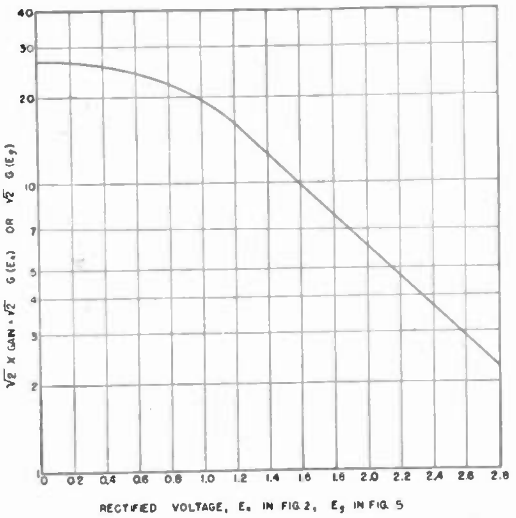

Fig. 3—Gain as a function of grid voltage.

Since the exact form of $G(E_a)$ is not known, the integration cannot be performed mathematically. However, values of $\sqrt{2}G(E_a)$ for a typical instrument were plotted against E_a (see Fig. 3). Equation (1) describes a family of curves, one for each assumed value of E_s. An explicit graphical integration of (1) was performed to obtain the relationship between t/RC and E_a/E_{ao}, for four values of E_s: 100, 20, 10, and 5 microvolts. The four curves corresponding to these values of E_s are plotted as A, B, C, and D in Fig. 4. The values of E_a determined for each curve were divided by E_{ao}, the steady-state value of E_a, and the ratio E_a/E_{ao} was used as the ordinate in plotting the curves. This was done to eliminate

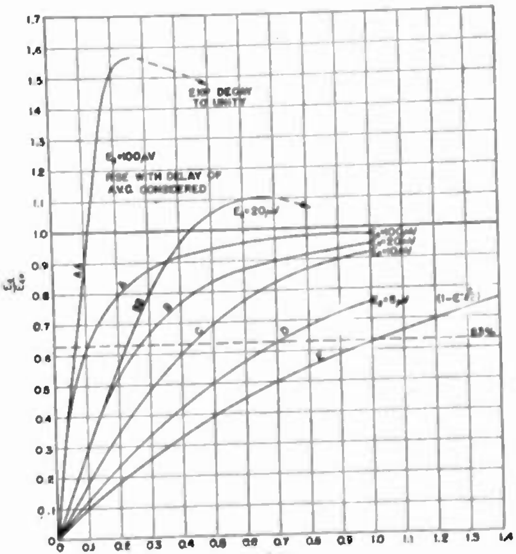

Fig. 4—Charge-time variation.

the relative magnitudes of the curves as a factor in studying the shapes of the curves. The curves A, B, C, and D progressively approach curve E, as E_s is made smaller. Curve E is plotted from the relation

$$E_a/E_{a0} = 1 - \exp.\,(-\,t/RC). \tag{2}$$

The gain curve, as shown in Fig. 3, shows a definite tendency to flatten at low signal levels, so that the direct-current voltage $\sqrt{2}E_sG(E_a)$ applied to R and C in Fig. 2 is approaching the condition of a constant applied voltage across the network. Equation (2) will be recognized as the solution of (1) for the condition $\sqrt{2}E_sG(E_a)$, equal to a constant. The relationship of curves A, B, C, and D to curve E in Fig. 4 shows that the over-all charge time becomes longer as the amplitude of the input signal is decreased and that it approaches the charge time constant of the resistance-capacitance network for low levels of input. This is the same as the relationship expressed by the experimental data in Fig. 1.

In the above graphical solution, several factors, in addition to the diode resistance, have been neglected. The Q of the amplifier circuits introduces a delay in the build-up of voltage at the diode. From published information concerning transient response of tuned circuits,[6] it is believed that this effect is small compared with a third factor which has also been neglected; this is the effect introduced by the delay in applying the rectified automatic-volume-control voltage E_a to the grids of the tubes. To study this effect the automatic-volume-control time-constant circuit must be added to the equivalent circuit, giving Fig. 5. The analysis is now more difficult than in the previous case. The product of R_1 and C_1 for the instrument under consideration was 0.0006. Using this value, an approximate solution for the relationship of voltage rise against time was obtained by the step-by-step method of calculation.[7] The result is shown by curves AA and BB of Fig. 4. The rise to maximum value was plotted and the decay to unity indicated. The over-shoot shown is characteristic of the actual results obtained by the photographic methods described above and has been present on all measurements, but varying in magnitude, and being dependent on the delay involved in applying the rectified voltage to the grids of the tubes controlled by automatic volume control. The value of t/RC for a rise to 63 per cent on curve AA is 0.067. The time constant of the weighting circuit was approximately three milliseconds. Based on curve AA, for E_s equal to 100 microvolts, the over-all charge time should be

$$t = 0.067 \times 3 = 0.201 \text{ milliseconds.}$$

The value actually measured was 225 microseconds (see Fig. 1), which is in fair agreement.

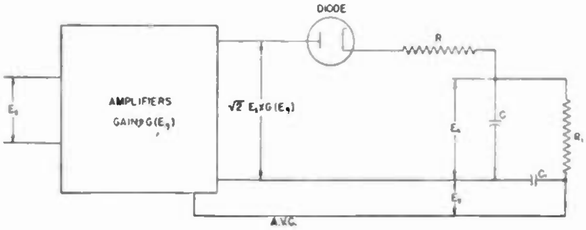

Fig. 5—Simplified charge circuit with R and C added.

Although several factors have been entirely neglected in this analysis, the results obtained by the graphical method agree so well with measurements that the factors considered, namely, the large variation in gain during the transient period and the delay in applying the automatic-volume-control voltage to the grids, appear to be of principal importance.

A pulsed radio-frequency signal was fed to the noise meter under test from a standard signal generator of known pulse wave form, for the purpose of observing the rectified voltage appearing across the diode load resistor on an oscilloscope and to determine the response of the equipment to various types of pulse signals. The pulsed wave form of the standard signal was first checked on a servoscope. The amplitude of the output was substantially constant whether pulsed or continuous. The pulses were essentially rectangular in shape. The rise time was approximately 0.5 microseconds and the decay time was less than 1 microsecond. Both were independent of either pulse length or repetition rate. The rectified pulse at the diode is illustrated in Fig. 6. The rise and decay times indicated are those of the pulsed radio-frequency signal at the input terminals of the noise meter and neglect any further delay that might result from the Q of the tuned circuits. To observe the variation in voltage appearing across the diode load resistor and associated network, the vertical plates of an oscilloscope were directly connected to the diode load resistor and the horizontal sweep was synchronized with the pulse-repetition rate. The shape of the pattern observed on the oscilloscope was similar to that formed by the lines marked E_a in Fig. 6. The variation in E_a for a 2-microsecond pulse at a repetition rate of 50 per second was less than 5 per cent of the average oscilloscope deflection, corresponding to E_a (average) in Fig. 6. Observation of the voltage E_a on the oscilloscope showed that the discharge time constant was adequate to maintain the voltage between successive pulses.

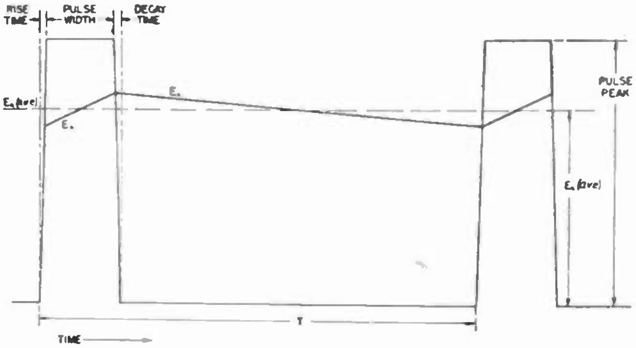

Fig. 6—Rectified pulse.

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[6] H. E. Kallman, R. E. Spencer, and C. P. Singer, "Transient response," Proc. I.R.E., vol. 33, pp. 169–195; March, 1945.

[7] Electrical engineering staff, M.I.T., "Electric Circuits," pp. 691–693, and pp. 705–711; John Wiley and Sons, Inc., New York, N. Y.

Fig. 7 shows the response of the noise meter to continuous waves and to various combinations of pulse length and repetition rate. It will be noted that successively greater input is required for the same meter reading as the pulse length and repetition rate are made smaller. The data in Fig. 7 are shown replotted in a different form in Fig. 8 to indicate better the variation in response with signal level. The ordinate of Fig. 8 corresponds to the value E_a (average) in Fig. 6, expressed as a percentage of the pulse peak voltage. The curve A, Fig. 8, for a continuous wave should be a horizontal straight line at 100 per cent. Its departure from a straight line expresses the meter scale-spread error of the instrument. The slopes of the other curves indicate the variation in response with signal level. The response of the curves $B, C, D,$ and E is much less at low signal levels than at high levels. The effect of the variation in over-all charge time given in Fig. 1 can be seen in these curves. Curves F, G, and H of Fig. 8 show the effect of amplifier and detector overload. Figs. 7 and 8 partially explain many of the anomalous readings observed with noise meters. The apparent erratic action of attenuators is accounted for. For example, on curve D of Fig. 7, it can be seen that if the pulsed radio-frequency input voltage is held constant at a value giving a meter indication of 50 on the "×1" step of the attenuator, and then the "×10" step of the attenuator is inserted, the reading will be 4.5×10 instead of 5×10. Or, from curve H, with other conditions the same, the reading would be 10.5×10 instead of 5×10. When this happens in practice, the observer is inclined to attribute the discrepancies directly to the attenuator rather than to the integrating action of the meter which is a character function of the noise. Measurements of pulses having small duty cycles will produce readings much below the actual peak irrespective of which step of the attenuator is used. The curves also help to explain why meters having different indicating scales may give different readings on the same noise input.

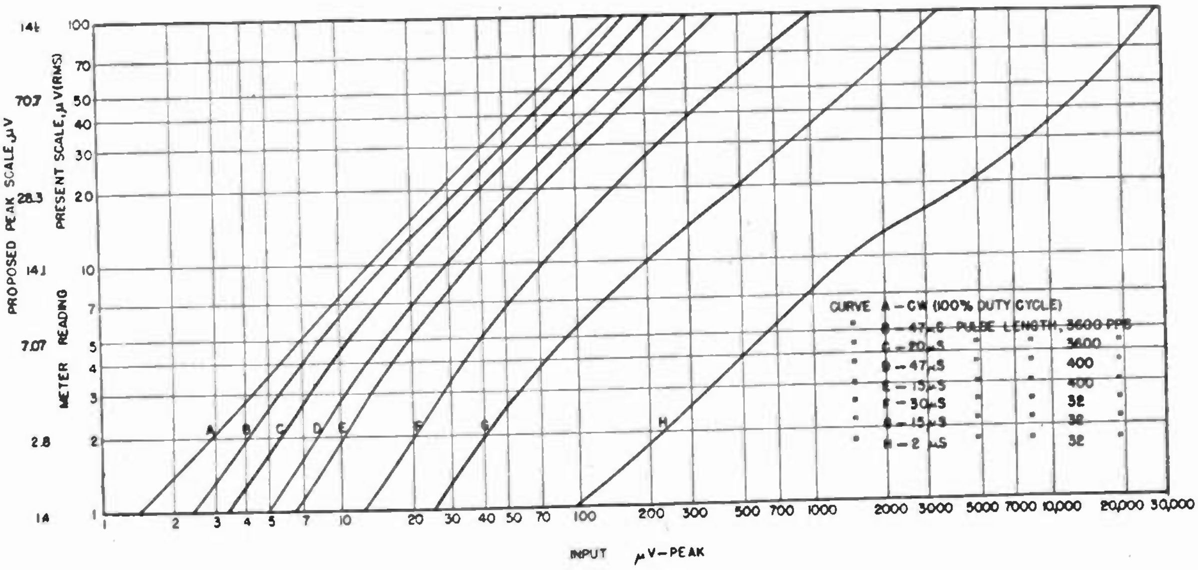

Fig. 7—Response of noise meter to pulse signals.

Indicating-Instrument Constants

The indicating instrument should be critically damped and its time constant should be short compared with the discharge time constant of the weighting circuit. It is usually required to have a damping factor of not less than 10 nor more than 100^2 as defined in the A.S.A. Standards for Electrical Indicating Instruments, C 39.1, 1938. The instrument is required to have a time constant of not less than 200 milliseconds nor greater than 500 milliseconds.

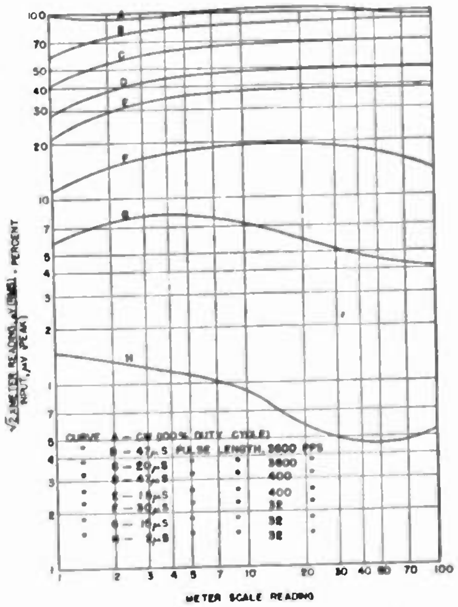

Fig. 8—Variation in percentage response with signal level.

Most radio noise meters have at least two scales on the indicating instrument, one of which is linear and calibrated in decibels above one microvolt, the other logarithmic and calibrated in microvolts. Either two or three decades are used for the latter scale. The logarithmic scale fulfills the psychological requirements of Weber's law and requires the use of fewer attenuator steps. The logarithmic action is usually obtained from the automatic-volume-control characteristics of the receiver, but in view of the foregoing discussion it may be desirable to obtain it by some other means having less effect on the response time.

In order to obtain uniform results between meters of different makes which use automatic volume control to obtain logarithmic response, it appears desirable to specify instrument scales that are very similar as to the number of decades used and the angular deflection per unit of signal.

MILLER EFFECT

The high degree of amplifier stability needed for the optimum accuracy in radio noise meters requires the practical elimination of Miller effect on the amplifier stages. The detuning effect on a preceding tuned circuit, caused by changes occurring in the input impedance of a tube, is well known.[8] The low grid-to-plate capacitance of multigrid tubes makes it possible to build moderate-gain amplifier circuits in which the Miller effect is so small that it is of little concern for many applications. However, the high gain obtained in some circuits using screen-grid or pentode tubes increases this effect to a point where it is quite troublesome, and corrective measures are necessary to obtain an amplifier with stable resonant frequency. This is particularly true of many noise meters employing automatic volume control. The presence of Miller effect in the amplifier stages of a noise meter produces the following adverse results: (a) the necessity of retuning for each input signal level. (b) inaccuracy due to variation in over-all bandwidth with signal level; (c) inaccuracy due to a variation in the output-meter scale tracking versus input voltage over the frequency range; and (d) variation in gain due to changes in the tracking of circuits.

Terman[8] states that the variation in the input capacitance of a typical radio-frequency pentode may detune an ordinary 450-kilocycle coupling circuit as much as 5 kilocycles. Changes of this order have been observed measurements on a commercial radio noise meter.

There are several ways in which neutralization of the Miller effect can be accomplished. In general this is done by the use of some form of feedback. A method of using an unby-passed portion of the cathode-bias resistor[9] as well as other feedback methods described in the literature[8,10] have been tried and found to be effective, not only for noise meters but also in conjunction with automatic recording of ordinary field intensities.

OTHER FACTORS

Mistracking of the Radio-Frequency and Mixer Circuits

The logarithmic scale on the indicating instrument is drawn for an average automatic-volume-control characteristic of the receiver. Improper tuning adjustments may change this characteristic and give poor scale tracking on the instrument scale.

Power-Supply Voltage Variations

The gain characteristic of the receiver is markedly influenced by changes in steady-state values of plate, screen, and heater voltages. Variations in these voltages are minimized by the use of ballast and voltage regulator tubes, but the tolerance found in commercial equipments is frequently large enough to cause serious errors.

Variations in Tube Characteristics

The automatic-volume-control characteristic is also, of course, a function of the individual transconductances of the amplifier tubes. Once the scale-spread adjustments are set for a particular complement of tubes, any changes in tube characteristics may necessitate readjustment of the scale-spread controls and recalibration of the noise meter. Tubes are selected for two different criteria; namely, sufficient gain to allow calibration, and proper shape of the transconductance curve.

Attenuator Inaccuracies

Although an apparent attenuator inaccuracy was explained above as being caused by other factors, actual attenuator errors are often present. A well-designed attenuator is essential.

Calibration Inaccuracies

The absolute accuracy of any measuring device can be no better than the accuracy with which it has been calibrated. The basic calibration of most radio noise meters is made as a two-terminal voltmeter connected to a standard signal generator through a suitable network. The calibration must be effected in such a manner that the input voltage-impedance relationships will be maintained. This requires an input network with characteristics governed by the generator and noise meter used.

When calibrating the "Field Intensity" position, the effective value of the signal-generator voltage is used. For the "Peak" (sometimes called the "Noise" or "Quasi-peak") position, the peak value of the signal generator voltage should be used. This latter point has often been over-looked and thus most equipments when calibrated according to their instruction manuals indicate a pseudo-effective value of the peak voltage.

The ideal method of calibration for radio noise values would be to calibrate both gain and meter scale spread with a suitable standard radio-noise generator. While there are several developmental projects for such a generator currently underway in various laboratories, there is as yet no suitable commercial model available. A method of using a square wave for the calibration that also holds some promise, at least for the lower frequencies, has been analyzed by Frick.[11]

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CONCLUSIONS

The measurement and evaluation of radio noise presents many unique problems. Several design compromises have been necessary, especially in field equipments, in order that the equipment does not become too complicated for ordinary use. Some of the more serious errors can be avoided by proper design, construction, calibration, and operation; others are still the subject of laboratory investigations. Further standardization will tend toward more uniformity in results obtained with different types of equipments. Increasing the absolute accuracy, however, is a much more difficult problem. Information is being accumulated from numerous experimental measurements which, it is hoped, will lead to more concrete design data.

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A Wide-Tuning-Range Microwave Oscillator Tube*

JOHN W. CLARK†, MEMBER, I.R.E., AND ARTHUR L. SAMUEL‡, FELLOW, I.R.E.

Summary—**This paper describes a reflex-type velocity-variation oscillator tube with a wide tuning range in the microwave band. The tube will oscillate from 2000 to 13,000 megacycles, but practical tuning considerations limit the band in any one circuit to a two-to-one frequency range. The problems involved in the design and a description of the various elements are given.**

AN IMPORTANT item of equipment needed for effective radar countermeasures is a search receiver which can detect enemy radar systems and determine their location and their operating frequency. A primary requirement which a search receiver must fulfill is that it cover an extremely wide band, at least a factor of two in frequency, to minimize the necessity of band switching. Microwave search receivers (like radar receivers) are usually of the superheterodyne type, using a crystal mixer. The present paper is concerned with the development of a wide-tuning-range oscillator tube suitable for use as a local oscillator in a search receiver in the microwave frequency range. A search receiver in which the tube was used has been described elsewhere.[1]

Ease and rapidity of tuning, wide tuning range, ease of manufacture, and mechanical ruggedness are important design considerations for a search-receiver local-oscillator tube. There are a number of electronic mechanisms capable of producing oscillations in the desired frequency range, but after careful study the single-cavity focused-beam reflex oscillator was chosen as most nearly meeting the above requirements. The resonant cavity is a coaxial structure, almost wholly external to the evacuated portion of the tube. Connection between the tube proper and the resonator is made by means of copper disk seals. Fig. 1 shows how the tube is placed in its associated resonator or cavity. The cavity center conductor, a piece of brass tubing, is soldered to the second disk and is mechanically an integral part of the tube. The first disk is clamped to the cavity, making a good high-frequency connection with the outer conductor of the coaxial resonator. Fig. 2 is

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‡ Formerly, Bell Telephone Laboratories Inc., New York 14, N. Y.; now, University of Illinois, Urbana, Illinois.

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a photograph of the tube. This tube has been given the Radio Manufacturers Association code number 2K48.

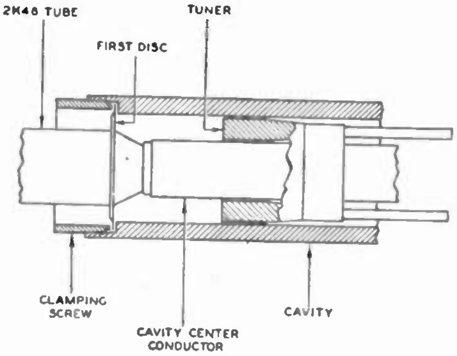

Fig. 1—Cross-section of coaxial cavity with wide tuning range, showing the method of clamping the tube.

Fig. 3 shows the electronically significant details of the tube. These are an electron gun which directs an electron beam through a high-frequency aperture or gap, two copper disks containing holes which define this gap, and a repeller electrode which produces a reflecting and focusing field to redirect the electron beam through the high-frequency gap after a critically valued time of transit in the reflection region. The determination of the optimum shape and size of these electrodes, and the mechanical problems associated with their construction and assembly to the required tolerances, formed a major part of the development work.

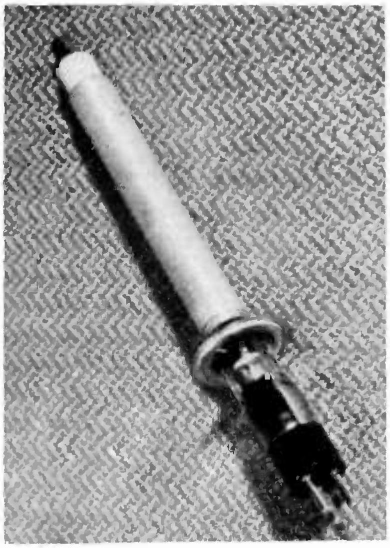

Fig. 2—The wide-range oscillator tube.

As the mechanism of velocity variation and reflection grouping is now well understood,[2] no detailed analysis of the theory will be given. Certain aspects of the theory relating primarily to these particular tubes do, however, warrant a brief consideration.

As mentioned above, the electron stream traverses the radio-frequency gap, is then retarded and reflected with a critically valued delay or transit time, and retraverses the radio-frequency gap. The beam is varied in velocity on its first transit of the gap. It then becomes "bunched" as a result of the shortened transit time of the slow electrons, which are more promptly reflected by the repeller field. Finally, the bunched beam retraverses the radio-frequency gap where it delivers energy to the radio-frequency field.

The electron transit time in the reflection region is critically valued and may be controlled by the potential of the repeller. However, the repeller field must also focus the reflected beam on the gap, which requires a definite repeller potential, conflicting with the transit-time requirement. A compromise is, therefore, necessary.

There exists a number of discrete values of the repeller potential for which oscillations at any specified frequency can be obtained. These values, designated as repeller modes, correspond to electron transit times in the repeller field centering around values given by

$$\theta = N + 3/4$$

where θ is the number of complete cycles at the operating frequency which correspond to the mean transit time, and N is any integer. Only certain of these modes will operate with sufficient power to be useful, and the range in frequency over which they are useful will vary from one to another. The most useful tube would be one in which one mode is particularly strong and in which this one mode persists over the entire frequency range to be covered. The situation is further complicated by the fact that there exists a multiplicity of different frequencies at which any given cavity will oscillate with the tuning adjustments in a fixed position. These different frequencies lie in different ranges which, for the coaxial-line cavity, correspond to operation in its $\lambda/4$, $3\lambda/4$, $5\lambda/4$, etc., ranges (where λ is the wavelength). While only the first two of these ranges are useful, operation in any other than the one desired will interfere with the use of the tube. The principal problem in the construction of a wide-range oscillator is, therefore, related to the proper distribution of the repeller modes and of the cavity ranges, so that unfortunate combinations of these will not give overlapping operating regions and thus interfere with the continuous coverage of the desired frequency ranges.

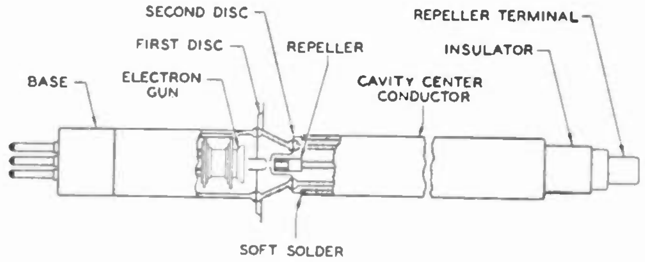

Fig. 3—Partial cross-section view of the tube.

The repeller must perform two distinct functions. It must reflect the electrons back into the interaction gap, and it must provide the correct electron transit time so that the electrons will return in the proper phase to

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

sustain oscillations. These requirements are not mutually compatible, since the first indicates a fixed repeller voltage while the second requires that this voltage be varied with the wavelength. For ease and convenience in operation an additional requirement is imposed, namely, that this change in voltage with wavelength be not too rapid, as this would require the use of an unreasonably large number of repeller modes to cover the required tuning range, since each repeller mode is useful only over a restricted range. The long-wavelength limit is set by the fact that the repeller must always be negative. The increase in cavity losses at short wavelengths is the principal cause in determining the short-wavelength end of the mode. Loss of focus may limit the mode at either the long-wavelength end or the short end, or both, depending upon the repeller design.

Finally, the returned electrons must not be allowed to re-enter the gun region or an undesirable hysteresis effect will be produced. This effect is manifested by a shift in the repeller-voltage range required for operation, depending upon the direction from which this range is approached. In the 2K48 this effect has been minimized by the use of a different hole size in the two gap electrodes.

The phenomenon referred to as electronic tuning, and usually associated with reflex oscillators, has been reduced to a negligibly small amount in the 2K48 as a result of the relatively short transit times employed and of the fairly high Q of the resonant cavities used.

The hole in the first disc should be as small as possible, limited only by the diameter of the electron beam. It has not proved practical to make this hole smaller than 0.030 inch. The hole in the second disc should be somewhat larger; the ideal situation would be for all electrons emitted from the cathode to pass through the hole in the first disc, then to be reflected by the repeller in such fashion that all pass through the hole in the second disc but none pass through that in the first. A number of diameters for the hole in the second disc were tried; 0.050 inch proved to be the most favorable value.

A series of experimental tubes was built to determine the most favorable operating current and voltage for the tube. An electron gun was then designed specifically for the tube. The principles employed in the design of such guns have been described elsewhere.[3] The method used involved the computation of the potential function from cathode to anode which would give the desired perveance, and the use of an electrolytic tank to determine the corresponding electrode shapes. Considerable care was taken to make the electrode shapes simple, to avoid critical dimensions, and to use stock sizes of tubing.

The problem of determining cavity dimensions is, in a sense, a circuit problem; but in this, as in any microwave tube, the tube and its circuit are so intimately related that the tube designer must give much detailed thought to the cavity dimensions to insure that the structure can be built and that it will have the required tuning range. The problem is really one of geometry; that is, to get a structure which will resonate over the required range, and which will have room in it for the necessary electronic mechanism to make the device oscillate.

In a cavity of the coaxial-conductor type, the diameter of the outer conductor is set by two conflicting requirements. Its maximum diameter is limited by the existence of parasitic resonant modes in which the field varies around the periphery of the cavity. Its minimum diameter is limited by loss considerations and by the severe mechanical problems of tube and tuner design which are encountered. A typical tuning curve is given in Fig. 4.

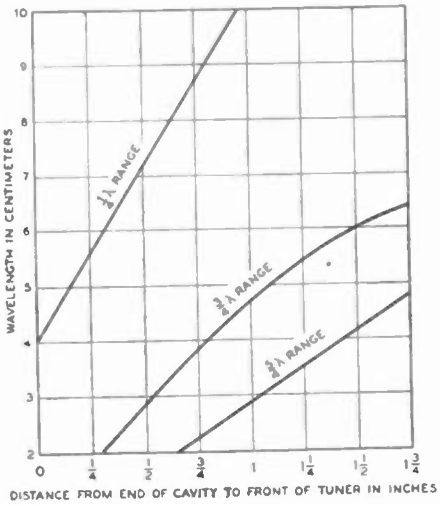

Fig. 4—A typical tuning curve.

Since the tuner is not a part of the tube, only enough consideration was given to it to enable the tube design to be completed. Most of the testing work has been done using a combination sliding-contact and resonant-choke tuner. Although this tuner is designed to resonate at about 3.2 centimeters, it has been used with good results at wavelengths varying from 2.3 to 11 centimeters.

The tube which is shown in Fig. 2 is capable of oscillating at any wavelength between 2.3 and 15 centimeters. It is impractical, however, to make a single cavity to scan over this range, because of the very formidable difficulty of making at uner and output-coupling mechanism which are useful over the entire range. Accordingly, two cavities are used. One is intended to be used between 3 and 6 centimeters wavelength and is used in its $\frac{3}{4}$-wavelength range so as to get short wavelengths without excessively small dimensions. The other is a $\frac{1}{4}$-wavelength cavity and tunes from 5 to 10 centimeters. In terms of frequency these cavities are useful from 5000 to 10,000 megacycles, and from 3000 to 6000 megacycles, respectively. Thus the entire spectrum from 3000 to 10,000 megacycles can be covered by a two-band receiver.

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[3] A. L. Samuel, "Some notes on the design of electron guns," Proc. I.R.E., vol. 33, pp. 233–240; April, 1945.

A Coaxial-Type Water Load and Associated Power-Measuring Apparatus*

R. C. SHAW†, SENIOR MEMBER, I.R.E., AND R. J. KIRCHER‡, SENIOR MEMBER, I.R.E.

Summary—**This paper presents a description of a coaxial-type water load and associated equipment suitable for measuring peak pulse powers of the order of a megawatt. Water-cell loads have been designed to operate at wavelengths of from 10 to 40 centimeters, where the average dissipation is of the order of 300 watts. Ordinary tap water is used in the load to dissipate the radio-frequency power.**

DESIGN

TO FACILITATE the testing of pulsed transmitters in the 10- to 40-centimeter transmission band where the ratio of peak power to average power is of the order of several thousand to one, a coaxial-type of water load has been developed with associated power-measuring apparatus. The apparatus is also useful in measuring continuous-wave power of the order of 300 watts.

The water-cell load represented in Fig. 1 consists of an appropriate length of 53-ohm coaxial line having an outside diameter of $1\frac{5}{8}$ inches and wall thickness of $\frac{1}{16}$ inch. The diameter of the center conductor is $\frac{5}{8}$ inch. One end of the line is made watertight with a metallic plug soldered to both inner and outer conductors. The inner conductor extends a short distance beyond this closure, and a water inlet tube is soldered into it axially. A small pipe soldered in the outer conductor near the shorted end provides a water exit. At the other end of the load section the chamber is made watertight by the insertion of a slug of dielectric (micalex or steatite) using the techniques to be described. Inside the center conductor is mounted a copper-encased immersion-type heating element which is insulated from the center conductor by a glass tube which fits around the heater. The heater element permits the substitution of 60-cycle alternating-current power for the radio-frequency power in determining the average power dissipated in the load. Ideally, the water heated by the heating element should not come in contact with the wall of the inner conductor of the coaxial line. Water enters through the axial tube of the center conductor, flows down the length of the line over the heater element, and is admitted to the main chamber through holes in the center conductor a short distance in front of the dielectric slug. Water fills the volume between the coaxial conductors to the length of the chamber and flows out through the exit pipe at the far end. It is in this volume of water that the radio-frequency power is dissipated. Appropriate fittings are used on the center and the outer conductor of this line so that it may be attached directly to the air-dielectric line which connects to the transmitter.

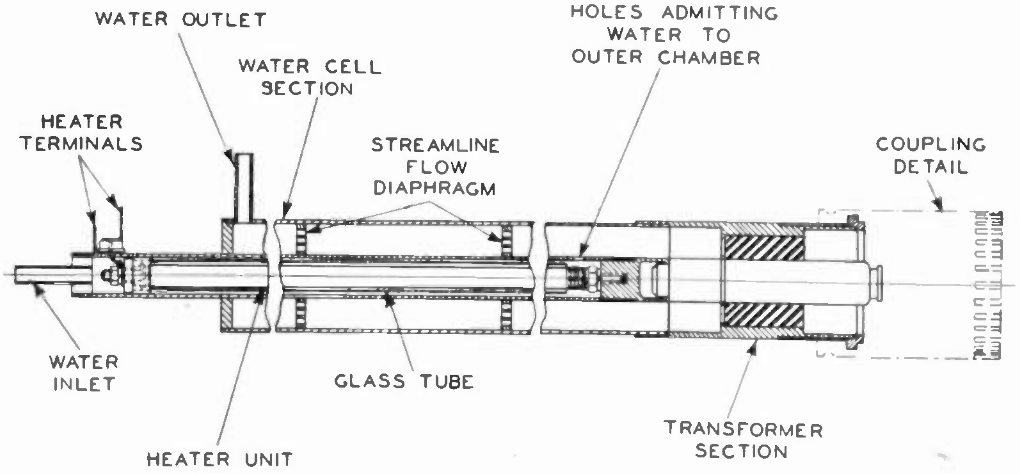

Fig. 1—Coaxial water load.

In order to match the water-filled section of line to the air-filled line, an impedance transformation is made by means of a slug of dielectric material of low loss and high dielectric constant which is electrically a quarter-wave long at the frequency of operation. The surge impedance Z_t of the section of line filled with the dielectric material is related to the surge impedance Z_a of the air-filled line and the surge impedance Z_ω of the water-filled line by the equation

$$Z_t = \sqrt{Z_a Z_\omega} \qquad (1)$$

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‡ Bell Telephone Laboratories, New York 14, N. Y.

and the equation relating the dielectric constants is

$$e_t = \sqrt{e_a e_\omega}. \qquad (2)$$

For a dielectric constant of water of 79, the dielectric constant for the transformer material should be, ideally,

$$e_t = \sqrt{79}$$
$$= 8.89.$$

Two materials suitable for the transformer, steatite and micalex, have dielectric constants with values between 5 and 7. Such materials may be used, however, by reducing the external diameter of the transformer slug, keeping the internal diameter that of the center conductor. This requires a metal sleeve over the transformer, or a corresponding reduction in the diameter of the line at this point. If a sleeve is used, it must be soldered into the external conductor of the transmission line. The surge impedance of the transformer section of the line is given by

$$Z_t = \frac{138 \log_{10} \dfrac{D}{0.625}}{\sqrt{e_t}} \qquad (3)$$

where D is the external diameter of the dielectric material, and 0.625 is the diameter of the inner conductor in inches. The surge impedance of a 53.5-ohm air line when filled with water is

$$Z_\omega = \frac{53.5}{\sqrt{e}}\,. \qquad (4)$$

If the dielectric constant for water is assumed to be 79, the surge impedance of the water line would be

$$Z_\omega = 6.00 \text{ (ohms)}$$

and the required surge impedance of the transformer section of line would be, from (1),

$$Z_t = 18 \text{ (ohms)}.$$

The external diameter of the dielectric material is now obtained from (3) when its dielectric constant e_t is known.

The wavelength in the water-filled line is related to the wavelength in the air-dielectric line by the equation

$$\lambda_\omega = \frac{\lambda_a}{\sqrt{e}} \qquad (5)$$

which, for $e_\omega = 79$, gives

$$\lambda_\omega = \frac{\lambda_a}{8.89}\,. \qquad (6)$$

At a wavelength of 10 centimeters the direct-current conductivity of water appears to have little, if any, effect on the absorption of radio-frequency power. The attenuation realized in the water-filled line is a function of its length and is roughly proportional to the frequency of operation. Calbick has measured values of an attenuation for tap water at 22 degrees centigrade of approximately 1.5 decibels per centimeter at a wavelength of 9.2 centimeters and 0.35 decibel per centimeter at a wavelength of 29.6 centimeters.

The time required to obtain a temperature equilibrium with a given power dissipation in a volume of water which is moving at a fixed rate is a function of the length of the line. In general, the shortest length of water line which will give adequate power attenuation should be used. For frequencies from 700 megacycles up, a length of water cell of approximately 10 water wavelengths (20 water wavelengths for the total attenuation path, direct and reflected) was found to be very satisfactory in that the reflected power returned to the input of the water cell had a negligible effect on the impedance presented to the air-filled line. A transmission bandwidth of several per cent of the operating frequency was realized for a standing-wave ratio of maximum-to-minimum voltage in the air line of 1 decibel at the edges of the band, and approximately one quarter of a decibel at the center of the band. By using auxiliary movable λ/4 transformers in the air line in front of the load, the useful band of operation with a given water load may be increased appreciably.

CONSTRUCTION

A technique was developed which permitted soldering the dielectric slug into place on both inner and outer conductors in order to insure a watertight seal. Both inner and outer peripheral surfaces of the dielectric were first coated with a deposit of platinum, which was then copper-plated. With sufficient care, the copper-plated surfaces could be soldered to the inner and outer conductors. This is a somewhat delicate operation, and, if unsuccessful, the process had to be started all over, the micalex piece being frequently unusable because of damage sustained in its removal. Because of the permanency of the installation, no corrective machining of the material for frequency adjustment is feasible once it is in place. A good degree of uniformity can be obtained with micalex pieces cut from the same stock and held to close tolerances on their dimensions. The precise machining of this material, which is easily damaged, is a drawback to its use.

In order to simplify the above process and obtain a more reliable method in the transformer-assembly procedure, molded steatite (F66) forms were obtained which could be dressed to close tolerances on all dimensions. The dielectric constant for this material at low frequencies was approximately 6.0±0.1. Experimental data at frequencies of 750 to 1250 megacycles indicated an apparent dielectric constant of approximately 5.30.

A further simplification in the design was to construct a brass outer housing for the steatite piece and a corresponding brass inner-conductor detail. This allowed the transformer section to be readily attached or removed from the water-load section. The inner conductors of the transformer and the water-load section were fitted together with a plug-and-jack type of connection. The outer conductors of these sections were also slipped one into the other and soft-soldered around the periphery to make a watertight connection.

As shown in Fig. 1, the inner and outer conductor details have, in each case, a shoulder on the periphery against which the steatite detail is pressed. No more than one-thousandth of an inch clearance is allowed between the inner and the outer conductors and the steatite. Such a close tolerance requires hand-fitting of the steatite piece to the inner and outer conductors. Before proceeding with the assembly the surfaces which come in contact are well cleaned, first with alcohol and then with a clean damp cloth. After cleaning it was necessary to avoid touching the surfaces with the fingers. In assembling, the inner and outer annular surfaces of the steatite were covered with a thin, uniform layer of industrial rubber cement and the piece was then pushed into place, first on the center conductor and then into the outer conductor. A drying period of twenty-four hours is recommended, after which a water pressure of not more than 8 pounds may be applied. The steatite may be removed from its shell by the application of 15 to 20 pounds of air pressure or by soaking the unit in benzene for several hours.

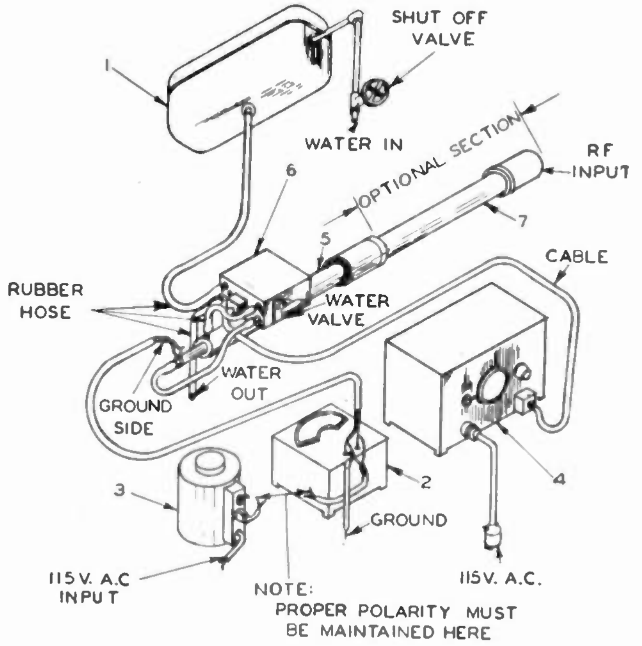

Fig. 2—Schematic of water load and associated equipment. Legend: (1) Water-head control chamber. (2) Voltmeter. (3) Variac. (4) Bridge box. (5) Coaxial water load. (6) Thermistor unit. (7) Auxiliary section of 53-ohm coaxial line.

With a constant water flow, the effect of varying the amount of power dissipated in the water on the smoothness of the impedance match obtained is comparatively small. Some variation in impedance match is noted over the range of zero to 200 watts dissipated in the water chamber because the absorption of power depends slightly on the temperature of the water. The conclusion is that, if the termination is good at very low average power inputs, it remains good for the higher average powers obtained in transmitter testing. These loads have been used to dissipate peak powers in excess of 1 megawatt and having an average dissipation of 300 watts or more.

MEASUREMENT

At this point the means for measuring the power supplied to the water load will be described. Refer to Fig. 2 for a pictorial view of the apparatus used. A reasonably constant flow of water is required which can be adjusted from a minimum rate of 7.5 cubic centimeters per second to a maximum rate of 15 cubic centimeters per second. Ordinary tap water is used to supply a "constant-head" chamber at eight to ten feet above the water load. A float valve maintains the water in the chamber at nearly constant level. The water is brought through small rubber tubing to a small box attached to the outside of the water load. From this point the water passes through a plastic tube with a thermistor unit inserted in the flow; then the water passes through a stopcock. This stopcock may be turned to supply the inlet to the water load, or to by-pass the water load.

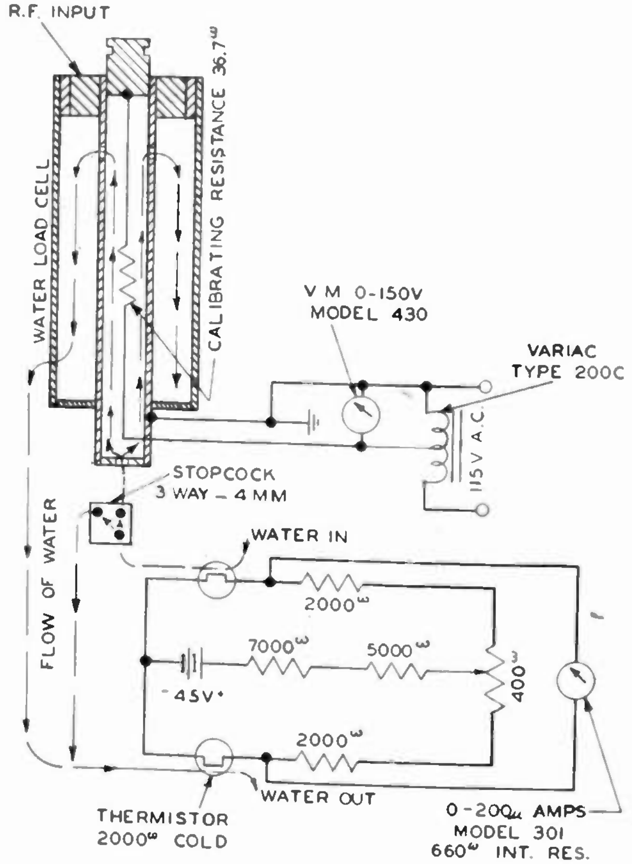

Fig. 3—Schematic of the power-measuring circuit.

Whether or not the load is by-passed, the water passes through an exit plastic tube in which a second thermistor is located.

The thermistor units of approximately equal resistance are in adjacent arms of a bridge circuit. The bridge is balanced for the same inlet and outlet water temperature with the stop-cock turned so that the water by-passes the water load. The stop-cock is then turned so that the water flows through the water load at a rate of about 10 cubic centimeters per second. A small temperature difference may be noted without power applied until the material of the water load is substantially at the temperature of the inlet water.

The schematic of the water circuit and the bridge circuit is given in Fig. 3. The power supply for the bridge may be either a battery or a built-in rectifier.

Alternating-current power may be supplied to the heater within the water load at approximately the value of the average power expected during the radio-frequency operation. A thermal lag of from one and one-half to three minutes occurs before an equilibrium condition is reached, depending on the power dissipated, the water flow, and the degree of thermal interchange between input and output water.[1] By means of a sensitivity control on the bridge circuit, a convenient reading may be obtained on the microammeter indicator for the power level being measured. A calibration may be obtained for the power range expected by taking readings for several values of power. Care must be taken that neither the water flow nor the sensitivity of the bridge be changed in using the calibration in subsequently measuring the radio-frequency power in the load. This method is useful over short intervals of time, and requires that the temperature of the inlet water remains constant at the value for which the bridge was balanced.

A direct-substitution method of power measurement is less subject to error. In this method the radio-frequency power is dissipated in the load until equilibrium is obtained; then the bridge meter reading is noted. The radio-frequency power is turned off, and 60-cycle power applied to the water-load heater and adjusted until the same meter reading is obtained. The average 60-cycle power required is equal to the average radio-frequency power.

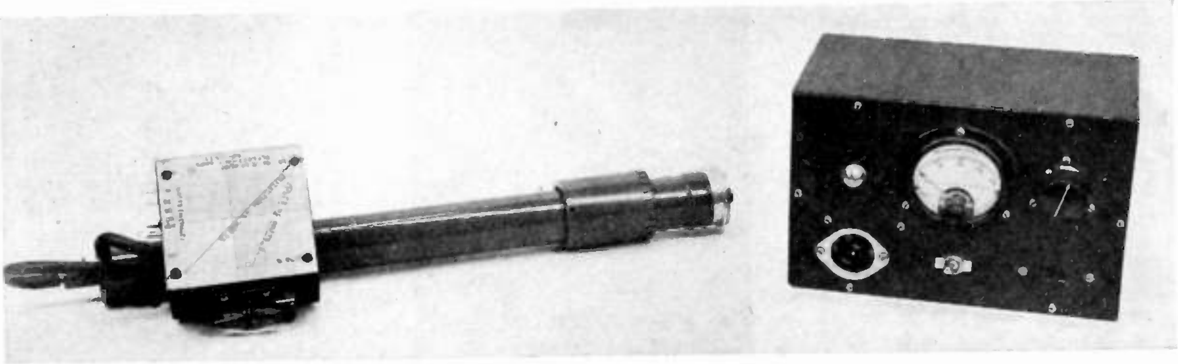

Fig. 4—A view of the water load, thermistor unit, and bridge box.

The conversion of average radio-frequency power to peak radio-frequency power involves the factors of pulse shape and repetition rate. The determination of such factors is beyond the scope of this paper. When these factors are accurately known, the accuracy of measurement of peak powers is between 5 and 10 per cent.

A photograph of the water-load section, thermistor box, and the bridge box is shown in Fig. 4. The water load shown was designed for operation at a wavelength of approximately 30 centimeters.

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[1] C. J. Calbick of the Bell Telephone Laboratories has shown in an unpublished memorandum that a heat interchange between the ingoing and the outcoming water should be minimized to avoid the possibility of an error in making power measurements due to the occurrence of an excessive time interval before obtaining a stable power indication.

Contributors to Waves and Electrons Section

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Raymond E. Lacy (SM'46) was born on March 17, 1916, at Camden, New Jersey. He was an undergraduate at Drexel Institute of Technology in Philadelphia, Pennsylvania, from 1933 to 1938, during which period he also served as a student engineer for five half-year periods with various concerns including the Potomac Electric Power and Light Company, Washington, D. C.; Philco Corporation, Philadelphia; and Chubbuck and Patrick, consulting engineers, Philadelphia. After receiving the B.S.E.E. degree in 1938 from Drexel, he joined the staff of New York University College of Engineering as graduate assistant in electrical engineering, in which capacity he served for two years. He received the M.E.E. degree from the N.Y.U. Graduate School in 1940. He joined the staff of the Signal Corps Laboratory at Fort Monmouth, New Jersey, as a radio design engineer, in 1940, and has been associated in the design of Army ground radio communication equipment during the ensuing period.

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Dr. Larsen was instructor in electrical engineering from 1937 to 1940 at Michigan College of Mining and Technology. In 1940 he became assistant professor, a post which he held until 1943, when he became associated with the research department of Stromberg-Carlson Company, in Rochester, New York, where he has remained to date.

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Lynn L. Merrill was born at Erin, New York, on October 19, 1902. He received the B.S. degree in electrical engineering at Clarkson College of Technology in 1924 and remained at Clarkson teaching mathematics until 1929. He then joined the General Electric Company at Schenectady and took the "A" portion of the advanced course in engineering. In 1930 he transferred to Pittsfield, where he worked as designer in the high-voltage bushing department and on theoretical research for the transformer department and the high-voltage laboratory. In 1932 he resumed graduate study at Rensselaer Polytechnic Institute, receiving a Ph.D. in mathematics in 1936 and teaching mathematics at Rensselaer from 1933 to 1944. Since 1944 he has been employed as mathematician in the research department of the Stromberg-Carlson Company.

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Arthur L. Samuel (A'24–SM'44–F'45) was born on December 5, 1901, at Emporia Kansas. He received the A.B. degree from the College of Emporia in 1923; the degrees of S.B. and S.M. in electrical engineering from the Massachusetts Institute of Technology in 1926; and has done additional graduate work both at M. I. T. and at Columbia University. Recently, he was awarded the honorary degree of Sc.D. from the College of Emporia.

He was employed by the General Electric Company intermittently from 1923 to 1927, and was an instructor in the electrical engineering department at M. I. T. from 1926 to 1928. Dr. Samuel joined the technical staff of the Bell Telephone Laboratories in 1928, where he was engaged in electronic research and development. From 1931 to 1946 his principal interest has been in the development of vacuum tubes for use at ultra-high frequencies. Dr. Samuel, at present, is a professor in the department of electrical engineering at the University of Illinois. He is a member of the American Physical Society, the American Institute of Electrical Engineers, and the American Association for the Advancement of Science.

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Harold E. Dinger (A'27–M'43–SM'43) was born on May 7, 1905, at Barberton, Ohio. He studied at the University of Akron and the University of Maryland. From 1917 until the present time he has been active in amateur and experimental activities. In 1921 he engaged in radio service work. From 1929 to 1939 Mr. Dinger was chief instructor at the McKim Technical Institute, and from 1939 to 1940 he was transmitter engineer for the Ohio Broadcasting Company. Since 1940 he has been serving as radio engineer at the Naval Research Laboratory, specializing in wave-propagation measurements, radiation-pattern measurements, and radio interference elimination. He also served as co-ordinator for the radio interference elimination program of the Bureau of Ships, and as Navy Department liaison for shipboard aspects on the NDRC Interference Reduction committee. He received the Navy "Meritorious Civilian Service Award" for work in connection with interference reduction aboard amphibious vehicles and landing vessels. Mr. Dinger is a member of the Combined Services committee for the standardization of filters and filter test methods, and a Navy Department representative for the American Standards Association subcommittee on radio noise meters working under the war committee on methods of measuring radio noise, C 63. He is an I.R.E.-Navy representative on the American Institute of Electrical Engineers electronic-heating committee.

Harold E. Dinger

⁂

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Harold G. Paine (A'43–M'46) was born on August 24, 1908, at Marshall, Oklahoma. He received the B.S. degree in electrical engineering from the University of Illinois in 1930. Beginning in 1930, he was employed as an electrical engineer by the Wagner Electric Corporation and later by the Carolina Power and Light Company. From 1942 until the present time, Mr. Paine has been serving as radio engineer at the Naval Research Laboratory engaged in research on problems connected with radio noise and field-intensity meters.

Abstracts and References

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ACOUSTICS AND AUDIO FREQUENCIES

534+536]:538.569.4.029.64 **3509**

Thermal and Acoustic Effects Attending Absorption of Microwaves by Gases—Hershberger. (*See* 3585.)

534.21 **3510**

The Propagation of Sound in the Atmosphere—Attenuation and Fluctuations—V. O. Knudsen. (*Jour. Acous. Soc. Amer.*, vol. 18, pp. 90–96; July, 1946.)

534.21 **3511**

Propagation of Sound Over Absorbent Surfaces—R. K. Cook. (*Jour. Acous. Soc. Amer.*, vol. 18, p. 252; July, 1946.) Summary of an American Acoustical Society paper.

534.212:534.321.9 **3512**

The Directional Characteristics of a Free-Edge Disk Mounted in a Flat Baffle or in a Parabolic Horn—F. H. Slaymaker, W. F. Meeker, and L. L. Merrill. (*Jour. Acous. Soc. Amer.*, vol. 18, p. 251; July, 1946.) When a parabolic horn is excited by a diaphragm at supersonic frequencies, the sharpest beams are obtained with the diaphragm radiating most of its energy towards the side walls. The differential equation for the vibration of a thin free-edge circular disk is solved, and the results tabulated so that the dynamic curve and directional radiation pattern can be calculated for any disk and frequency. Summary of American Acoustical Society paper.

534.32 **3513**

A Hundred-Element Tone Synthesizer—E. C. Wente, C. A. Lovell, and J. F. Muller. (*Jour. Acous. Soc. Amer.*, vol. 18, p. 253; July, 1946.) A complex electric current is generated by a combination of up to one hundred sine-wave currents derived from magnetic records, each of which is variable in amplitude and substantially free from harmonics. Summary of American Acoustical Society paper.

534.321.9:534.241 **3514**

Echo Formation on Simple Surfaces—C. E. Mongan Jr. (*Jour. Acous. Soc. Amer.* vol. 18, p. 255; July 1946.) A laboratory technique has been devised for the observation of supersonic echo formation at plane cylindrical and spherical discontinuities. Summary of American Acoustical Society paper.

534.321.9:549.514.51 **3515**

Refinements in Supersonic Reflectoscopy. Polarized Sound—F. A. Firestone and J. R. Frederick. (*Jour. Acous. Soc. Amer.* vol. 18, pp. 200–211; July 1946.) Supersonic waves of various types can be radiated by suitable excitation of quartz crystals and can be used in a reflectoscope for the detection of flaws in metals by the reflection they produce. The reflectoscope has also been used extensively for studying the laws of reflection and refraction of polarized sound in solids.

534.41+534.781 **3516**

The Portrayal of Visible Speech—J. C. Steinberg and N. R. French. (*Jour. Acous. Soc. Amer.* vol. 18, pp. 4–18; July 1946.) The object is to produce a satisfactory time-intensity-frequency (three-dimensional) speech pattern. The best scales for such a diagram and other possible developments suggested by the physiological characteristics of speech and reading are discussed. Methods of evaluating pattern legibility are also suggested.

534.41+534.781 **3517**

The Sound Spectrograph—W. Koenig H. K. Dunn and L. Y. Lacy. (*Jour. Acous. Soc. Amer.*, vol. 18, pp. 19–49; July, 1946.) The mechanical arrangements and electrical circuits are described. Time-intensity-frequency records of a wide variety of sounds are shown and the problems arising in their analysis are discussed.

534.41+534.781 **3518**

Basic Phonetic Principles of Visible Speech—G. A. Kopp and H. C. Green. (*Jour. Acous. Soc. Amer.*, vol. 18, pp. 74–89; July, 1946.) The visible characteristics of the various fundamental phonetic sounds, classified according to their method of production, are shown pictorially and discussed. Racial and individual differences of speech are also considered.

534.41+534.781]:535.37 **3519**

Visible Speech Translators with External Phosphors—H. Dudley and O. O. Gruenz, Jr. (*Jour. Acous. Soc. Amer.*, vol. 18, pp. 62–73; July, 1946.) Speech patterns, analyzed on an intensity-frequency-time basis, are displayed on an endless moving phosphorescent belt. Any pattern is visible for about 1½ seconds after which it is erased by red lamps.

534.41+534.781]:621.385.832 **3520**

Visible Speech Cathode-Ray Translator—R. R. Riesz and L. Schott. (*Jour. Acous. Soc. Amer.*, vol. 18, pp. 50–61; July, 1946.) The special cathode-ray tube has a persistent screen in the form of a cylindrical band, and is rotated about a vertical axis. The electron beam always excites the screen in the same vertical plane, so the speech patterns, portrayed both as a spectrum and as a pitch analysis, appear along a horizontal time axis. The translator has been used to study patterns of speech sounds and their combinations. Trained observers can interpret the pictures of conversational speech at 90 to 120 words per minute.

534.417:534.88 **3521**

Echo Ranging Sonar [Model QCS/T]—Evans. (*See* 3905.)

534.7 **3522**

Effects of Distortion on the Intelligibility of Speech at High Altitudes—G. A. Miller and S. Mitchell. (*Jour. Acous. Soc. Amer.*, vol. 18, p. 250; July, 1946.) "Sounds spoken into the closed cavity of an oxygen mask at high altitudes differ substantially in quality and intensity from normal speech at sea level." The most satisfactory method of correcting this appears to be sharp, symmetrical limiting ("peak clipping"). Summary of American Acoustical Society paper.

534.7 **3523**

The Effects of High Altitude on Speech and Hearing—H. W. Rudmose, K. C. Clark, F. D. Carlson, J. C. Eisenstein, and R. A. Walker, (*Jour. Acous. Soc. Amer.*, vol. 18, pp. 250–251; July, 1946.) Measurements of speech energy, made at simulated altitudes from sea level to 40,000 feet show that the energy decrement of vowels and semivowels, expressed in decibels, varies uniformly with altitude but there is little change in unvoiced consonants, or the threshold of hearing. Summary of American Acoustical Society paper.

534.78 **3524**

Effects of Amplitude Distortion upon the Intelligibility of Speech—J. C. R. Licklider. (*Jour. Acous. Soc. Amer.*, vol. 18, p. 249; July, 1946.) Amplitude distortion of various forms was introduced into otherwise high-quality audio system by the use of nonlinear circuits. Tests indicated that cutting of peaks, resulting from overdriving the amplifiers, had little detrimental effect on intelligibility whereas cutting of the low amplitude part of the speech had a serious effect. The reduction in intelligibility caused by distortion increases when noise is mixed with the speech, except when the noise consists of pulses and the distortion limits them. Summary of Acoustical Society paper.

534.78 **3525**

Effects of Frequency Distortion upon the Intelligibility of Speech—J. P. Egan and F. M. Wiener. (*Jour. Acous. Soc. Amer.*, vol. 18, pp. 249–250; July, 1946.) Bandwidths used for frequency-distortion tests ranged from half an octave to the whole of the speech frequency range. Articulation increased with the intensity level of the received speech, but less rapidly the smaller the bandwidth. When the speech is mixed with a uniform spectrum of noise, and speech and noise are together passed through a filter, articulation differs very little from that obtained by filtering the speech only, before mixing with the noise. When, however, the noise level falls rapidly as a function of frequency, articulation can be improved by filtering both noise and speech. Summary of American Acoustical Society paper.

534.78 **3526**

Correction of the Audio Characteristics of Communication Systems with Measured Articulation Scores—L. L. Beranek. (*Jour. Acoust. Soc. Amer.*, vol. 18, p. 250; July, 1946.) A method has been devised for determining the ability of a communication system to transmit speech intelligibly. "The results indicate that it is possible to estimate from physical data the performance of a system in an assumed ambient noise field." Summary of American Acoustical Society paper.

534.78 **3527**

The Masking of Speech by Sine Waves, Square Waves, and Regular and Randomized Pulses—S. S. Stevens, J. Miller, and I. Truscott. (*Jour. Acous. Soc. Amer.*, vol. 18, p. 250; July, 1946.) The most effective frequency for the masking of speech by sine waves is about 300 cycles for intense waves and 500 cycles for weak waves. Square waves are approximately equally effective between 80 cycles and 400 cycles. Regular pulses with a duration of 10 microseconds are most effective at about 200 pulses per second, but when the pulses occur at random time intervals the masking is greatly increased. Summary of American Acoustical Society paper.

534.84 **3528**

Note on Normal Frequency Statistics for Rectangular Rooms—R. H. Bolt. (*Jour. Acous. Soc. Amer.*, vol. 18, pp. 130–133; July, 1946.) A graphical method for determining the average number and spacings of normal frequencies, up to a given frequency, for a rectangular room of known dimensions.

534.84 **3529**

The Design and Construction of Anechoic Sound Chambers—L. L. Beranek and H. P. Sleeper, Jr. (*Jour. Acous. Soc. Amer.*, vol. 18, pp. 140–150; July, 1946.)

534.844.1 **3530**

The Measurement of Reverberation—W. Tak. (*Philip Tech. Rev.*, vol. 8, pp. 82–88; March, 1946.) In rooms and auditoria. Theoretical discussion of principles of measurement; apparatus used will be described in a subsequent article.

621.394/.396].645 **3531**

The Multiamp—Jackson. (*See* 3559.)

621.395.61:534.64 **3532**

Motional Impedance Analysis Applied to a Dynamic Microphone—J. E. White. (*Jour. Acous. Soc. Amer.*, vol. 18, pp. 155–160; July, 1946.) A method whereby the pressure sensitivity at low-audio frequencies, the microphone mechanical impedance, and other characteristics may be calculated.

621.395.613.38 **3533**

Magnetic Throat Microphones of High Sensitivity—D. W. Martin. (*Jour. Acous. Soc. Amer.*, vol. 18, p. 253; July, 1946.) Summary of American Acoustical Society paper.

621.395.623.6 **3534**

Development of Midget Earphones for Military Use—H. A. Pearson, A. B. Mundel, R. W. Carlisle, W. Knauert, and M. Zaret. (*Jour. Acous. Soc. Amer.*, vol. 18, p. 253; July, 1946.) Summary of American Acoustical Society paper.

621.395.623.7 **3535**

The Output Stage—Stanley. (*See* 3727.)

621.395.623.8:621.396.932 **3536**

Marine Loudspeaking Gear—P. Hickson. (*Wireless World*, vol. 52, pp. 254–255; August, 1946.) Description of Ardente "Loud Hailer" Type 431, which was designed for intership communication to work off a 12 volt accumulator, high voltage being supplied by a small motor generator. The audio amplifier gives an output of 15 watts, which was found sufficient for the purpose and enabled the apparatus to be kept small; it weighed 23 pounds. The frequency response curve of the amplifier is flat from 300 to 10,000 cycles, attenuation being introduced at the lower end to match the characteristics of the loudspeaker unit.

621.395.667 **3537**

Fundamental Tone Control Circuits—A Reference Sheet—(*See* 3728.)

621.395.92 **3538**

Desirable Frequency Characteristics for Hearing Aids—H. Davis, C. V. Hudgins, G. E. Peterson, and D. A. Ross. (*Jour. Acous. Soc. Amer.*, vol. 18, p. 247; July, 1946.) Characteristics investigated included one with uniform acoustic gain from 100 to 7,000 cycles, and others with a slope of either 6 decibels or 12 decibels per octave towards either the high or the low frequencies. Tests on hard-of-hearing subjects showed that the best articulation was achieved with the flat characteristic or one with a 6 decibel per octave slope increasing towards the high frequencies. This was also true in the presence of static noise. Summary of American Acoustical Society paper.

AERIALS AND TRANSMISSION LINES

621.392 **3539**

A Fractional Termination for Ladder Networks—W. R. LePage. (*Trans. A.I.E.E.* (*Elec. Eng.* August–September, 1946), vol. 65, pp. 530–536; August–September, 1946.) A theoretical survey of the possibilities of using a new termination for a uniform ladder network, which renders it more nearly like its corresponding circuit of distributed constants than either the mid-shunt or mid-series terminated types, is given. The improvement, which holds only over a limited frequency range, is small, becoming less as the number of sections is increased.

621.392 **3540**

Anomalous Attenuation in Waveguides—D. A. Bell. (*Wireless Eng.*, vol. 23, pp. 287–288; October, 1946.) A note on a pictorial and qualitative method of showing the variation with frequency of attenuation in waveguides, prompted by a paper by Kemp. A correction to this paper appears at the bottom of page 289; the original paper was abstracted in 2818 of October.

621.392 **3541**

A Theory of the Narrow Resonant Slit in a Wave Guide Partition—E. S. Akeley. (*Phys. Rev.*, vol. 69, p. 697; June 1–15, 1946.) "The ratio r of the amplitude of the transmitted radiation at infinity to the incident amplitude is determined. Resonance occurs at $k=\pi/2L$ for infinitely narrow slit and $r=1$ at resonance. These results are independent of the position of slit in partition. Formulas are given for the determination of the shape of the resonance maximum and shift of same from $k=\pi/2L$." Summary of American Physical Society paper.

621.392:538.566 **3542**

The Scattering of Electromagnetic Radiation by a Thin Circular Ring in a Circular Wave Guide—P. Feuer and E. S. Akeley. (*Phys. Rev.*, vol. 69, p. 697; June 1–15, 1946.) "The ratio of the scattered field to the incident field at a great distance from the obstacle is computed and also the position and shape of the resonance maximum." Summary of American Physical Society paper.

621.392:621.315.1 **3543**

Flexible Wave Guides—A. R. Anderson and A. M. Winchell. (*Electronics*, vol. 19, pp. 104–109; August, 1946.) Three types of flexible waveguides are described, the first consisting of a metal strip wound spirally about a rectangular former; the second, a series of circular sections held by an internally ribbed synthetic rubber jacket; and the third, a thin-walled, seamless corrugated rectangular tube. The construction of each is given in detail and their characteristics tabulated. Their possible uses are discussed and the mechanical limitations of each type are considered. There is appreciable leakage of energy from the second type which also suffers from the disadvantage of restricted frequency coverage.

621.396.67 **3544**

More about Aerials—Polarization: Gain: Reflectors—"Cathode Ray"—(*Wireless World*, vol. 52, pp. 251–253; August, 1946.) An elementary discussion of electromagnetic fields and the above—mentioned properties of aerials. For part 1 see 2821 of October.

621.396.67.013 **3545**

The Magnetic Antenna—L. Page. (*Phys. Rev.*, vol. 69, pp. 645–648; June 1–15, 1946.) The antenna investigated is an infinitely long cylindrical rod (radius a, permeability μ, permittivity κ) with a few turns wound around the center. The increase of flux is given by equation (14) and is shown graphically as a function of λ/a in Figs. 1 and 2. For constant κ there is a critical wavelength approximately proportional to $\sqrt{\mu}$, at which the flux ratio exhibits a sharp maximum, while for constant μ the flux ratio increases strongly with increase of κ. A $\kappa=25$, $\mu=100$, and $a=2$ centimeters, then $\lambda_m=3$ meters, and the value of the flux ratio is 7.2. This effect has a possible practical application.

621.396.67.08.011.2 **3546**

Antenna Impedance Measurement—King. (*See* 3984.)

621.396.671 3547

The Cylindrical Antenna: Current and Impedance—R. King and D. Middleton. (*Quart. Appl. Math.*, vol. 4, pp. 199–200; July, 1946.) Correction to 1453 of June.

621.396.677 3548

A Comparison of the Efficiencies of Rhombic Type Aerials—I. M. Ruschuk. (*Vestnik Elektropromyshlennosti*, no. 2, pp. 13–18; 1946. In Russian.) Mathematical discussion showing that the gain of a variation of the broadside rhombic aerial, as proposed by Eisenberg, is 1.8 times higher on short waves and 1.53 times higher on long waves than in the case of an ordinary rhombic aerial.

621.396.679.4.012.2 3549

Solving Feeder Problems Graphically—R. E. Kelley. (*QST*, vol. 30, pp. 25–27, 140; September, 1946.) Explains the use of a circle diagram to calculate the impedance of a line for given loads and for given lengths.

CIRCUITS

621.3.014/.015].33:517.942.82 3550

Switching Problems and Instantaneous Impulses—J. C. Jaeger. (*Phil. Mag.*, vol. 36, pp. 644–651; September, 1945.) Discussion of the application of the Laplace transform to circuits or linear systems in which an instantaneous change of conditions is imposed; the work of Ghizzetti (*R. C. Circ. Math. Palermo*, vol. 61, p. 339; 1937) is critically considered. The paper includes a treatment of the case of an impulsive current or electromotive force having a very large magnitude and very small duration, but finite time-integral.

621.314.015.33 3551

Pulse Transformer Ratings Based on Energy Considerations, and Methods of Design Based on Thermodynamical Considerations—W. H. Bostick. (*Phys. Rev.*, vol. 69, p. 697; June 1–15, 1946.) Summary of American Physical Society paper.

621.315.14.011.3/.4 3552

Geometric Mean Distances for Rectangular Conductors—H. B. Dwight. (*Trans. A.I.E.E.* (*Elec. Eng.* August–September, 1946), vol. 65, pp. 536–538; August–September, 1946.) Values of geometric mean distances are given in graphical form for the calculation of reactance of parallel conductors of rectangular cross-section, for a wide range of dimensions and spacings. The formulas used and method of application to practical cases are illustrated by examples.

621.392 3553

A Fractional Termination for Ladder Networks—LePage. (*See* 3539.)

621.392:621.315.59 3554

Nonlinear Circuit Element Applications—H. E. Kallmann. (*Electronics*, vol. 19, pp. 130–136; August, 1946.) Applications considered comprise bridge circuits in which nonlinear elements are combined with ordinary resistors to improve power supply regulation; use as limiters to facilitate pulse squaring; and logarithmic scaling. The application of the latter in electrical computing operations is briefly described.

621.392.012.2.029.64 3555

A Circle Diagram for Resonant Microwave Systems—W. Alter. (*Phys. Rev.*, vol. 69, p. 697; June 1–15, 1946.) "The diagram permits the mapping, in the complex plane of load impedances or reflection coefficients, of such contour lines as loaded resonator *Q*, frequency pulling, and the circuit efficiency." Summary of American Physical Society paper.

621.394/.397].645 3556

Minimal Noise Amplifiers—E. J. Schremp. (*Phys. Rev.*, vol. 69, pp. 695–696; June 1–15, 1946.) "The present unified method, employing a simple principle valid at all frequencies, consists of two successive steps: (1) for each independent internodal tube noise generator, apply the constraint that the ratio vanish between its output response and that of the signal source; and (2) introduce small variations from the resulting constraints, to secure a compromise in gain, bandwidth, stability, and noise figure." Summary of American Physical Society paper.

621.394/.397].645 3557

Cathode Follower Coupling in D.C. Amplifiers—Y. P. Yu. (*Electronics*, vol. 19, pp. 99–103; August, 1946.) The use of phase inverter circuits employing a high-value common cathode resistor is discussed, and the reduction of circuit noise considered. A high-gain (60 decibels) amplifier was designed for a single 250-volt power supply, using a cathode follower as an interstage coupling element, and with injection on to the screen grid: the circuit noise level was "almost unmeasurably low."

621.394/.397].645.34 3558

Termination Effects in Feedback Amplifier Chains—A. J. Ferguson. (*Canad. Jour. Res.*, vol. 24, sec. A, pp. 56–78; July, 1946.) An analysis is carried out of a typical intermediate-frequency or video-feedback-amplifier chain to investigate the influence of the input and output circuits on the over-all frequency response. The analysis is similar to that of a four-terminal passive network and the solution to the network equations can be resolved into a wave advancing from the input to the output and another reflected back through the amplifier to the input. Performance can thus be expressed in terms of the reflection coefficients of the source and terminal impedances. Reflection coefficients are calculated for typical simple circuits used with amplifiers of three, four, and five stages, and many of the derived frequency response curves are unsatisfactory. Considerably better curves may be obtained by using rather more complex terminations, and the best curves result from matching the amplifier chain to both the source and the load. Matching, however, limits the over-all gain to half the value which could otherwise be achieved.

621.394.396].645 3559

The Multiamp—C. E. Jackson. (*Radio News*, vol. 36, pp. 58, 155; September, 1946.) An amplifier giving 30 watts high-fidelity output over the entire audio range, with a total harmonic distortion of less then 2 per cent, and noise level 60 decibels below 30 watts on the microphone channels. Higher outputs, up to 90 watts, can be obtained by adding compact booster units using 6V6 tubes; the power gain of the tubes is increased by applying abnormally high anode and screen voltages and controlling the anode current by a simple network. The output transformer is provided with tappings giving fifteen possible impedance values for matching purposes. The input channels include two 125-decibel gain microphone stages and two 85-decibel gain program stages.

621.395.667 3560

Design of Attenuation Equalizers—H. N. Wroe. (*Wireless Eng.*, vol. 23, pp. 272–280; October, 1946.) From theoretical considerations a method of accurately calculating the performance of an equalizer is deduced. Its application in particular to telephone cables is illustrated. Equalizer performance is found to depend on two quantities which can be calculated without making conventional assumptions concerning its internal structure. A general method of equalizer design is deduced.

621.396.611 3561

The Tapped Inductor Circuit—J. E. Haworth. (*Electronic Eng.*, vol. 18, pp. 284–286; September, 1946.) The tapped inductor, in conjunction with a suitable capacitor, is considered as a two-terminal network possessing both resonant and anti-resonant properties. Expressions for impedance are derived and results illustrated graphically. The application of the theory in the design of tube oscillators is discussed.

621.396.621.53.018.41.012.7 3562

Mixer Frequency Charts—Badessa. (*See* 4037.)

621.396.645.014.332 3563

Peak Pulse Currents in Class B Amplifiers—K. R. Sturley. (*Wireless Eng.*, vol. 23, p. 286; October, 1946.) A simpler method than that involving Fourier analysis of the waveform, and only slightly less accurate, for estimating the peak pulse anode current taken by a valve in Class B operation, including the effect of the "standing feed."

GENERAL PHYSICS

531.18:531.15 3564

Is Rotation Relative or Absolute?—G. W. O. H. (*Wireless Eng.*, vol. 23, pp. 263–264; October, 1946.)

534.2:550.34 3565

Studies on Seismic Waves: Part 1—Reflection and Refraction of Plane Waves. Part 2—Rayleigh Waves in a Superficial Layer—C. Y. Fu. (*Geophys.*, vol. 11, pp. 1–23; January, 1946.)

535.215.1+621.383.2 3566

Complex Photocathodes—Khlebnikoff. (*See* 4080.)

535.247.4 3567

The Specification of a Spectral Correction Filter for Photometry with Emission Photocells—J. S. Preston. (*Jour. Sci. Instr.*, vol. 23, pp. 211–216; September, 1946.)

535.312.2 3568

Reflection [of Light] in a Corner Formed by Three Plane Mirrors—J. L. Synge. (*Quart. Appl. Math.*, vol. 4, pp. 166–176; July, 1946.) Consideration of the case where the three planes are not mutually perpendicular.

536:621.3.012.8 3569

Electrical Solution of Thermal Problems—F. G. Willey. (*Electronics*, vol. 19, pp. 190, 198; August, 1946.) By setting up an analogous electrical circuit, difficult thermal problems can sometimes be rapidly solved. Typical examples are given.

536.21:517.942.9 3570

Heat Conduction in Elliptical Cylinder and an Analogous Electromagnetic Problem—N. W. McLachlan. (*Phil. Mag.*, vol. 36, pp. 600–609; September, 1945.) The electromagnetic problem is that of a long uniform solenoid with a core of elliptical cross section in which a constant current I_0 is applied at $t=0$. The magnetizing force at any point in the core and the (variable) inductance of the solenoid due to the core flux are calculated.

537.13 3571

Production and Annihilation of Negative Protons—(*Nature*, (London), vol. 158, pp. 280–281; August 24, 1946.) A short account of a theoretical paper by J. McConnell published in *Proc. Roy. Irish Acad.*, discussing the possibility of observing negative protons which may be produced by collisions of mesons in cosmic rays.

537.2:621.317.72 3572

The Investigation of Electrostatic Fields by Means of a Valve Voltmeter—R. Street. (*Jour. Sci. Instr.*, vol. 23, pp. 203–204; September, 1946.) "The electrostatic field between two overlapping parallel planes has been investigated by means of a radio-active source and a valve electrometer. The equipotentials so obtained are in satisfactory agreement with the theoretical curves calculated from a Schwarz-Christoffel transformation. The experiment can be adapted as a laboratory exercise."

537.52:621.3.015.5.029.64 **3573**

Electrical Breakdown in Air at Microwave Frequencies—D. Q. Posin. (*Phys. Rev.*, vol. 69, p. 696; June 1–15, 1946.) "The Paschen Law was found to be approximately valid in this range, as in direct current, though dependent upon the length of time that the microwave field is applied." Summary of American Physical Society paper.

537.52:621.3.029.64 **3574**

Complex Conductivity of Electrical Discharge in Gas at Microwave Frequencies—M. A. Herlin and S. C. Brown. (*Phys. Rev.*, vol. 69, p. 696; June 1–15, 1946.) "The complex conductivity as a function of power gives the voltage and current characteristic of the discharge. Voltage and complex current characteristics have been studied as a function of cavity dimensions, tube dimensions, and pressure." Summary of American Physical Society paper.

537.521.6:533.5 **3575**

The Insulation of High Voltages in Vacuum—J. G. Trupp. (*Phys. Rev.*, vol. 69, p. 692; June 1–15, 1946.) Field emission theory is inadequate to account for insulation breakdown. Other mechanisms investigated include emission of particles by impact, and photoelectric emission. Summary of American Physical Socity paper.

537.525 **3576**

A Generalization of the Conception of Electron Plasma—A. A. Vlasoff. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 8, no. 5, pp. 248–266; 1944. In Russian.) The importance of the interaction between electrons over distances exceeding their average spacing ("distant forces") is emphasized. If this interaction is taken into account, new dynamic properties of polyatomic systems become apparent, and the conceptions of "gas," "liquid" and "solid body" are modified towards greater unification with the conception of plasma. The problem of the transition from "micro" to "macro" is also understood in a different light. The main result is the proof that if the fundamental kinetic equation (1) is used, in which the distant interaction is taken into account, the spontaneous appearance of the crystalline structure from the gas (under suitable conditions with regard to density and temperature) becomes evident without the use of any additional hypotheses. The same applies to the appearance of peculiar "vibrational" properties in polyatomic systems. Conditions necessary for the appearance of the crystalline structure are discussed in detail.

537.525 **3577**

The Behaviour of Electron-Ion Plasma in Magnetic Fields—G. V. Spivak and O. N. Repkova. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol 8, no. 5, pp. 275–279; 1944. In Russian.) The following effects of magnetic fields on the electron-ion plasma were investigated experimentally: (a) the variation in the concentration of charged particles in a magnetic field symmetrical with respect to the axis. Curves are plotted in Fig. 1 for various discharge currents (from 0.1 to 1.2 amperes) through mercury vapor in a "quasi-uniform" magnetic field, and in Fig. 2 for discharge currents of 100 and 400 milliamperes through mercury vapor in the field of a magnetic lens; (b) the variation in the average electron energy. A curve is plotted in Fig. 3; and (c) the variation in the electron distribution. The curve plotted in Fig. 4 refers to discharges through neon between two coaxial cylinders.

537.525:621.3.015.532 **3578**

The Variation of the Mobility of Negative Ions in Strong Electric Fields and the Rôle of This Phenomenon in Corona Discharges—N. A. Kaptsoff. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 8, no. 5, pp. 280–285; 1944. In Russian.) Experiments were conducted to explain the contradiction existing between the accepted values of K for the mobility of negative ions in humid air, and experimental values derived from volt-ampere characteristics of the corona discharge between two coaxial cylinders. The circuit used in these experiments is a simplified version of those proposed by Tyndall (1929 Abstracts, p. 146) and Van de Graaf (1928 Abstracts, p. 525) and the results obtained are plotted in Fig. 4. The contradiction is ascribed to the fact that the table values of K are given for values of E lower than those normally met in the case of corona discharges, and that complex ions are formed in humid air with these lower values of E. An abstract in English was noted in 1814 of July.

537.525.3:621.316.722.078.3 **3579**

Characteristics of the Pre-Corona Discharge and Its Use as a Reference Potential in Voltage Stabilizers—Brown. (*See* 3758.)

537.53 **3580**

Spontaneous Emission Probabilities at Radio Frequencies—E. M. Purcell. (*Phys. Rev.*, vol. 69, p. 681; June 1–15, 1946.) Summary of American Physical Society paper.

537.533:621.385.833 **3581**

Extraction of Electrons by an Electric Field—P. I. Lukirski. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 8, no. 5, pp. 226–231; 1944. In Russian.) A survey of literature on the Fowler-Nordheim theory of the electron emission from metals under the action of a strong electric field. Some of the experimental results obtained in Russia are also discussed. An abstract in English was noted in 1952 of July.

537.591.8 **3582**

Production of Mesons by Electrons—H. Feshbach. (*Phys. Rev.*, vol. 69, p. 690; June 1–15, 1946.) Summary of American Physical Society paper.

538.569.4+621.396.11].029.64:551.57 **3583**

Absorption of Microwaves by Water Vapor—Autler, Becker and Kellogg. (*See* 3719.)

538.569.4.029.64 **3584**

Absorption of Microwaves by Gases: Part 2—J. E. Walter and W. D. Hershberger. (*Phys. Rev.*, vol. 69, p. 694; June 1–15, 1946.) "The absorption coefficients and dielectric constants of sixteen gases have been measured at the two wavelengths $\lambda = 1.24$ centimeters and $\lambda = 3.18$ centimeters. The gases are H_2S, SO_2, COS, $(CH_3)_2O$, C_2H_4O, NH_3, six halogenated methanes, and three amines." For part 1 see 3234 of November. Summary of American Physical Society paper.

538.569.4.029.64:]534+536 **3585**

Thermal and Acoustic Effects Attending Absorption of Microwaves by Gases—W. D. Hershberger. (*Phys. Rev.*, vol. 69, p. 695; June 1–15, 1946.) "The conversion of microwave energy into thermal and acoustic energy by the use of any of the gases which absorb microwaves is described. . . . " "A gas-filled resonator capable of detecting 10 milliwatts of microwave power is described which consists of a cavity which resonates electrically to the microwave frequency and acoustically to the modulating frequency." Summary of American Physical Society paper.

539.16.081 **3586**

New Units for the Measurement of Radioactivity—E. U. Condon and L. F. Curtiss. (*Rev. Sci. Instr.*, vol. 17, p. 249; June, 1946.)

539.389.3:536.7 **3587**

Thermodynamics of Relaxation Processes—G. E. Kimball. (*Phys. Rev.*, vol. 69, p. 688; June 1–15, 1946.) Summary of American Physical Society paper.

541.13 **3588**

Physical Intersection of Electrons with Liquid Dielectric Media. The Properties of Metal-Ammonia Solutions—R. A. Ogg, Jr. (*Phys. Rev.*, vol. 69, pp. 668–669; June 1–15, 1946.) A discussion in terms of quantum mechanics. The electrical and magnetic properties of such solutions are explained.

621.3.011.2+621.3.011.5]:546.3[-13+-14 **3589**

On the Relationship between the Liquid and Gaseous States in Metals—Ya. Zel'dovich and L. Landau. (*Zh. Eksp. Teor. Fiz.*, vol. 14, pp. 32–34; 1944. In Russian.) It is pointed out that even at absolute zero there is a definite difference between a metal and a dielectric, and that a transition from one state into another can only be effected with emission or absorption of latent heat and an abrupt change in the properties of the material. A similar question of the transition from a metallic liquid state into a dielectric gaseous state is then examined, and the conclusion reached is that there should exist a nonmetallic liquid phase which changes abruptly into a metal, when pressure increases, or into a gas, when pressure decreases.

621.314.63:539.233 **3590**

Contact Potential Difference as a Tool in the Study of Adsorption—R. C. L. Bosworth. (*Jour. Roy. Soc. N.S.W. of 1945*, vol. 79, part 2, pp. 53–62; June 19, 1946.) A description of apparatus for using the contact potential difference method to measure the work function of a surface and an account of its application to the study of the properties of electro-positive and electro-negative films and to the measurement of vapor pressures.

621.314.632 **3591**

Theory of Crystal Rectifiers—Sachs. (*See* 3629.)

621.314.632.029.6 **3592**

High Frequency Rectification Efficiency of Radar Crystal Detectors—Lawson, Goodman, and Schiff. (*See* 3636.)

621.317.332:537.312.62 **3593**

High-Frequency Resistance of Superconductors—A. B. Pippard. (*Nature*, (London), vol. 158, pp. 234–235; August 17, 1946.) Measurements have been made in the temperature range 2 to 4 degrees Kelvin on mercury and polycrystalline tin using the specimen to form a twin, separately shielded quarter-wave transmission line. The measured specific conductivity for tin at 3.8 degrees Kelvin in the region of 1200 megacycles is about one seventh the direct-current value. These results, together with some obtained in the presence of a superposed constant magnetic field, are related to those of previous observers (See especially 208 of 1941).

621.384.2 **3594**

The Technique of Gamma Radiography—R. Halmshaw. (*Engineering*, (London), vol. 162, pp. 169–170; August 23, 1946.)

621.386.77 **3595**

Secondary Radiation from X-Ray Filters—N. M. Morrow. (*Canad. Jour. Res.*, vol. 24, sec. A, pp. 46–55; July, 1946.)

537(075) **3596**

Principles of Electricity Illustrated [Book Review]—R. C. Norris. Odhams Press, London, 380 pp., 8s. 6d. (*Electronic Eng.*, vol. 18, p. 293; September, 1946.)

538.1 3597

Le Magnétisme. Vol. 1: Généralités et Magnéto-Optique; Vol. 2: Ferromagnétisme; Vol. 3: Paramagnétisme [Book Notice]—Collection Scientifique, Institut International de Coopération Intellectuelle, Paris; Columbia Univ. Press, 1940, 184+280+348 pp., $2.00+$2.50+$2.50. (*Science*, vol. 104, pp. 70–73; July 26, 1946.) Report by S. J. Barnett on the proceedings of an international conference on magnetism held in Strasbourg in 1939 under the presidency of the Professor Weiss, the volumes constituting a wide survey of current magnetic research and summaries of earlier work.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16:621.396.822 3598

Interpretation of Cosmic Noise—Radio Waves from Extraterrestrial Sources—C. H. Townes. (*Phys. Rev.*, vol. 69, p. 695; June 1–15, 1946.) Assuming cosmic noise is due to radiation from interstellar ions of density n and at temperature T, the effective temperature of the Milky Way is found to depend on frequency ν as

$$Ta = T[1 - \exp(-8 \times 10^{-8} \frac{n^2 S}{\nu^2})]$$

where S is the distance from the observer to the Milky Way boundary in the direction of observation. This gives qualitative agreement with some of the observational data. Summary of American Physical Society paper.

523.165:621.396.822 3599

Fluctuations in Cosmic Radiation at Radio-Frequencies—J. S. Hey, S. J. Parsons, and J. W. Phillips. (*Nature*, (*London*), vol. 158, p. 234; August 17, 1946.) Continuation of experiments referred to in 3270 of November. "Short period [order of several seconds] irregular fluctuations have been found to be associated with the direction of Cygnus. . . . The average amplitude of the fluctuations is 15 per cent of the mean power received." The source appears to subtend an angle not exceeding 2 degrees. A brief discussion of the significance of the results is given.

537.591.5 3600

Production of Mesotrone in the Stratosphere—I. Bloch. (*Phys. Rev.*, vol. 69, pp. 575–585; June 1–15, 1946.) Theoretical derivation of the intensity of the penetrating cosmic-ray particles as a function of altitude, magnetic latitude and energy, assuming mesotrons to be produced in multiples by primary protons in fields of atomic nuclei.

537.591.8 3601

The Ionization Spectrum of Cosmic-Ray Electrons and Mesotrons—P. B. Weisz and W. F. G. Swann. (*Phys. Rev.*, vol. 69, p. 690; June 1–15, 1946.) The average ionization produced by mesotrons was found experimentally to be about 68 per cent of that produced by electrons, but the experimental spread in values is unexpectedly large. Summary of American Physical Society paper.

538.71:629.123.011.22 3602

Measurement of Magnetic Fields Beneath Ships—H. A. Miller. (*Jour. R. Soc. Arts*, vol. 94, pp. 327–329; April 12, 1946.) A comparison of the magnetic field existing beneath a ship in England and at a measurement range constructed off Colombo Harbour in the region of the magnetic equator.

550.34:534.2 3603

Studies on Seismic Waves: Part 1—Reflection and Refraction of Plane Waves. Part 2—Rayleigh Waves in a Superficial Layer—Fu. (See 3565.)

551.51.053.5:621.396.11 3604

Short-Wave [Ionospheric] Forecasting—Bennington. (*See* 3721.)

LOCATION AND AIDS TO NAVIGATION

534.417:534.88 3605

Echo Ranging Sonar [Model QCS/1T]—R. J. Evans. (*Electronics*, vol. 19, pp. 88–93; August, 1946.) A device for accurate location of underwater targets. A principle analogous to radar is employed, with pulses of sound energy: echoes are received from obstacles. Frequencies employed are in the range 10 kilocycles to 30 kilocycles. A 600-watt pulse of 0.1 to 0.2 seconds duration is transmitted at intervals of several seconds by a self-excited oscillator and amplifier feeding a magnetostriction projector-receiver, housed in a dome under the ship. Receiving equipment comprises a superheterodyne receiver, a range indicator and a loudspeaker for providing an audible signal. Each of these units is described in detail with block and circuit diagrams.

621.396.7:621.396.9 3606

Decca Navigator Stations—M. G. Scroggie. (*Wireless World*, vol. 52, pp. 260–262; August, 1946.) The essential requirement of the service is the maintenance over the whole of its area of a phase pattern that is stationary and permanent. In order that the two interfering wave trains may be separately received, for purposes of phase checking, the transmissions are on two exact submultiples of the phase comparison frequency; one of a pair (or more) of transmitting stations provides the master drive for the others, thus ensuring relative frequency constancy, while a receiver automatically corrects the phase of one of each pair of transmissions. See also 1242 of May and back reference.

621.396.9 3607

Factors Affecting the Range of Radar Sets—L. R. Quarles and W. M. Breazeale. (*Trans. A.I.E.E.* (*Elec. Eng.* August–September, 1946), vol. 65, pp. 546–548; August–September, 1946.) The relation between received and transmitted powers is derived for the case of transmitting and receiving aerials fitted with parabolic reflectors, in terms of the wavelength, power gain of the aerials, and the range and effective area of the target. By considering the magnitude of the thermal agitation noise in the receiver an expression is obtained for the maximum range at which an echo is detectable. The formula is discussed and illustrated by a numerical example. A brief discussion is given of the nature of the polar diagram of aerials using parabolic reflectors.

621.396.9 3608

Navigating by Loran—C. J. Pannill. (*Telegr. Teleph. Age*, vol. 64, pp. 17–18; September, 1946.) A brief account of experiences on the liner Drottningholm during a voyage from New York to Shanghai. Dependable ranges of 700 miles by day and 1400 miles by night were obtained.

621.396.9 3609

Successor to the Sextant—H. Manchester. (*Sci. Amer.*, vol. 174, pp. 264–267; June, 1946.) An elementary account of loran.

621.396.9 3610

Radar—E. Aisberg. (*Cah. Toute la Radio*, pp. 5–12; September, 1945.)

621.396.9:621.396.932 3611

Metrovick Marine Radar Set—Metropolitan-Vickers Electrical Co. (*Engineering*, (London), vol. 162, pp. 79–80; July 26, 1946; *Wireless World*, vol. 52, p. 269; August, 1946.) A production model, type MR1, which consists of three units: (1) a console containing the receiver, plan-position indicator, timebase strobe unit, control panel, power unit, and modulator; (2) the transmitter proper, signal mixer, head amplifier and mixer, intermediate-frequency amplifier, and discriminator circuit for automatic-frequency control; (3) the aerial unit of parabolic "cheese" type, normally driven at 20 rotations per minute and mounted in a perspex dome.

The frequency is 9500 megacycles with a peak power of 50 kilowatts, ½-microsecond pulses are used with a repetition rate of 1000 per second. Three ranges are provided: 3000 yards, 10,000 yards, and 60,000 yards; the plan-position indicator displays obstructions as near as 50 yards, and an audible warning signal is incorporated.

621.396.9:621.396.932 3612

Navigational Radar in Merchant Ships—E. H. Hart. (*Electronic Eng.*, vol. 18, pp. 265–267; September, 1946.) A detailed description of modified naval radar equipment, demonstrated on H. M. S. Fleetwood. See also 1246 and 1247 of May.

621.396.9:621.397 3613

Has Radar Influenced Television Development?—(*See* 3799.)

621.396.9:623.454.25 3614

Radio Proximity Fuze Design—Hinman and Brunetti. (*See* 3713.)

621.396.932.25 3615

Auxiliary Pilot Guides Ships—J. H. Jupe. (*Electronics*, vol. 19, pp. 154, 160; August, 1946.) A buoyant tank containing radar receiving and transmitting equipment is placed in position by a skilled navigator. The circuit consists basically of a supersonically quenched radio-frequency amplifier, (made to oscillate for transmitting) with a demodulator and pulse amplified. A pulse signal from a ship-borne interrogator is received and returned greatly amplified. On immersion, power supplies are connected automatically and the buoy is armed for self-destruction.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 3616

A Sensitive Vacuum Gauge with Linear Response—J. R. Downing and G. Mellen. (*Rev. Sci. Instr.*, vol. 17, pp. 218–223; June, 1946.) Ionization gauge for measuring pressures in the range between zero and 10 millimeters of mercury, with linear response.

531.788.7 3617

A Radium Source Ion Gauge—G. L. Mellen. (*Phys. Rev.*, vol. 69, p. 691; June 1–15, 1946.) For measurement of pressures between 1 micron and 10 millimeters. Summary of American Physical Society paper.

533.5:621.3.032.53 3618

Glass-to-Metal Seal Design—W. J. Scott. (*Jour. Sci. Instr.*, vol. 23, pp. 193–202; September, 1946.) The physics and chemistry of "oxide" and other seals are discussed from a theoretical and practical standpoint. Particular attention is paid to the formation of cracks and a nomogram correlates stress, strain and glass thickness. The relevant physical properties of various glasses and metals are tabulated. Excerpts from this paper were noted in 97 of January.

549.514.51:534.321.9 3619

Refinements in Supersonic Reflectoscopy. Polarized Sound—Firestone and Frederick. (*See* 3515.)

535.215.1:546.28 3620

The Velocity of Propagation of the Transmitted Photo-Effect in Silicon Crystals—F. C. Brown. (*Phys. Rev.*, vol. 69, p. 686; June 1–15, 1946.) The relative magnitude of the transmitted effect is a function of the velocity, coefficient of recapture and distance travelled. The velocity, measured directly for distances of travel from 0.4 to 1 centimeter is 400 meters per second ±5 per cent. Summary of American Physical Society paper.

535.37 3621

Corpuscular vs. Undulatory Excitation of Phosphors—H. W. Leverenz. (*Phys. Rev.*, vol. 69, p. 686; June 1–15, 1946.) Summary of American Physical Society paper.

535.37:535.61-15 3622

Some Properties of Infra-Red Sensitive Phosphors—R. T. Ellickson. (*Phys. Rev.*, vol. 69, pp. 685–686; June 1–15, 1946.) Alkaline earth sulphides and selenide phosphors activated with rare earths show a marked increase in phosphorescence under infra-red stimulation. Summary of American Physical Society paper.

537.226.8:621.319.4 3623

Temperature Coefficients of Interfacial Polarization in Dielectrics—R. F. Field. (*Phys. Rev.*, vol. 69, p. 688; June 1–15, 1946.) Summary of American Physical Society paper.

541.147.4:546.683.22 3624

The Photoelectric Mechanism of the Thallous Sulfide Photo-Conductive Cell—A. von Hippel, F. G. Chesley, H. S. Denmark, P. B. Ulin, and E. S. Rittner. (*Phys. Rev.*, vol. 69, p. 685; June 1–15, 1946.) Thermoelectric measurements show that pure thallous sulphide is "an 'excess,' semiconductor changing to a 'defect' conductor on oxidation." The photosensitization of thallous sulphide is due to the presence of oxygen. Summary of American Physical Society paper.

546.431.826:621.3.011.5 3625

Oscillographic Study of the Dielectric Properties of Barium Titanate—A. de Bretteville, Jr. (*Phys. Rev.*, vol. 69, p. 687; June, 1946.) Barium titanate and solid solutions of barium and strontium titanates exhibit saturation of the dielectric flux density with increasing field strength. These properties have been studied by an oscillograph method throughout the charging and discharging cycle. Summary of American Physical Society paper.

546.45:669.018 3626

Beryllium: Workaday Metal—F. P. Peters. (*Sci. Amer.*, vol. 174, pp. 249–251; June, 1946.) A nontechnical account of its principal properties and uses as an alloying metal.

549.514.51 3627

Recrystallization of Quartz as a Result of Flexure—D. D. Eustachio. (*Phys. Rev.*, vol. 69, p. 687; June 1–15, 1946.) Quartz crystals change from single crystals to a "poly-crystalline" structure when they are thinner than 25 microns but after cold working appear to recrystallize. See also 3310 and 3311 of November. Summary of American Physical Society paper.

621.314.632 3628

Rectification Series—W. H. Brattain. (*Phys. Rev.*, vol. 69, p. 682; June 1–15, 1946.) Rectification takes place between one crystal and another on which a point has been made. Using the pointed crystal as reference, a series may be made such that any crystal in it will rectify in one direction with those above it and in the other direction with those below it. Summary of American Physical Society paper.

621.314.632 3629

Theory of Crystal Rectifiers—R. G. Sachs. (*Phys. Rev.*, vol. 69, p. 682; June 1–15, 1946.) The observed direct-current characteristics of Ge and Si rectifiers agree approximately with those calculated on the multicontact theory which assumes that (1) the contact potential ϕ varies continuously over the surface of contact, (2) the total current is the sum of the partial currents flowing through regions of varying ϕ, (3) the area of a region with a contact potential ϕ may be a function of ϕ, and (4) the number of spots having contact potential between ϕ and $\phi+d\phi$ is a function of ϕ. Summary of American Physical Society paper. See also 3630 and 3633.

621.314.632 3630

Semi-Quantitative Explanation of D. C. Characteristics of Crystal Rectifiers—V. A. Johnson, R. N. Smith, and H. J. Yearian. (*Phys. Rev.*, vol. 69, pp. 682–683; June 1–15, 1946.) "The multicontact theory is employed to explain the observed direct-current versus voltage characteristics between metal and semiconductor (Ge and Si). . . . The proper slopes are obtained for the forward current." A graphical method is developed for the rapid synthesis of any characteristic. Summary of American Physical Society paper. See also 3629 and 3633.

621.314.632 3631

Contact Capacity of Crystal Rectifiers—R. N. Smith. (*Phys. Rev.*, vol. 69, p. 683; June 1–15, 1946.) The capacitance is measured indirectly from the decrease of rectification efficiency at microwave frequencies assuming that the crystal can be represented by a "spreading" resistance in series with a parallel combination of capacitance and the nonlinear contact resistance, these elements being independent of frequency. The dependence of capacitance on direct-current bias is in agreement with theory for silicon crystals but not for germanium units. Summary of American Physical Society paper.

621.314.632 3632

Image Force and Tunnel Effect in Crystal Rectifiers—E. D. Courant. (*Phys. Rev.*, vol. 69, p. 684; June 1–15, 1946.) The divergence of the experimental i—V curve from the theoretical (equation given) is explained by assuming that the height of the potential barrier is reduced by the image force, and that some electrons tunnel through the barrier. Summary of American Physical Society paper.

621.314.532:546[.28+.289 3633

D. C. Characteristics of Ge and Si Crystal Rectifiers—H. J. Yearian. (*Phys. Rev.*, vol. 69, p. 682; June 1–15, 1946.) The slope of the i—V characteristic is less than that predicted. A consideration of the difficulties in explaining the discrepancy. Summary of American Physical Society paper. (See also 3629 and 3630.)

621.314.632:546.289 3634

High Voltage and Photo-Sensitive Characteristics in Germanium—S. Benzer. (*Phys. Rev.*, vol. 69, p. 683; June 1–15, 1946.) Photo-effects of two types are observed in the visible and near infra-red range: (1) a saturated i—V characteristic in which the saturation current varies with illumination and temperature, and (2) a triple-valued characteristic with a voltage peak and negative resistance region. Some Ge rectifiers can withstand high inverse voltages. Summary of American Physical Society paper.

621.314.632:546.289 3635

Effect of Various Atmospheres on Germanium Crystal Rectifiers—R. M. Whaley and K. Lark-Horovitz. (*Phys. Rev.*, vol. 69, pp. 683–684; June 1–15, 1946.) Relatively high-purity germanium which provided poor rectification in a vacuum was unchanged by admission of gas, samples of relatively high conductivity, due to impurity content, which gave good rectification in vacuum, showed irreversible increases in back resistance upon admitting air. The changes can be accounted for by multicontact theory. Summary of American Physical Society paper.

621.314.632.029.6 3636

High Frequency Rectification Efficiency of Radar Crystal Detectors—A. W. Lawson, B. Goodman, and L. I. Schiff. (*Phys. Rev.*, vol. 69, p. 682; June 1–15, 1946.) If a finite time is required to ionize the semiconductor donator levels in the blocking layer, the rate of decrease of efficiency with frequency may be explained. Summary of American Physical Society paper.

621.314.632.029.6 3637

Noise in Radar Crystal Detectors—Schiff. (*See* 3725.)

621.315.59 3638

The Effect of Grain Structure on the Electrical Conductivity of Semiconductors—B. Goodman. (*Phys. Rev.*, vol. 69, p. 687; June 1–15, 1946.) Only the effect due to the discontinuities introduced into the periodic lattice potential is considered. The added resistivity may sometimes be comparable with that part of the room temperature resistance due to lattice vibrations, but it is small compared with the total resistivity which is mostly caused by the impurity ions. Summary of American Physical Society paper.

621.315.61:546.4 3639

Materials with High and Super-High Permittivities—B. M. Vul. (*Elektrichestvo*, no. 3, pp. 12–17; 1946. In Russian.) Experimental investigation of titanates of metals in the second group of Mendeleeff's table. It is shown that a barium titanate can be used as an electrical insulating material with a permittivity exceeding 1000.

621.315.612:621.319.4.029.5 3640

High Frequency Ceramic Capacitors—Vul and Skanavi. (*See* 3779.)

621.315.616 3641

Synthetic Rubbers and Plastics: XI. (Part 3) Water and the High Polymer—F. T. White. (*Distrib. Elec.*, vol. 19, pp. 170–173; October, 1946.) A discussion on moisture permeability, and the effect of including plasticisers, resins, or waxes in the basic materials. For part 2 (Section XI) see 2936 of October.

621.315.616:621.38/.39 3642

Plastics in the Electronic and High Frequency Industries—W. S. Penn. (*Electronic Eng.*, vol. 18, pp. 280–281; September, 1946.) Improvements and future possibilities. Tables give properties of dielectrics of various types.

621.315.616.9:621.3 3643

Plastics and the Electrical Industry: Parts 1-4—P. R. S. Gibson. (*Electrician*, vol. 137, pp. 443–445, 517–520, 582–585, and 649–652; August 16, and September 6, 1946.) In part 1 the conductivity, dielectric strength, permittivity, and power factor of plastic insulating materials in general are defined and discussed; in subsequent parts the properties and uses of specific materials are described.

621.318.22:620.179.14 3644

Magnetic Testing of [Ferromagnetic] Metals—P. E. Cavanagh, E. R. Mann, and R. T. Cavanagh. (*Electronics*, vol. 19, pp. 114–121; August, 1946.) The examination of the metallurgical properties of metals by testing their magnetic qualities. An extensive description of the technique with practical details and applications is given.

621.383.4+535.215.1:546.28 3645

A New Bridge Photo-Cell Employing a Photoconductive Effect in Silicon. Some Properties of High Purity Silicon—Teal, Fisher, and Treptow. (*See* 3784.)

621.396.611.21.032.2:546.59 3646

Gold Film Electrodes for High Frequency Quartz Plates—Spears. (*See* 3794.)

669.45.778:621.315.22 3647

F-3 Lead Alloy—an Improved Cable Sheathing—L. F. Hickernell and C. J. Snyder. (*Trans. A.I.E.E.* (*Elec. Eng.* August–September, 1946) vol. 65, pp. 563–569; August–September, 1946.) Details of a new arsenical lead alloy for cable sheathing, including chemical composition and physical properties. It is superior to materials hitherto used for power cables, notably in respect of its increased resistance to bending fatigue, creep resistance, and bursting strength.

621.315.614.72 3648

Varnished Cloths for Electrical Insulation [Book Review]—H. W. Chatfield, and J. H. Wreddon. J. and A. Churchill, London, 1946, 255 pp., 21s. (*Electronic Eng.*, vol. 18, p. 292; September, 1946.) A "clear and orderly exposition of the technical features of varnished fabrics. . . . The properties of the finished products . . . are very fully dealt with."

621.315.616 3649

Collection "Matériaux de Synthese." Aminoplastes [Book Review]—P. Talet. **Résines Vinyliques [Book Review]**—H. Gibello. Dunod, Paris, 280 fr. and 260 fr. (*Engineering*, (London), vol. 162, p. 171; August 23, 1946.)

MATHEMATICS

517.432.1 3650

Some Notes on the Operational Calculus—L. Jofeh. (*Jour. Brit. Instn. Radio Eng.*, vol. 6, pp. 73–77; March–May, 1946. Discussion, pp. 77–79.) A short review is given of the various systems of operational calculus and the Heaviside system is discussed in greater detail. The basis of the operational form of a differential equation is explained and the interpretation of the operational equation illustrated by an example. A bibliography of 14 items is given.

517.512.2 3651

Fourier Series—M. M. Levy. (*Jour. Brit. Instn. Radio Eng.*, vol. 6, pp. 64–73; March–May, 1946.) An outline of Fourier analysis and its application to the study of periodic and nonperiodic functions. Fourier series are given in trigonometrical and exponential form and simplified forms derived which are applicable to periodic curves having various types of symmetry. Expressions are given for denoting the coefficients of the series from the discontinuities in a periodic curve. The theory is extended to the treatment of nonperiodic functions such as the unit pulse and unit impulse functions, and the response of a low-pass filter to a pulse of given width is determined. Brief mention is made of the harmonic analysis of periodic functions with special reference to methods of determining very high harmonic components.

517.942.82:621.3.014/.015].33 3652

Switching Problems and Instantaneous Impulses—Jaeger. (*See* 3550.)

518.2 3653

Mathematical Tables—(*Jour. Res. Nat. Bur. Stand.*, vol. 37, p. 73; July, 1946.) A list of mathematical tables made available through the National Bureau of Standards, including powers of integers, exponentials, circular functions and associated integrals, logarithms, and probability functions.

518.4:676.31 3654

Utility of Log-Log vs. Arithmetic Co-Ordinate Grids—A. A. Merrill. (*Gen. Elec. Rev.*, vol. 49, pp. 20–22; July, 1946.)

518.5:621.385.001.8 3655

Electronic Computers—W. Shannon. (*Electronics*, vol. 19, pp. 110–113; August, 1946.) The fundamental circuits for performing elementary mathematical calculations electrically are described, with brief remarks on the military and industrial use of electrical computers.

534+537].014.2:518.61 3656

On the Numerical Treatment of Forced Oscillations—A. C. Sugar. (*Quart. Appl. Math.*, vol. 4, pp. 193–196; July, 1946.) An approximation to the Duhamel integral is used as the solution of the differential equation of a harmonic oscillator. Vector methods are used to deduce the maximum displacement and acceleration.

536.21:517.942.9 3657

Heat Conduction in Elliptical Cylinder and an Analogous Electromagnetic Problem—McLachlan. (*See* 3570.)

621.396.679.4.012.2 3658

Solving Feeder Problems Graphically—Kelley. (*See* 3549.)

516+517 3659

Analytical Geometry and Calculus [Book Review]—H. B. Phillips. J. Wiley, New York: Chapman and Hall, London, second edition, 504 pp., 27s. (*Jour. Sci. Instr.*, vol. 23, p. 218; September, 1946.) For courses in science and engineering.

518.2 3660

An Index of Mathematical Tables [Book Review]—A. Fletcher, J. C. P. Miller and L. Rosenhead. Sci. Computing Service, London, 1946, 451 pp., 75 s. (*Nature*, (London), vol. 158, pp. 252–253; August 24, 1946.) See also 3339 of November.

MEASUREMENTS AND TEST GEAR

531.71+.76]:621.383 3661

A Photoelectric Method of Indicating Small Displacements and of Timing a Moving Body—D. S. Perfect and R. M. J. Withers. (*Jour. Sci. Instr.*, vol. 23, pp. 204–207; September, 1946.)

531.76 3662

The Measurement of Ultra-Short Time Differences—S. H. Neddermeyer. (*Phys. Rev.*, vol. 69, p. 702; June 1–15, 1946.) A device ("chronotron") which uses a bent coaxial line to produce superposition of pulses whose time interval is being measured. Accuracy to $\pm 3.10^{-11}$ seconds is claimed. Summary of American Physical Society paper.

531.76:621.317.755 3663

A Note on the Measurement of Pulse Duration by Anode-Current Form-Factor—L. H. Ford. (*Jour. Sci. Instr.*, vol. 23, p. 216; September, 1946.) Pulse shape is intermediate between a rectangle and a triangle. In the former case pulse duration t is given by $t = T/F^2$, and in the latter by $t = 1.33T/F^2$ where T is the pulse-recurrence time and F the ratio of root-mean-square (measured thermally) to mean value of oscillator-anode current supply. The assumption of a perfect rectangle will give durations too small by not more than 10 per cent.

In an experimental verification the results were compared with those obtained using a cathode-ray oscilloscope over a range of pulse width 2 to 1000 microseconds and pulse recurrence 50 to 5000 microseconds. Agreement was within 5 per cent.

538.214:621.317.44 3664

An Electrodynamic Balance for the Measurement of Magnetic Susceptibilities—T. S. Hutchison and J. Reekie. (*Jour. Sci. Instr.*, vol. 23, pp. 209–211; September, 1946.) The magnetic forces are compensated by passing known currents through a coil system. The balance is directly calibrated and can measure forces as low as a few hundredths of a dyne to an accuracy of about ½ per cent.

539.16.08:537.591.8 3665

Fluctuations in Measurements of Ionization per Centimeter of Path in Proportional Counters—W. F. G. Swann. (*Phys. Rev.*, vol. 69, p. 690; June 1–15, 1946.) Summary of American Physical Society paper.

621.317.33 3666

Measuring Mutual Inductance and Capacitance—A. W. Simon. (*Electronics*, vol. 19, pp. 142, 154; August, 1946.) The method used "is based on the Campbell modification of the Felici circuit. . . . It is particularly suitable for the accurate measurement of mutual inductance between coils with small coefficients of coupling such as are frequently encountered in radio work. . . ." Measurements of intercoil capacitances can also be made.

621.317.333 3667

Measuring Insulation Resistance—J. Piggott. (*Wireless World*, vol. 52, pp. 263–264; August, 1946.) The model 40 and model 7 "Avometers" are fitted with an automatic cut-out which requires the use of a protective rectifier. The circuit thus formed with the inductance of the moving coil of the instrument resonates at about 450 kilocycles. If this frequency is picked up it may be rectified and produce false readings.

621.317.333.027.3 3668

High Voltage D.C. Testing of Rubber-Insulated Wire—W. N. Eddy and W. D. Fenn. (*Trans. A.I.E.E.* (*Elec. Eng.*) August–September, 1946), vol. 65, pp. 576–578; August–September, 1946.) The results of experiments on the limits suitable for the high voltage direct-current testing of plain insulated wire immersed in water.

621.317.7.017.5 3669

Instrument Bearing Friction—A. L. Nylander. (*Gen. elec. Rev.*, vol. 49, pp. 12–17; July, 1946.) Formulas are given for computing the bearing-system frictional torque in instruments requiring high sensitivity (e.g. microammeters). To reduce this torque, minimum pivot radius, maximum jewel-radius and minimum end-play consistent with small side-play error are required.

621.317.715.085.39 3670

A Simple Galvanometer Amplifier with Negative Feedback—J. S. Preston. (*Jour. Sci. Instr.*, vol. 23, pp. 173–176; August, 1946.) An intense beam of light reflected from the mirror of an ordinary d'Arsonval galvanometer is focused on a pair of selenium rectifier photocells connected in series opposition, the net current from which is measured by a secondary galvanometer. Negative feedback is introduced by interconnection of the two galvanometer circuits to give a more linear over-all response, and a sensitivity less influenced by changes in individual elements. A red-absorbing filter is used in the light beam to eliminate photocell fatigue. Sufficient gain is obtained without tube amplification to give an over-all sensitivity of 15 to 29 meters per microampere with zero repetition to 10^{-10} ampere.

537.2:621.317.72 3671

The Investigation of Electrostatic Fields by Means of a Valve Voltmeter—Street. (*See* 3572.)

621.317.72:621.384.6 3672

A Megavoltmeter for Induction Electron Accelerators—W. F. Westendorp. (*Rev. Sci. Instr.*, vol. 17, pp. 215–217; June, 1946.)

621.317.725:621.385 3673

Inverse Vacuum-Tube Voltmeter—S. H. Dike. (*Electronics*, vol. 19, pp. 140, 142; August, 1946.) Specially designed for measuring high voltages and maintaining a high input impedance. Calibration curves and a circuit diagram are given.

621.317.725:621.385 3674

Valve Voltmeter—F. Haas. (*Cah. Toute la Radio*, pp. 18–20; September, 1945.)

621.317.726 3675

Fighting Vehicle Exhibition—(*Engineer*, (London), vol. 182, pp. 148–149; August 16, 1946.) Includes description of a battery-operated peak voltmeter designed for the measurement of sparking-plug voltages. It is basically a resistance potential divider followed by a triode rectifier with a milliammeter in the anode calibrated directly in kilovolts.

621.317.738 3676

Increasing the Sensitivity of the Schering Bridge for the Measurement of Small Loss Angles at Low Voltage—G. Sella. (*Alta Frequenza*, vol. 15, pp. 15–27; March, 1946. With English, French and German summaries.) A push-pull preamplifier is inserted between the bridge and the output transformer in such a way as to preserve symmetry to earth; balance adjustments are provided on one of the values by a variable cathode resistor and a variable capacitance from anode to earth. The effective amplification at 50 cycles is roughly unity and the asymmetry is about .5. 10^{-6}. The preamplifier is followed by a two-stage amplifier feeding a bridge-type copper-oxide rectifier connected to a direct-current amplifier. Tests show that at 50 cycles loss angles of 10^{-6} can be measured to 5 per cent even with a test capacitance of 50 micromicrofarads.

621.317.755 3677

Elements of a New Oscilloscope Design—E. C. Simmons. (*Elec. Ind.*, vol. 5, pp. 96–97, 118; September, 1946.) Design features of a commercial oscillograph for general laboratory use. A variety of timebases and synchronizing arrangements are considered. A calibration circuit for the determination of input signal amplitude is also included, together with a pulse generator for triggering external circuits.

621.317.761.029.62/.64 3678

The Determination of Very High Frequencies—F. Dickson. (*Proc. I.R.E.* (Australia), vol. 7, p. 20; July, 1946.) "Fundamentally, the method is to determine two or more relatively low frequencies, the harmonics of which are related to the frequency to be checked in such a way that the latter is the l. centimeter of the observed frequencies," i.e. $F=nf_1=(n-1)f_1$ etc. For a suitable instrument see 679 of March.

621.317.763.029.62/.63 3679

Frequency Measurements at U.H.F.—R. Endall. (*Radio News*, vol. 36, pp. 50–52, 100; September, 1946.) Description of types of wavemeter available for amateur use in the range 100 to 3000 megacycles, including absorption wavemeters, butterfly types, Lecher wire systems, and heterodyne frequency meters.

621.317.763.029.62 3680

A Wavemeter for the Ultra-Short Band—J. Banner. (*Electronic Eng.*, vol. 18, pp. 268–269; September, 1946.) A compact, robust instrument, with frequency coverage 155 to 255 megacycles and accurate to within ±0.25 megacycle. Provision is made for use as a signal generator.

621.317.784.029.4 3681

Hypso-Wattmeter [Variable-Impedance Output Meter]—C. M. Laurent. (*Cah. Toute la Radio*, pp. 9–11; October, 1945.) Compares the power outputs of amplifiers or receivers up to 20 watts with a reference power of 1 milliwatt, at frequencies between 30 and 20,000 cycles. Design details are given.

621.317.79:621.396.619 3682

The Time-Delay [Amplitude] Modulation Meter. TH 3077—M. Sollima. (*Rev. Tech. Comp. Franç. Thomson-Houston*, pp. 45–58; January, 1944.) A brief discussion of various types of meter for measuring modulation depth; the advantages of incorporating a time delay (described in French Patent 858,680) to give a suitable measure of peak values with small error are stated. A description of the TH 3077 includes details of circuit design. The low-frequency amplifier is provided with negative feedback which does not function until the input level exceeds a certain value. The peak voltmeter is followed by a logarithmic direct-current network using three biased diodes enabling a nearly linear decibel scale to be obtained over range of 50 decibels. The error of measurement is less than 1 decibel for frequencies of 30 to 15,000 cycles.

621.385.832.088 3683

C.R. Tube Quality Measuring Apparatus—A. H. Spooner. (*Electronic Eng.*, vol. 18, pp. 273–276, September, 1946; *Jour. Telev. Soc.*, vol. 4, pp. 251–254; June, 1946.) Circuits are described for the determination of distortion in either magnetically or electrostatically deflected television tubes, and in their associated timebases. Pulses, synchronized to the line and frame frequencies are fed to the picture grid to produce on the screen an array of dots. The defocusing and symmetrical spacing of the dots give an indication of faulty circuits. Variation of spot size with peak beam current may also be determined.

621.396.67.08.011.2 3684

Antenna Impedance Measurement—D. D. King. (*Phys. Rev.*, vol. 69, p. 696; June 1–15, 1946.) Summary of American Physical Society paper.

621.317.029.3/.6 3685

Alternating Current Measurements at Audio and Radio Frequencies [Book Review]—D. Owen. Methuen, London, second edition, 120 pp., 5s. (*Jour. Sci. Instr.*, vol. 23, p. 218; September, 1946.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.714.7:621.317.39 3686

Electronic Gaging—P. H. Hunter. (*Elec. Ind.*, vol. 5, pp. 68–71; September, 1946.) Describes two micrometers, one using a tube circuit for indicating the moment of contact (between the micrometer screw and the material measured) and the other using a photoelectric technique for continuous gauging of soft materials without physical contact.

535.61-15:536.51.072.2 3687

Thermal Detectors—(*Elec. Ind.*, vol. 5, pp. 87, 118; September, 1946.) Survey of the various principles used, especially superconductivity, and description of two bolometers, one of which can detect a temperature change as small as 10^{-6} degrees centigrade, and has a time constant of the order of 1 millisecond. Sensitive apparatus of this type can be used for measurements of stellar radiation, atmospheric humidity, small radio-frequency currents, for physiological studies and in navigation.

535.61-15:621.383 3688

Investigations of Near Infra-Red Radiations by Means of Image Converters—A. Vasko. (*Nature*, (London), vol. 158, p. 235; August 17, 1946.) "The spectrum under test is projected on to a photo-electric cathode of the type (Ag)-Cs_2O, Cs, Ag-Cs. . . . [The photo-electrons] are focused by an electronic lens and form a picture on the fluorescent screen. . . ."

539.16.08 3689

Various paper on Geiger-Müller and other Particle Counters. (*See* 3759 to 3762.)

621.316.7 3690

Industry Studies New Circuit Technics—R. R. B. (*Elec. Ind.*, vol. 5, pp. 66–67, 130; September, 1946.) A general article comparing electronic control with pneumatic, hydraulic, and mechanical systems.

621.317.39:620.172.222 3691

Electrical Resistance Wire Strain Gauges—(*Engineering*, (London), vol. 162, pp. 121–123; August 9, 1946.)

621.317.49 3692

Magnetism and the Testing of Materials—R. V. Baud. (*Engineering* (London), vol. 162, pp. 41–42; July 12, 1946.) The theory is given of (a) the X-ray absorption method, (b) the magnetic method, used in flaw detection, and the merits of each discussed. An abridged translation from *Schweiz. Tech. Z.*, p. 515; October 11, 1945.

621.365.5:669.14:621.785.6 3693

The Surface Treatment of Metals and in Particular the Surface Hardening of Steel by High-Frequency Currents—R. Casti. (*Brown Boveri Rev.*, vol. 31, pp. 306–308; September, 1944.) A discussion on the main factors which determine the thickness of the hardened zone, with brief notes on the design of a suitable heating coil. The results of tests on surface hardening of a chromium-steel disk, and of a carbon-steel shaft are given. These prove that variation of heating time provides a means of fixing the thickness of the hardened layer without affecting the degree of hardness.

621.365.92:674 3694

Joints in a Jiffy [by Means of Electronic Heating of Wood]—J. Markus. (*Sci. Amer.*, vol. 174, pp. 245–248; June, 1946.)

621.365.92:678 3695

Capacity-Current Heating in the Rubber Industry—T. H. Messenger. (*Electronic Eng.*, vol. 18, pp. 270–272, 276; September, 1946.)

621.365.92:679.5 3696

Hardening of Plastics by High-Frequency Power—H. Stäger and F. Held. (*Brown Boveri Rev.*, vol. 31, pp. 298–305; September, 1944.) The changes in a plastic when subject to hardening by hot plate presses and by high-frequency power are outlined.

A series of tests carried out on pure resins and on laminated wood is described. The 23-megacycle generator provided 1.5 kilowatts for heating small cylindrical specimens of resin and 3-centimeter cubes. The temperature of the specimen was measured by thermocouples inserted at the center and at 0.5 centimeter from the surface. The relations between applied voltage, heating time, and temperature are shown graphically. A comparison was made of the hardening produced by hot-plate and high-frequency power methods, and the results are given.

621.38:655.324.5 3697

Electronic Register Control for Multicolor Printing—W. D. Cockrell. (*Trans. A.I.E.E.* (*Elec. Eng.* August–September, 1946), vol. 65, pp. 617–622; August–September, 1946.)

621.38.001.8 3698

Tubes on the Job—(*Elec. Ind.*, vol. 5, pp. 98–99; September, 1946.) Describes briefly the use of electronic apparatus in (a) mobile radio for trucks, (b) rubber weighing, (c) a wheel balancer, (d) a coin rejector, (e) a wind-up reel regulator, and (f) stress measurements on steel trusses.

621.38.078:6 3699

What Industry Seeks in Electronic Control—P. Ewald, L. C. Roess, H. K. Steele. (*Elec. Ind.*, vol. 5, pp. 85–86, 116; September, 1946.) Chemists need reliable and corrosion-proof electronic measuring instruments.

In the oil industry, mass spectrometers, absorption spectrometers, electronic computers, knock meters, and devices for measuring electrical conditions causing corrosion could usefully be developed. The food industry needs improved devices for color determination, rapid heating, and determination of moisture.

621.383.001.8 3700

Reading Aid for the Blind—V. K. Zworykin and L. E. Flory. (*Electronics*, vol. 19, pp. 84–87; August, 1946.) A light synchronized with a frequency-modulation audio-oscillator moves over a line. Reflected light actuates an amplifier by means of a photoelectric cell when the scanning spot is over the black portion of a character. A distinctive sound thus corresponds to each letter.

621.384 3701

Various Papers on Electron Accelerators—(*See* 3786 to 3789.)

621.385:6 3702

Electronic Inspection—V. Zeluff. (*Sci. Amer.*, vol. 174, pp. 59–61; February, 1946.) Use of cathode-ray tubes and other electronic devices for production testing where speed and accuracy are needed.

621.385.001.8:518.5 3703

Electronic Computers—Shannon. (*See* 3655.)

621.385.833+535.317.6 3704

A Magnetic Lens with a Minimum Spherical Aberration—A. G. Vlasoff. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 8, no. 5, pp. 235–239; 1944. In Russian.) The spherical aberration of a magnetic lens is considered, and a formula (1) for calculating it is given. Methods are indicated for deriving conditions under which spherical aberration would be a minimum. The case of a "short" magnetic lens is discussed in greater detail, and the shape of the pole shoes satisfying the required conditions is determined. An abstract in English was noted in 1946 of July.

621.385.833 3705

Calculation of the Fields of Simple Electrostatic Lenses—A. G. Vlasoff. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 8, no. 5, pp. 240–242; 1944. In Russian.) Lenses are considered which represent systems of (a) a number of plane metallic electrodes perpendicular to the optical axis, and having circular apertures with their centers on the optical axis, and of (b) a number of cylindrical surfaces with their axes coinciding with the optical axis. A function is found satisfying Laplace's equation within the space bounded by the electrodes, and passing through given values at the electrodes. It is shown that the problem can be reduced to that of Dirichlet for the case of a cylinder, and, starting from Laplace's equation, a solution (10) is found which satisfies all conditions of Dirichlet's problem. An abstract in English was noted in 1945 of July.

621.385.833 3706

The Electrostatic Electron Microscope—H. Bruck and P. Grivet. (*Onde Élect.*, vol. 26, pp. 175–227; June, 1946.) From the Electron Optics Laboratory of the Compagnie Générale de Télégraphie Sans Fil. Details of the theory and design, illustrated by examples of photomicrographs. It is claimed that the electrostatic type, though slightly inferior to the magnetic type in resolving power, gives images richer in contrast, and is less exacting in its power-stabilization requirements.

621.385.833:537.533 3707

Extraction of Electrons by an Electric Field—Lukirski. (*See* 3581.)

621.389:535.61-15 3708

Sight at Night—V. Zeluff. (*Sci. Amer.*, vol. 175, pp. 21–23; July, 1946.) Possible uses of infra-red beams for travel at night or in fog.

621.39.083.7:539.89 3709

Telemetering for Project Crossroads—J. W. Colton. (*Elec. Ind.*, vol. 5, pp. 76–79, 135; September, 1946.) Engineering details of the equipment for recording the air and eater pressures in the Bikini atomic bomb tests.

622.19:621.396 3710

The Impedance Method of Radio Prospecting. Practical Applications: Parts 1 and 2—V. Fritsch. (*Arch. Tech. Messen*, pp. T75–76, and T90; July and August, 1940.) The measurement of resistance and capacitance between electrodes provides geological data which are of particular use in tectonic regions. Fissures, faults, cavities, and water pockets can be detected from the capacitance variations observed as the electrodes are moved over the region. In interpreting measurements on glaciers the dependence of resistivity and dielectric constant of ice on temperature, frequency, and impurities must be taken into account. The detection of coal, salt, and ores is discussed. A table of the resistivities of various geological materials in the dry state is given. Other parts have been referred to in 1942 and 1943 abstracts.

622.19:623.26:621.396.9 3711

Treasure Finding Modernized—W. E. Osborne. (*Radio News*, vol. 36, pp. 30, 94; September, 1946.) General article on the adaptation of mine detectors to the location of metals and other materials, with mention of the various types of equipment used.

623.26:621.396.9 3712

The Problem of Land-Mine Detection. Detection of Metallic Masses of Small Dimensions—H. Grumel and P. Morel. (*Ann. Radioélect.*, vol. 1, pp. 160–167; January, 1946.) A review of the problems of mine detection and of the electrical, military, and practical requirements for a detector. The S.F.R. model 451 is described in detail. It contains an exploring head consisting of two tuned iron-cored coils having mutually perpendicular axes to give zero coupling. One coil is part of the resonant circuit of a push-pull oscillator cat about 280 cycles and the other is followed by a 2-tube tuned amplifier (gain about 70 decibels) feeding headphones. Two magnetic zero adjustments are provided to eater for in-phase and quadrature effects. With this detector the German "Schuhmine" could be detected easily at a depth of 10 centimeters, and an anti-tank mine at 55 centimeters.

623.454.25:621.396.9 3713

Radio Proximity Fuze Design—W. S. Hinman, Jr., and C. Brunetti. (*Jour. Res. Nat. Bur. Stand.*, vol. 37, pp. 1–13; July, 1946.) Particularly for bombs, rockets and mortars. The fuses operate on the principle that the Döppler beat between the electromagnetic radiation from an antenna on the projectile and the reflected radiation from a target is received by the same apparatus, amplified and used to control the detonator.

A description is given of the aerials, oscillating detectors, amplifiers, and power supplies used, with a short account of the production, laboratory testing, and field testing processes.

621.317.755:6 3714

The Cathode-Ray Oscillograph in Industry [Book Review]—Wilson. (*See* 3797.)

621.385.833 3715

Introduction to the Electron Microscope [Book Review]—F. E. J. Ockenden. Williams and Northgate (Norgate), London, 1946, 24 pp., 27 figs., 2s. 6d. (*Elect. Rev.*, (London), vol. 139, p. 430; September 13, 1946.) ". . . [this monograph, the second of a series] can be recommended to those who need a clear explanation . . . of the three principles on which electron microscopy is based."

621.386.1 3716

X-Rays in Practice [Book Review—W. T. Sproull. McGraw-Hill, London, 615 pp., 30s. (*Jour. Sci. Instr.*, vol. 23, p. 218; September, 1946.)

PROPAGATION OF WAVES

538.566.029.63:535.42 3717

Diffraction Pattern of a Circular Aperture at Short Distances—C. L. Andrews. (*Phys. Rev.*, vol. 69, p. 684; June 1–15, 1946.) Measurements were made for wavelength of 12.8 centimeters with aperture of 1 to 6 wavelengths. Fresnel zone theory could be used as an approximate guide. Checks were made against Kirchhoff's theorem. Summary of American Physical Society paper.

538.566.029.64:535.43:551.48 3718

The Frequency Dependence of Radar Echoes from the Surface of the Sea—H. Goldstein. (*Phys. Rev.*, vol. 69, nos. 11/12, p. 695; June 1–15, 1946.) The scattering cross-section σ per unit area of sea surface has been measured on wavelengths of 9.2, 3.2, and 1.26 centimeters. Although the $1/\lambda^4$ Rayleigh law, which would indicate scattering by spray droplets, is not observed, the changes of σ with polarization and with frequency are difficult to explain on the basis of scattering from large sea surfaces. Summary of American Physical Society paper.

621.396.11+538.569.4].029.64:551.57 3719

Absorption of Microwaves by Water Vapor—S. H. Autler, G. E. Becker, and J. M. B. Kellogg. (*Phys. Rev.*, vol. 69, p. 694; June 1–15, 1946.) A new method, using a cubical copper cavity of $8\frac{1}{2}$-foot side, within which thermocouples are placed at random, enables the loss of water vapor at various humidities to be measured. Over the range wavelengths from 0.7 to 1.69 centimeters there is an absorption peak at wavelength of 1.33 centimeters where the attenuation is 0.024 decibel per kilometer for 1 gram per cubic meter of water vapor. Summary of American Physical Society paper.

621.396.11:551.51.053.5 3720

Wave-Treatment of Propagation of Electromagnetic Waves in the Ionosphere—M. N. Saha and B. K. Banerjea. (*Indian Jour. Phys.*, vol. 19, pp. 159–166; October, 1945.) "Wave-equations for the propagation of electromagnetic waves through the ionosphere have been obtained by the use of a new mathematical method involving the use of dyadic analysis introduced by Gibbs. Expressions for steady-current conductivity of the ionosphere have been obtained by this method and the results agree with those of Chapman; an extra term for the conductivity which is more prominent in the F_2-layer, has been obtained.

"It has been shown that the wave is split up into three waves, as in Zeeman effect, one of which is ordinary, the other two extraordinary, in accordance with observations by Toshniwal and Harang."

621.396.11:551.51.053.5 3721

Short-Wave [Ionospheric] Forecasting—T. W. Bennington. (*Wireless World*, vol. 52, pp. 246–250; August, 1946.) As a result of wartime needs, ionospheric stations were set up all over the world. By means of their observations, world contour charts both of existing ionospheric conditions and of predicted conditions for any month of the year were constructed. The charts are customarily drawn in terms of the maximum usable frequency for 2500 miles miles (the maximum distance of travel of a radio wave with only one reflection by the ionospheric layers). It was found that the ionization of the layers causing reflection depended on both the geographic and the geomagnetic latitude and longitude of the observing station. To take account of this the world is divided into three zones, bounded by certain geomagnetic meridians, and separate charts are issued for each zone. To be continued.

RECEPTION

534.78 3722

Effects of Amplitude Distortion upon the Intelligibility of Speech—Licklider. (*See* 3524.)

534.78 3723

Effects of Frequency Distortion upon the Intelligibility of Speech—Egan and Wiener. (*See* 3525.)

534.78 3724

Correlation of the Audio Characteristics of Communication Systems with Measured Articulation Scores—Beranek. (*See* 3526.)

621.314.632.029.6 3725

Noise in Radar Crystal Detectors—L. I. Schiff. (*Phys. Rev.*, vol. 69, p. 682; June 1–15, 1946.) A calculation of noise due to shot effect of electrons crossing the blocking layer. A frequency-independent spectrum is deduced. Summary of American Physical Society paper.

621.395.623.6 3726

Development of Midget Earphones for Military Use—Pearson, Mundel, Carlisle, Knauert, and Zaret. (*See* 3534.)

621.395.623.7 3727

The Output Stage—A. W. Stanley. (*Wireless World*, vol. 52, pp. 256–259; August, 1946.) Curves are given of the variation with frequency of resistance and reactance of a typical moving-coil loudspeaker. By considering these and the equivalent circuit of the output stage of an audio amplifier, the theoretical effects of correct matching are deduced. Frequency response curves taken under practical conditions are reproduced, showing that a knowledge of the response curve of a loudspeaker is useless unless the impedance of the driving source is known.

621.395.667 3728

Fundamental Tone Control Circuits—A Reference Sheet—(*Electronic Eng.*, vol. 18, pp. 278–279; September, 1946.) Full circuit details of resistance and capacity networks, for obtaining the four chief types of tone control.

621.395.92 3729

Desirable Frequency Characteristics for Hearing Aids—Davis, Hudgins, Peterson, and Ross. (*See* 3538.)

621.396.619 3730

Comparison of Frequency Modulation and Amplitude Modulation—T. J. Weijers. (*Philips Tech. Rev.*, vol. 8, pp. 89–96; March, 1946.) The relative merits of amplitude and frequency modulation in relation to noise and interference, and the advantages of frequency modulation with wide frequency sweep and adequate receiver limiting for high-fidelity broadcasting, are discussed.

621.396.619.018.41:621.396.621 3731

Linear Frequency Discriminator—J. R. Tillman. (*Wirless Eng.*, vol. 23, pp. 281–286; October, 1946.) The disadvantages of discriminators which include inductances (e.g. variability with temperature) are described, and the design of a resistance-capacitance discriminator is discussed. This design is based on the output versus frequency characteristic of the Wien bridge near the balance frequency, and consists of two twin-T networks having suitably staggered balance frequencies. The linearity is good over the whole required range of frequency. This is true even for wide-band circuits if additional networks are used. The performance can probably be made substantially independent of temperature.

621.396.62.029.62 3732

Mobile Receiving Equipment for Two, Six and Ten Meters—E. P. Tilton. (*QST*, vol. 30, pp. 28–35; September, 1946.) The system described uses three units; (a) An intermediate-frequency amplifier using a frequency of 11 megacycles, super regenerative detector and audio stage; (b) A converter to cover two frequency ranges, 27 to 30 megacycles and 50 to 54 megacycles; (c) a converter for the 144 to 150 megacycles band.

621.396.621+621.396.61]:621.396.932 3733

Transmitting Equipment for the Merchant Navy—Grumel. (*See* 3805.)

621.396.621 3734

An Amateur-Built Eight-Valve Communications Superhet—E. W. Nield. (*R.S.G.B. Bull.*, vol. 22, pp. 50–54, 57; October, 1946.) A circuit design incorporating what are considered to be the most desirable features of various existing designs. The performance claimed is well up to commercial standards; 1 watt is delivered by the loudspeaker for an input voltage to the receiver of less than 1 microvolt (30 per cent sine wave modulation). The automatic volume control is substantially flat above an input of about 5 microvolts.

621.396.621 3735

Modern Home Receiver Design—Z. Benin. (*Electronics*, vol. 19, pp. 94–98; August, 1946.) The design of a commercial receiver suitable for frequency-modulation operation in the 88 to 108 megacycles portion of the spectrum as well as in the usual broadcast bands is considered in some detail. Necessary compromises in stage design to ensure satisfactory performance in the two roles are discussed. Particular attention is paid to the aerial circuits, intermediate-frequency transformer design and frequency stability of the conversion oscillator.

621.396.621.5 3736

Frequency Deviation Reception—D. A. Griffin. (*Proc. Radio Club Amer.*, vol. 23, pp. 3–7; April, 1946.) A description of a method of achieving increased selectivity and signal to noise ratio in the reception of keyed continuous-wave telegraphy signals. It is based upon the frequency modulation of the signals in the receiver itself in such a way that the desired signal is brought periodically within a narrow pass band while nearby unwanted signals remain just outside this pass band. An additional possibility is that by the suitable location of the mean frequency of the wanted signal relative to a narrow pass band, its frequency modulation can be converted into an amplitude modulation with a major component of twice the modulation frequency, whereas that of nearby unwanted signals remains at the fundamental. Thus audio-frequency discrimination can be used to gain additional selectivity.

Means of realizing these possibilities by the use of a double frequency-change process, with frequency modulation of the first conversion oscillator, are described.

621.396.621.53.018.41.012.7 3737

Mixer Frequency Charts—R. S. Badessa. (*Electronics*, vol. 19, p. 138; August, 1946.) The charts deal with either sum or difference frequency mixers, and show the combinations of high order frequency components which are capable of beating together to give a resultant frequency coinciding with the desired output value.

621.396.645:629.135 3738

Isolation Amplifiers for Aircraft—G. F. Rogers. (*Electronics*, vol. 19, pp. 122–123; August, 1946.) A device enabling members of an aircraft crew to listen as required to the various aircraft receivers without mutual interference. Each member is provided with his own single tube amplifier, to the control grid of which the receiver outputs are connected through decoupling resistors and selector switches. The grid circuit constant and gain of the amplifier are chosen so that satisfactory operation is achieved with up to five input channels.

621.396.82 3739

BCI [Broadcast Interference]—G. G. (*QST*, vol. 30, pp. 54–55; September, 1946.) Interference caused by amateur transmitters in the vicinity of broadcast receivers has been found due in certain cases to radio-frequency voltages being rectified at the grid of the first audio amplifier in receivers having poor shielding such as the midget a.c.-d.c. types. The trouble may be reduced or entirely overcome by inserting a suitable resistance in the grid circuit so that the grid capacitance and this resistance form a low-pass filter to the radio-frequency voltages. Alternative methods are mentioned using suitable by-pass capacitors in the grid circuit.

621.396.822.029.6 3740

Signal-Noise Ratio at V.H.F.—M. J. O. Strutt and A. van der Zeil. (*Wireless Eng.*, vol. 23, pp. 241–249; September, 1946.) Analyses are given of the signal to noise ratios of the grounded-cathode and grounded-grid triode amplifiers, assuming the use of equivalent noise current generators. The partial coherence to the induced grid noise and the anode to cathode noise determines the optimum values of the grid circuit detuning and aerial coupling; experimental confirmation is given (cf. Strutt and van der Ziel, 749 of 1945). Similar results are obtained for the velocity-modulation amplifier (Müller, 48 of 1943). The paper concludes with a discussion of tube and circuit design for maximum sensitivity The authors' noise ratio *w* is compared with other definitions of noise factor.

621.396.823:621.43.04 3741

Motor-Car Ignition Interference—C. C. Eaglesfield. (*Wireless Eng.*, vol. 23, pp. 265–272; October, 1946.) A simple theory is given in which the spark gap is replaced by a switch, and the radiating part of the ignition system by a small loop close to the plane earth. The initial current in the loop and hence the radiated field and the duration of the pulse are obtained. Comparison of the field strength calculated by this means with values measured by other observers shows agreement. The waveforms of the radiated field from a spark plug and from cars were investigated on a television receiver connected to a second cathode-ray tube from which photographs of the pulse trains were taken. From these photographs the number of pulses per train and its average duration for several makes of car were tabulated.

STATIONS AND COMMUNICATION SYSTEMS

621.391.1 3742

Multichannel Communication Systems—F. F. Roberts and J. C. Simmonds. (*Alta Frequenza*, vol. 14, pp. 236–238; September—December 1945.) Long summary in Italian of 183 of January and 416 of February.

621.395.332.029.64(44) 3743

First Microwave Telephone System Operating on Busy Paris Route—(*Telegr. Teleph. Age*, vol. 64, pp. 5, 6; September, 1946.) An announcement of the introduction into regular service by the French Ministry of Posts, Telegraphs and Telephones of a 12-channel, 3000 megacycles frequency-modulated radio link between Paris and Montmorency.

621.395.44 3744

The Unit Bay 1B Coaxial Cable Transmission System: Part 4—R. A. Brockbank and C. F. Floyd. (*P.O. Elec. Eng. Jour.*, vol. 39, part 2, pp. 64–65; July, 1946.(Continuation of 2000 of July, dealing with the installation and operation.

621.396:061.5 3745

High-Frequency and Communications Engineering—K. Sachs and W. G. Noack. (*Brown Boveri Rev.*, vol. 30, pp. 59–64; January–April, 1943.) Review of the Brown Boveri Company's work in 1942 including notes on (a) development of microwave oscillators, and applications of microwaves to distance-measurement and communication problems; (b) secrecy equipment on telephone and telegraphic systems; (c) medium and short-wave transmitters using remote control; (d) broadcast studio equipment; and (e) supervisory remote-control systems applied to power-stations, railways, and industrial problems.

621.396:061.5 3746

High-Frequency and Communications Engineering—K. Sachs and W. G. Noack. (*Brown Boveri Rev.*, vol. 31, pp. 86–93; January—February, 1944.) A review of research and development work undertaken in 1943 by the Brown Boveri Company, including notes on filters using artificially cultivated crystals, frequency-modulation and unidirectional aerials applied to the design of sets for multichannel operation; magnetron developments, and an impedance-measuring device for ultra-high-frequency work; medium-wave broadcast transmitters, and equipment used in police wireless communication; a telemetering system for supervisory control, and a low-voltage network telecontrol system.

621.396.029.56/.62 3747

Amateur Bands—(*Wireless World*, vol. 52, p. 271; August, 1946.) The frequency bands now available to British amateur transmitters are 1.8 to 2.0, 7.15 to 7.30, 14.1 to 14.3, 28.0 to 30.0, and 58.5 to 60.0 megacycles. These may be used for continuous wave, modulated continuous wave, and radiotelephony. The power limitations are given.

621.396.13 3748

A Preview of the Western Union System of Radio Beam Telegraphy: Part 2—J. Z. Millar. (*Jour. Frank. Inst.*, vol. 242, pp. 23–40; July, 1946.) A description of the centimeter-wavelength beam system developed by RCA for inter-city communications. Thirty-two voice-frequency bands in the range 500 cycles to 150 kilocycles modulate the frequency of a 1-megacycle sub-carrier with a peak deviation of 400 kilocycles. This signal in turn modulates the frequency of the radio-frequency transmitter, which has a peak deviation of 2 megacycles. The transmitter has a power output of 50 milliwatts and, with a parabolic reflector to the antenna, gives a range of 50 miles. The effects of atmospheric conditions on the choice of operating frequency are considered, and vertical diversity reception is proposed to overcome interference fading. The repeater stations, which are mounted on steel towers, demodulate the signal to the 1-megacycle sub-carrier, which then modulates a transmitter on a nearby frequency. The proposed relay networks are shown with examples of the terrain profile. For part 1 see 3048 of October.

621.396.13 3749

Reuters' Wireless Services—W. West. (*P.O. Elec. Eng. Jour.*, vol. 39, pp. 48–52; July, 1946.) An account of the development of Reuters' service of news distribution from British Post Office transmitters, including the long-wave "European," and the short-wave "World" services, together with a description of the German Hellschreiber telegraph system.

621.396.24.029.64 3750

Radiotelephone Links on Ultra-Short Waves—Vecchiacchi. (*Alta Frequenza*, vol. 15, pp. 3–14; March, 1946. With English, French and German summaries.) Ultra-high-frequency links can be used advantageously in multichannel long-distance telephony in Italy where the mountains provide natural intermediate relay points. The choice of wavelength, station spacing, modulation system, repeating system, and multiplexing system is discussed. Profile curves illustrate three possibilities for the Milan-Turin link. The Milan-Rome link which is nearly 500 kilometers direct requires only two intermediate stations.

621.396.619.018.41 3751

Frequency Modulation: B.B.C. Field Trials—H. L. Kirke. (*B.B.C. Quart.*, vol. 1, pp. 62–80; July, 1946.) A detailed description of results obtained from the following tests: "1. propagation tests on 45 and 90 megacycles (field strength *versus* distance, for both horizontal and vertical polarization), 2. fading measurements at various distances, 3. comparative tests on frequency modulation and amplitude modulation, 4. signal to noise ratio tests, and 5. practical listening tests with different types of receivers at various distances and in the homes of ordinary listeners." It is thought that the optimum conditions for a British ultra-short-wave frequency modulation broadcasting service are: maximum deviation 75 kilocycles; pre-emphasis 50 microseconds; carrier-channel spacing 200 kilocycles but 400 kilocycles between transmitters serving the same area. See also *Wireless World*, vol. 52, pp. 316–320; October, 1946.

621.396.619.16 3752

Pulse Modulation—A. S. Gladwin. (*Wireless Eng.*, vol. 23, pp. 288–289; October, 1946.) A further contribution to the discussion given in various letters to *Wireless Engineer* since the article by Robert and Simmonds [See 3054 of October and back references). A single general formula is here derived, in terms of which all types of modulated pulse trains with constant amplitude pulses can be represented.

621.396.82 3753

Adjacent Channel Interference—A. G. Dunn. (*R.S.G.B. Bull.*, vol. 22, pp. 55–57; October, 1946.) To relieve congestion on amateur frequencies, it is essential (a) to improve apparatus, e.g. by using beam aerials and the least possible power, (b) to improve operating procedure: operators should be able to read morse well, choose frequency correctly, and use variable frequency oscillators where feasible; unnecessary "netting" should be avoided, (c) to share time and frequency more drastically.

621.396.9.015.33 3754

Pulse Technique and Its Application [to Radar]—A. de Gouvenain. (*Cah. Toute la Radio*, pp. 2–4; September, 1945.) An elementary description.

SUBSIDIARY APPARATUS

534.41+534.781 3755

The Sound Spectrograph—Koenig, Dunn, and Lacy. (*See* 3517.)

534.41+534.781]:535.37 3756

Visible Speech Translators with External Phosphors—Dudley and Gruenz. (*See* 3519.)

534.41+534.781]:621.385.832 3757

Visible Speech-Cathode-Ray Translator—Riesz and Schott. (*See* 3520).

537.525.3:621.316.722.078.3 3758

Characteristics of the Pre-Corona Discharge and Its Use as a Reference Potential in Voltage Stabilizers—S. C. Brown. (*Phys. Rev.*, vol. 69, pp. 696–697; June 1–15, 1946.) Summary of American Physical Society paper.

539.16.08 3759

Radioactivity Meter for Nuclear Research—A. G. Bousquet. (*Elec. Ind.*, vol. 5, pp. 88–89; September, 1946.) A commercial Geiger-Müller counter and associated amplifier for rate counting with tube quenching and meter indication.

539.16.08 3760

Small Mica Window Geiger-Müller Counter for Measurements of Radioactive Isotopes in Vivo—E. Strajman. (*Rev. Sci. Instr.*, vol. 17, pp. 232–234; June, 1946.)

539.16.08 3761

System for High Speed Counting on Nuclear Particles—H. L. Schultz. (*Phys. Rev.*, vol. 69, p. 689; June 1–15, 1946.) Summary of American Physical Society paper.

539.16.08:547.2 3762

Organic Vapors for Self-Quenched G.M. Counters—E. der Mateosian and H. Friedman. (*Phys. Rev.*, vol. 69, p. 689; June 1–15, 1946.) Summary of American Physical Society paper.

621-526 3763

Parallel Circuits in Servomechanisms—H. T. Marcy. (*Trans. A.I.E.E.* (*Elec. Eng.* August–September, 1946), vol. 65, pp. 521–529; August–September, 1946.) A general mathematical technique for the analysis of servomechanisms in which the inclusion of a component in the feedback circuit modifies the controlled quantity before comparison with the desired quantity. A bibliography of 8 items is given.

621-26 3764

The Frequency Response of Automatic Control Systems—H. Harris, Jr. (*Trans. A.I.E.E.* (*Elec. Eng.* August-September, 1946), vol. 65, pp. 539–546; August-September, 1946.) The accepted method of transient response analysis of servomechanisms becomes unwieldy with complex systems. A frequency response method based on Fourier analysis is advocated which gives equivalent results more simply. A numerical examples is given of a torque amplifier with motor generator control.

621.314.12:621.394.396].66 3765

The Amplidyne Electrical Control System—British Thomson-Houston Company. (*Engineering*, (London), vol. 162, pp. 103–105; August 2, 1946.) The machine is essentially a direct-current generator with an unusually low field-power, which can be used for automatic control of voltage, current, or power factor with good transient response. A description of several applications is given.

621.314.2:621.396.619 3766

Modulation Transformers for Broadcasting Transmitters—M. G. Favre. (*Brown Boveri Rev.*, vol. 31, pp. 323–326; September, 1944.) For 100 per cent modulation, the highest quality is obtained with anode modulation of the final radio-frequency amplifier. The requirements for a class-B modulator and the design of the modulation transformer are given. The response curve of an 8.5-kilowatt modulator indicates a loss of 0.5 decibel at 30 and 10,000 cycles when the transformer has an efficiency of 95 per cent.

621.314.53 3767

Rotary Converter for Portable Power Supplies—(*Electronics*, vol. 19, p. 142; August, 1946.) Coils rotating in a magnetic field carry with them an evacuated glass sphere containing mercury, which makes contact successively with tungsten electrodes. "The complete cased unit resembles a conventional vibrator in appearance."

621.314.6+621.319.4+621.383]:669.018 3768

Light Alloys in Metal Rectifiers, Photocells and Condensers—A series of anonymous articles under this or similar title has appeared in various issues of *Light Metals* since April, 1944:

(i) vol. 7, pp. 162–172; April, 1944. "... the theory and practice of the use of aluminium and magnesium [in metal rectifiers] are examined;"

(ii) vol. 7, pp. 276–298; June, 1944. "Particular attention is [here] paid to the selenium rectifier, and the use made of light metals in its construction;"

(iii) vol. 7, pp. 437–458; September, 1944. "Concluding ... a study of the selenium rectifier, and introducing a comprehensive discussion on photocells and the role of light metals in their construction;"

(iv) vol. 7, pp. 505–512; October, 1944. "Continuing ... a discussion on photocells. The copper-oxide and caesium types are here dealt with";

(v) vol. 7, pp. 525–529; November, 1944. "Apparatus, auxiliary materials, and technique employed in preparing and handling alkali metals for photocells are described. Physical and chemical properties of these metals and methods for their extraction are then briefly discussed."

To be continued.

621.317.755 3769

Modification to Cossor Oscilloscope Model 339 to Enable Modulation Measurements to be Made at Carrier Frequencies above 20 Mc/s—A. J. Muir and J. W. Whitehead. (*Jour. Sci. Instr.*, vol. 23, No. 8, p. 189; August, 1946.)

621.317.755 3770

Fast Sweep Synchroscope—D. F. Winter. (*Phys. Rev.*, vol. 69, p. 695; June 1–15, 1946.) The basic problems involved in building a sealed tube cathode-ray oscillograph for measuring time intervals of 10^{-9} seconds to within 10 per cent are listed. Summary of America Physical Society paper.

621.317.755 3771

Elements of a New Oscilloscope Design—Simmons. (*See* 3677.)

621.317.755.027.3 3772

The Precision-Type Quadruple-Beam High-Voltage Oscillograph—G. Induni (*Brown Boveri Rev.*, vol. 30, pp. 222–223; September–October, 1943.) One cathode is used and there are four independent deflection assemblies for voltages up to 3 kilovolts and two deflection assemblies for voltages up to 50 kilovolts. The high voltage is supplied to the cathode through a screened lead from an oil-immersed 50-kilovolt direct-current rectifier plant. The cast-iron body of the oscillograph is highly vacuum-tight and provides excellent screening, rubber packing helps to maintain the vacuum, and one molecular pump suffices. The beams are independently adjusted for intensity and shift, and a common focusing coil is used.

621.318.24 3773

Condenser-Discharge Magnetiser for Permanent Magnets—F. Brailsford. (*Engineering*, (London), vol. 162, pp. 145–146; August 16, 1946.) Permanent magnets may be magnetized conveniently by passing "a high unidirectional current for a short time through a single conductor threading the magnet". This has been done by discharging a capacitor through the primary winding of a transformer with the secondary connected to the single magnetizing conductor. The circuit must be critically damped to obtain maximum peak current without oscillation. See also 2719 of September.

621.318.42+621.396.662.21 3774

Properties and Application of Standard-*Q* Coils at High Frequency—G. Opitz. (*Arch. Tech. Messen*, pp. T106–107; September, 1940.) The self-capacitance C_s of a coil of standard $Q(=\omega L/r)$ can introduce errors according to the method of measurement. The construction of a series of iron-cored standards from 1 microhenry to 1 henry ($Q=50$–250) is described, and the influence of the iron on the loss and inductance of the coil (as a function of magnetization) is discussed. A 20 per cent change of Q is possible for a ±10 degrees centigrade temperature change or a variation of 0–80 per cent of relative humidity.

621.318.5 3775

Electric Relays: a General Survey—J. Sorge. (*Arch. Tech. Messen*, pp. T40–41; April, 1940.)

621.318.572 3776

Electronic True Decade Counters—H. G. Shea. (*Elec. Ind.*, vol. 5, pp. 82–84, 136; September, 1946.) Four double triode multivibrators have neon lamps in each anode circuit. By means of feedback to the third and fifth triode sections it is possible to count up to ten input pulses directly thus avoiding conversion to and from the binary scale. Another counter, described in 2496 and 2497 of September, uses ten directly-coupled twin pentodes arranged in a ring and variations of this circuit may be triggered by alternating-current.

621.319.3.027.3 3777

A Compact High Voltage Electrostatic Generator using Sulphur Hexafluoride Insulation—W. W. Buechner, R. J. Van de Graaff, A. Sperduto, E. A. Burrill, L. R. McIntosh, and R. C. Urquhart. **Preparation and Physical Properties of Sulfur Hexafluoride [for use in Generator]**—W. C. Schumb. (*Phys. Rev.*, vol. 69, p. 692; June 1–15, 1946.) The generator can produce over 5 millivolts, and has been used for investigating the dielectric properties of various gases including air. Summary of American Physical Society papers.

621.319.4:620.193.91 3778

Capacitor Life Testing—J. R. Weeks. (*Bell Lab. Rec.*, vol. 24, pp. 296–299; August 1946.) Relationships discovered in the last fifteen years between voltage, temperature, and life form the basis of accelerated tests in which the probable life can be determined in two weeks. The testing apparatus is described.

621.319.4.029.5:621.315.612 3779

High Frequency Ceramic Capacitors—B. M. Vul, and G. U. Skanavi. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 8, no. 4, pp. 194–199; 1944. In Russian.) For practical purposes the following two types are required: (a) compensating and (b) of high stability. The temperature coefficient of permittivity should be negative in the first, and as near as possible to zero in the second. As a result of the present investigation, materials were developed with positive and negative temperature coefficients by combining TiO_2 with MgO ("timag") and $MgC(CO_3)_2$ ("tidol") respectively. A predetermined value of the coefficient can be obtained by using these materials in various proportions. Experimental curves are plotted showing permittivity ϵ and its temperature coefficient, as measured in completed capacitors (Fig. 1), the effect of temperature on capacitance (Fig. 2), and the effect of frequency (Figs. 3 and 4) and temperature (Fig. 5) on the loss angle. Formulas are also quoted for calculating the heating of flat and cylindrical capacitors. A summary in English was noted in 2761 of 1945.

621.383.2+535.215.1 3780

Complex Photocathodes—N. S. Khlebniko (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 8, no. 5, pp. 286–289; 1944. In Russian.) According to de Boer the external photoeffect of complex cathodes is determined by two elementary processes, the photo-ionization of adsorbed atoms, and the movement of the replenishing electrons through the intermediate layer. Investigations of the Sb-Cs type of cathode and later of the Cs-O-Ag type have led the author, however, to the conclusion that complex photocathodes should be regarded as semiconductors with a relatively low value of the work function, operating by the photo-emission of electrons from the depth of the intermediate layer. The advantages of the new theory are discussed, and further possible developments indicated. An abstract in English was noted in 1819 of July.

621.383.2 3781

Certain Physical Properties of Caesium Oxide Photocathodes—P. M. Morozoff, and M. M. Butsloff. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 8, pp. 291–303; 1944. In Russian.) A preliminary report on a detailed experimental investigation carried out to determine the spectral distribution of sensitivity, the energy distribution of photo-electrons, and the work function of ceasium oxide photocathodes. Curves are plotted in Figs. 2 and 3 showing the spectral sensitivity when the photocathode is illuminated from the rear and the front respectively. The spectral distribution of sensitivity is closely connected with the thickness of the silver layer, and, therefore, with the thickness of the photocathode. Photocathodes of a given thickness, and therefore of a predetermined spectral sensitivity, can be obtained by varying the depth of the oxidation of the silver layer, and Fig. 4 gives sensitivity curves for photocathodes of different thickness. The difficulties of ensuring the required depth of oxidation are pointed out, and the structure of the photocathodes is discussed in detail. The electrical and optical properties of the silver layer are greatly affected by the temperature of the glass envelope during the deposition of the silver. The volt-ampere characteristics of the photocathodes illuminated from the front (thick lines) and from the rear (dotted lines) are shown in Fig. 9. It appears from these curves that the energy distribution of photoelectrons in this type of cathode is similar to that in the case of pure metals. An irregularity in the energy distribution is apparent, however, on wavelengths of 750 millimicrons and more. The reasons for this irregularity are discussed.

Measurements of the work function are described. The value was found to be of the order of 0.78 to 0.90 volts, but may vary with time by as much as ±1 volt over a period of 10 to 500 hours. An abstract in English was noted in 2033 of July.

621.383.2 3782

The Energy Distribution of Electrons and the Relationship Between Photocurrent and the Angle of Incidence of Light in the Case of Caesium Oxide Photocathodes—A. M. Pyatnitski. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 8, pp. 304–308; 1944. In Russian.) Results of an experimental investigation are shown in the following curves; Fig. 1 (left)—the photocurrent against the angle of incidence of light; Fig. 1 (right)—reflection, transmission and absorption of light against the angle of incidence: Fig. 2 (upper curves)—the photocurrent for an angle of incidence of 70 degrees against the wavelength of light; Fig. 2 (lower curves)—reflection transmission and spectral characteristics against the wavelength; Fig. 3 (left)—reflection, transmission and absorption, and Fig. 3 (right)—the photocurrent for different structures of the cathode and different angles of incidence; Fig. 4—volt-ampere characteristics for infra-red light falling at different angles from the front and the rear of the cathode.

Conclusions: (a) The photocurrent depends on the angle of incidence of light. The relationship is determined by the structure of the cathode and the wavelength of the light. (b) The spectral characteristic also depends on the angle of incidence. (c) The maximum of energy distribution, independently of the wavelength, is shifted towards a greater energy value when the cathode is illuminated from the rear. An abstract in English was noted in 2032 of July.

621.383.2 3783

New Photocells with Antimony-Caesium Cathodes—N. S. Khlebnikoff and A. E. Melamid. (*Bull. Acad. Sci. U.R.S.S., sér. phys.*, vol. 8, pp. 309–312; 1944. In Russian.) Two types of photocell were developed, one for use with ultra-violet radiation, and the other possessing constant sensitivity and capable of operating under both weak and intense illumination. For the first type it is difficult to manufacture an envelope transparent to ultra-violet rays, and methods adopted to overcome this are described. The spectral characteristics of the photocells so produced are shown in Fig. 3. The cells are almost free from fatigue, and possess a very high resistance (of the order of 10^{13} Ω) between the anode and the cathode.

The production of the second type is based on the fact that the spectral sensitivity of the Sb-Cs cathodes decreases as a result of fatigue down to 50 per cent of the original value, and is not restored after rest. A photocell can, therefore, be artificially fatigued by exposing it to the illumination of the sky. Photocells with a constant sensitivity (with no greater deviation than ±10 per cent) and capable of operating under such an intense illumination as 10^4 lux were produced in this way. An abstract in English was noted in 2031 of July.

621.383.4+535.215.1:546.28 3784

A New Bridge Photo-Cell Employing a Photo-Conductive Effect in Silicon. Some Properties of High Purity Silicon—G. K. Teal, J. R. Fisher, and A. W. Treptow. (*Phys. Rev.*, vol. 69, p. 686; June 1–15, 1946.) An apparatus is described for making bridge type photocells by reaction of silicon tetrachloride and hydrogen gases at ceramic or quartz surfaces at high temperatures. The variation of the conductivity of pyrolitic silicon films on porcelain with temperature is described and explained. Summary of American Physical Society paper.

621.383.5 3785

Blocking-Layer Photocells—W. C. van Geel. (*Philips Tech. Rev.*, vol. 8, pp. 65–71; March, 1946.) An account of the structure and functioning of photocells formed from a layer of semiconductor such as cuprous oxide or selenium separated from a metal electrode by a very thin insulating layer, the blocking layer. The internal and external photo-effects are explained as the action of light quanta in enabling the electrons to pass from one energy band to another. Examples of the characteristics of typical cells are given.

621.384 3786

The Synchro-Betatron Electron Accelerator Guide Fields—H. F. Kaiser and E. C. Greanias. (*Phys. Rev.*, vol. 69, pp. 536–537; May 1–15, 1946.) Discussion of the possibility of modifying the Kerst betatron for operation as a synchrotron, by the synchronized application of an intense guide field which increases in strength with the energy of the particle. A few calculations show the practicability of the scheme, and possible conductor guide systems are considered.

621.384:537.291 3787

Electron Orbits in the Synchrotron—D. S. Saxon and J. Schwinger. (*Phys. Rev.*, vol. 69, p. 702; June 1–15, 1946.) Equations for the orbital motion are derived on the assumption that the localized accelerating field can be replaced by an equivalent rotating electric field. Summary of American Physical Society paper.

621.384:621.385.16 3788

Preliminary Studies on the Design of a Microwave Linear Accelerator—J. Halpern, E. Everhart, R. A. Rapuano, and J. C. Slater. (*Phys. Rev.*, vol. 69, p. 688; June 1–15, 1946.) Pulsed magnetrons operating at about 3000 megacycles and feeding into high-Q cavities have been used to obtain electron accelerating voltages of the order of 2000 volts. Summary of American Physical Society paper.

621.384:621.392 3789

Wave Guide Acceleration of Particles—E. L. Hudspeth. (*Phys. Rev.*, vol. 69, p. 671; June 1–15, 1946.) The disadvantages of a linear accelerator may be overcome by using the transverse field inside a waveguide to accelerate the particles. The guide is bent into a spiral and holes in the walls allow the particles to move along a diameter. Phasing is obtained by adjusting the position of the holes and the spacing of the turns. The guide is short-circuited and operated as a cavity resonator.

621.385.18.029.64 3790

Various Papers on TR Switches—(*See* 3814 to 3819.)

621.385.832.088 3791

C. R. Tube Quality Measuring Apparatus—(*See* 3683.)

621.394.624 3792

An Electronic Code Translator—H. W. Babcock. (*Electronic Eng.*, vol. 18, pp. 282–283; September, 1946.) Slightly abbreviated reprint of 2730 of September.

621.395.636 3793

Dialling Selection Signals at Voice Frequency—F. Lucantonio. (*Alta Frequenza*, vol. 14, pp. 195–217; September to December, 1945. With English, French and German summaries.) General discussion of systems of dialling signals with particular reference to Italian long-distance underground cables. Schematic and detailed circuits of the relay chains are given.

621.396.611.21.032.2:546.59 3794

Gold Film Electrodes for High Frequency Quartz Plates—R. A. Spears. (*Jour. Brit. Instn. Radio Eng.*, vol. 6, pp. 50–59; March–May, 1946. Discussion pp. 59–62.) A theoretical and practical discussion of the design of gold film electrodes and their effect on the natural frequency. The behavior of electrodes is analyzed to explain the advantages of adherent films. Factors in the design of film electrodes are the surface displacement of the crystal and the optimum area and location of the electrodes. A short account of a practical method of gold sputtering includes details of the preparation of surface and methods of varying quality, color, durability, adherence, thickness, and weight of the film. An analysis is given of the effect of the gold deposit on frequency and activity and usual methods of mounting the crystals are described.

621.396.662 3795

A Device for [Periodic] Variation of Reactance—J. Bernhardt. (*Cah. Toute la Radio* pp. 21–23; September, 1945.) A flexible vibrating blade has one end fixed, a soft iron armature at the center and, at the free end, the element for varying reactance, which can be either (a) a closed wire loop in the field of a coils, (b) a capacitor plate, or (c) a piece of soft iron in the air gap of a magnet. The most convenient frequency of oscillation is 50 to 100 cycles. Various possible applications are mentioned.

621.396.682:621.316.722 3796

Multi-Voltage Regulated Power Supplies—J. R. Mentzer. (*Electronics*, vol. 19, pp. 132–133; September, 1946.) A description of circuits for obtaining outputs at two regulated voltages by the use of standard gas-filled tubes. Any voltage that is a multiple of 15 volts up to the maximum voltage of the unregulated source, can be obtained by suitable additive or subtractive combinations of the tubes. A procedure is given for computing circuit values in relation to voltage and current requirements.

621.317.755:*0* 3797

The Cathode-Ray Oscillograph in Industry [Book Review]—W. Wilson. Chapman and Hall, London, second edition, 1946, 18s. (*Engineering* (London), vol. 162, p. 124; August 9, 1946. *Jour. Sci. Instr.*, vol. 23, p. 218; September, 1946.)

TELEVISION AND PHOTOTELEGRAPHY

621.385.832.088 3798

C. R. Tube Quality Measuring Apparatus—(*See* 3683.)

621.397:621.396.9 3799

Has Radar Influenced Television Development?—(*Jour. Telev. Soc.*, vol. 4, pp. 220–222; March, 1946.) A discussion before the Television Society led by F. R. W. Strafford. It was generally agreed that radar owed a great deal to television, but had so far had little influence on television development, it may have more in the future if higher frequency systems are introduced. Improved feeders and smaller and lighter components produced for radar could be adapted for television, but would not result in great reduction of cost to the user.

621.397:778.53 3800

Film—The Backbone of Television Planning—R. B. Austrian. (*Jour. Telev. Soc.*, vol. 4, p. 226; March, 1946.) Summary of an address given before the Society of Motion Picture Engineers. The high cost of programs employing individual artists makes it probable that films will eventually be the major source of television transmissions.

621.397.26 3801

Stratosphere Television—(*Jour. Telev. Soc.*, vol. 4, p. 227; March, 1946.) Usable signals have been transmitted over a distance of 240 airline-miles from an altitude of 25,000 feet using only 250 watts of power. Those results agree almost exactly with pre-flight calculations. Transmissions to date have been on three frequencies between 100 and 550 megacycles with one channel devoted to studies of television "ghosting"; another to frequency-modulation transmission; and the third for communications incident to test operation.

A suitable plane for the purpose is a low-wing all-metal monoplane about the size of the B-29, but weighing only one third as much. "Each plane would weigh 20 tons fully loaded." The original proposals were noted in 3970 of 1945.

621.397.611 3802

Portable Video Pickup Equipment—W. A. Howard. (*Electronics*, vol. 19, pp. 124–129; August, 1946.) A portable lightweight television camera and associated control, monitoring and synchronizing equipment, designed to give an output of video and standard synchronizing signals. A detailed description of the camera and each auxiliary unit is given with performance specifications and method of operation.

621.397.645 **3803**

Television V. F. Stage—W. T. Cocking. (*Wireless World*, vol. 52, pp. 265–268; August, 1946.) The advantages of feeding the vision signal on to the cathode of the display tube rather than to the grid are pointed out. A basic circuit, and a suitable synchronization pulse separator stage, are described.

621.397.645 **3804**

Television V. F. Stage—H. Wood. (*Wireless World*, vol. 52, p. 346; October 1946.) Comments on 3803 above.

TRANSMISSION

621.396.61+621.396.621]:621.396.932 **3805**

Transmitting Equipment for the Merchant Navy—H. Grumel. (*Ann. Radioélect.* vol. 1, pp. 264–273; January, 1946.) Describes a telephony transmitter for use by unskilled persons giving an aerial power of 30 watts on 80 to 220 meters. Circuits and details of construction of the transmitter and the associated receiver (mounted in the same unit) are given. Mention is also made of three transmitters for use by skilled persons: (a) 300 watts (A_1 communication or 400 watts (A_2) on wavelengths of 580 to 820 meters, (b) 150 to 200 watts (A_1) on 18, 24, 36 and 48 meters, (c) 75 watts on wavelengths of 600 meters—distress. Details of these equipments will be published later.

621.396.61.029.54 **3806**

The "Monobloc" 10-kW Broadcast Transmitters Type TH. 1308—C. Beurtheret. (*Rev. Tech. Comp. Franç. Thomson-Houston*, pp. 45–52; October 1945.) Description of a medium-wave transmitter suitable for rapid serial production to replace those of the French broadcasting system destroyed during the war. The anode efficiency of the power stage is 40 per cent for carrier or 80 per cent at peak, and the over-all efficiency of the transmitter is 23 per cent on carrier. The harmonic distortion is less than 2 per cent at frequencies of 50 to 4000 cycles for 80 per cent modulation and less than 3 per cent at 95 per cent modulation. The noise modulation is 55 decibels below the level corresponding to 80 per cent modulation. The advantages of the assembly of the transmitter in the form of a single rectangular block are stated.

621.396.611.21.029.62:621.396.662.078.3 **3807**

Crystal Control on 144 Mc—W. W. King. (*QST*, vol. 30, pp. 46–50; September, 1946.) Three transmitters are described having output powers of 5, 20, and 60 watts. The frequency is controlled by a 12 megacycles A.T.-cut crystal operating on an overtone, frequency multiplication being carried out in stages which drive the output tubes. All the transmitters described are plate-modulated.

621.396.615.12.029.5 **3808**

A Simple V.F.O. Crystal Substitute—D. Mix. (*QST*, vol. 30, pp. 13–16; September, 1946.) Practical details of a variable frequency oscillator covering a frequency range from 3500 to 4000 kilocycles. In order to obtain high frequency stability an electron coupled oscillator having a large capacitance in its frequency-determining circuit is followed by two untuned amplifier stages. The oscillator tube is 6SK7 as this is well screened. The amplifier uses two 6F6's, which give little trouble from parasitic oscillations. The high-voltage supply to the oscillator and the first amplifier is stabilized from a regulator tube. The oscillator screen may be keyed without causing serious defects.

621.396.619:621.314.2 **3809**

Modulation Transformers for Broadcasting Transmitters—Favre. (*See* 3766.)

VACUUM TUBES AND THERMIONICS

621.385.1.032.216 **3810**

The Measurement of Differences of Contact Potential and of Saturation Current in Vacuum Tubes Using Oxide Cathodes—R. Champeix. (*Ann. Radioélect.*, vol. 1, pp. 208–235; January, 1946.) The characteristic log $I=f(V)$ (the "residual current") is used, and is obtained when the electrode considered has a retarding potential relative to the cathode. This measurement of contact potential drop requires a knowledge of the saturation current, and a method for measuring it has been studied and evolved for oxide cathodes by using the discharge of a capacitor controlled by thyratrons. The method of interpreting Schottky's law to deduce the true saturation current is indicated. Evidence is given of the new phenomenon of the modification of the slope of the line log $I=f(V)$ when the condition of the receiving electrode is changed. Methods are proposed for stablizing the contact potential drop during the industrial production of radio tubes.

621.385.1.032.216:621.386.1 **3811**

A Study of Oxide Cathodes by X-Ray Diffraction Methods: Part 1—Methods, Conversion Studies, and Thermal Explansion Coefficients—A. Eisenstein. (*Jour. Appl. Phys.*, vol. 17, pp. 434–443; June 1946.) Two methods are described, one studying the conversion process in forming the cathode, the other for measuring the thermal expansion coefficients of barium, strontium, and thorium oxides. "The conversion of an equal molar barium strontium carbonate solid solution, $(BaSr)CO_3$ involves (1) crystal growth in the carbonate, (2) decomposition to the mixed oxides, BaO and SrO, (3) formation of the oxide solid solution, (BaSr)O, and (4) crystal growth in the oxide. A similar sequence of events is observed in the conversion of a mixed carbonate, $BaCO_3+SrCO_3$. Crystal and particle size, growth of carbonates, and crystal growth of oxides are investigated, and possible relationships are discussed."

621.385.16+621.396.9 **3812**

Radar and the Magnetrons—J. T. Randall. (*Jour. Roy. Soc. Arts*, vol. 94, pp. 303–312; April 12, 1946. Discussion, pp. 312, 323.) A paper read before the Royal Society of Arts giving a brief account of the basic principles and history of radar, and, in greater detail, of the development of the cavity magnetron.

The development of high powers at centimeter wavelengths began with the klystron (Varian and Varian, 1939) and was followed by the cavity magnetron (Randall and Boot, 1939–40) to which the improvement of strapping was added in 1941 by Sayers.

621.385.16:621.384 **3813**

Preliminary Studies on the Design of a Microwave Linear Accelerator—Halpern, Everhart, Rapuano, and Slater. (*See* 3788.)

621.385.18.029.64 **3814**

Gas Discharge Switches for Controlling Lower Power Microwave Signals—T-S. Ke and L. D. Smullin. (*Phys. Rev.*, vol. 69, p. 698; June 1–15, 1946.) The keep-alive electrode in a transmit-receiver switch was placed unusually near the radio-frequency gap. With this arrangement, "in a modified 1B24 filled with 12-millimeter nitrogen, 44-decibel attenuation was obtained when the keep-alive current was 0.4 milliampere." Summary of American Physical Society paper.

621.385.18.029.64 **3815**

Phenomenological Theory of the TR Switch Spike—T. Holstein. (*Phys. Rev.*, vol. 69, p. 698; June 1–15, 1946.) Summary of American Physical Society paper.

621.385.18.029.64 **3816**

The Band-Pass TR Switch: Part 1—The Switching Action—M. D. Fiske. (*Phys. Rev.*, vol. 69, p. 699; June 1–15, 1946.) The switch consists of a waveguide section with a number of uniformly spaced resonant breakdown gaps within it. With argon at a pressure of 10 millimeters of Hg breakdown times of 5×10^{-9} seconds are possible. Recovery time with pure argon is long (100 microseconds) but may be reduced by a small addition of water vapor. Summary of American Physical Society paper.

621.385.18.029.64 **3817**

The Band-Pass TR Switch: Part 2—Linear Electrical Characteristics—W. C. Caldwell. (*Phys. Rev.*, vol. 69, p. 699; June 1–15, 1946.) By the combination of several resonant circuits in the waveguide, band-pass characteristics over as 12 per cent wavelength range can be obtained. Summary of American Physical Society paper.

621.385.18.029.64:537.5 **3818**

Cross Sections for Capture of Electrons from TR-Tube Recovery Measurements—C. G. Montgomery, F. L. McMillan, I. H. Dearnley, and C. S. Pearsall. (*Phys. Rev.*, vol. 69, p. 699; June 1–15, 1946.) Summary of American Physical Society paper.

621.385.18.029.64:537.525 **3819**

Low Pressure Gas Discharges in Microwave TR Tube—L. D. Smullin. (*Phys. Rev.*, vol. 69, p. 698; June 1–15, 1946.) Summary of American Physical Society paper.

621.385.2 **3820**

Theory of the Diode—J. K. Knipp. (*Phys. Rev.*, vol. 69, p. 700; June 1–15, 1946.) The behavior of the parallel plate diode for radio frequencies is discussed theoretically, taking account of space charge effects and of the distribution of velocity amongst the electrons. Summary of American Physical Society paper.

621.385.3 **3821**

The Application of Dimensional Analysis to Triode Valves at Very High Frequencies—G. Lehmann. (*Onde Élect.*, vol. 26, pp. 175–187; May, 1946.) After a brief discussion of the advantages and the principles of the dimensional method, it is shown that the performance of similar tubes depends only on the single parameter $\phi=Fd/\sqrt{V}$ where F=frequency, d=a linear dimension of the tube and V=a voltage of the system. The output, gain, voltage manification, etc., of similar tubes can all be expressed in terms of ϕ, which is proportional to the transit angle of the electrons. For tubes having the same type of cathode, the products F^3d and F^4V must be maintained constant for a constant output.

By simplified study of the movements of electrons in a tube in the class-B regime, it can be shown that the characteristics of a tube at low frequencies are conserved without material deterioration up to values of ϕ in the region of 2.5, where F is in megacycles, d is the anode-cathode distance in centimeters, and V is the anode voltage in volts. Also, for this value, the Q of the output circuit will be about 18.

621.385.4.029.6 **3822**

Principles of Operation of the Resnatron—F. W. Boggs. (*Phys. Rev.*, vol. 69, p. 700; June 1–15, 1946.) A powerful ultra-high-frequency tetrode oscillator with substantial transit time, unusual grid, and resonant cavities in the vacuum envelope. Summary of American Physical Society paper.

621.385.5:621.317.723 3823
A New Electrometer Valve—J. A. Darbyshire. (*Electronic Eng.*, vol. 18, p. 277; September, 1946.) A double tetrode, in which each section has the characteristic of a single tetrode of type FP54, for use in a balanced bridge circuit. It has better grid insulation than earlier double-triode tubes.

621.385.832 3824
Ion Burn in Cathode-Ray Tubes—G. Liebmann. (*Electronic Eng.*, vol. 18, pp. 289–290; September, 1946.) Experimental evidence that the ions responsible for fluorescent screen destruction are generated by the activation process of the cathode. Recent suggestions for the suppression of ion burn are outlined. See also a paper by C. H. Bachman (*Gen. Elec. Rev.*, vol. 48, p. 13; 1945.) and 2403 of August.

621.396.611:621.385.1 3825
Modulation and Tuning of Cavity Oscillators by Electron Beams—D. S. Saxon. (*Phys. Rev.*, vol. 69, p. 700; June 1–15, 1946.) A theoretical study of tuning by passing the electron beam through an auxiliary cavity tightly coupled to the oscillator. Summary of American Physical Society paper.

621.396.615.17:621.317.755 3826
Time-Base Converter and Frequency-Divider—P. Nagy and M. J. Goddard. (*Wireless Eng.*, vol. 23, pp. 286–287; October, 1946.) A reply to criticisms of earlier papers. See 2093 of July (Moss) and back references.

621.396.645.014.332 3827
Peak Pulse Currents in Class B Amplifiers—Sturley. (*See* 3563.)

MISCELLANEOUS

001.89 3828
Royal Society Empire Scientific Conference—(*Nature* (London), vol. 158, pp. 136–141; July 27, 1946.) Report of discussions and recommendations on a number of subjects including the use of radar in map-making, cosmic-ray research, improvement in scientific information services, standards of measurement, and commonwealth co-operation in science.

5+6]:778.53 3829
The Future of Scientific Films—A. S. C. Lawrence. (*Jour. Roy. Soc. Arts*, vol. 94, pp. 461–469; June 21, 1946.) Discussion of the requirements for satisfactory films, and their value for various purposes.

620.193:669.14 3830
Corrosion of Stainless Steel Sheet in Marine Atmospheres—(*Jour. Frank. Inst.*, vol. 241, pp. 372–373; May, 1946.) Short note only.

621.3.078 3831
Robot Dynamics—Theory of Non-Linear Automatic Control Systems—M. Avramy. (*Phys. Rev.*, vol. 69, p. 697; June 1–15, 1946.) Summary of American Physical Society paper.

621.317 3832
Scientific instruments in Britain—C. Darwin. (*Engineer* (London), vol. 182, pp. 78–79; July 26, 1946.) Lecture delivered at the Exhibition of British Scientific Instruments in Stockholm, describing the development and outstanding achievements of the industry during the present century, important war-time instruments, and the function of the (British) National Physical Laboratory.

621.396.001.6 3833
Research and Development in Radio Technology—R. A. Collacott. (*Electronic Eng.*, vol. 18, pp. 287–288; September, 1946.)

5+6]:41.3=00 3834
Dictionary of Science and Technology [Book Review]—. Newmark. Pitman, London, 386 pp., 30s. (*Jour. Sci. Instr.*, vol. 23, p. 219; September, 1946.) Intended for use in the fields of chemistry, physics, and engineering. The French, German, and Spanish languages are covered.

621.3(031) 3835
Whittaker's Electrical Engineer's Pocket Book [Book Review]—R. E. Neale (Editor). Pitman and Sons, London, seventh edition, 1946, 938 pp., 30s. (*Electrician*, vol. 137, p. 100; July 12, 1946. *Wireless World*, vol. 52, p. 332; October, 1946.) New edition, almost completely rewritten to bring it into line with modern practice.

621.396 3836
The Wireless Trader Year Book. [Book Notice]—Iliffe and Sons, London, 1946, 160 pp., 10s. 6d. (*Electrician*, vol. 137, p. 185; July 19, 1946.)

621.396(075) 3837
An Experimental Course in the Fundamental Principles of Radio [Book Review]—R. H. Humphrey. Pitman, London, 194 pp., 12s. 6d. (*Electronic Eng.*, vol. 18, p. 292; September, 1946.)

621.396(075) 3838
Radio Communications [Book Reviews]—W. T. Perkins and R. W. Barton. George Newnes, London, 312 pp., 12s. 6d. (*Elect. Rev.* (London), vol. 139, p. 296; August 23, 1946; *Wireless World*, vol. 52, p. 307; September, 1946.) ". . . primarily a manual of the 'question and answer type,' intended for students. . . . "

621.396.3(075) 3839
Handbook of Technical Instruction for Wireless Telegraphists [Book Review]—H. M. Dowsett and L. E. Q. Walker. Iliffe and Sons, London, eighth edition, 660 pp., 30s. (*Electronic Eng.*, vol. 18, pp. 292–293; September, 1946.) ". . .It is an essential requirement for passing the P.M.G. Certificate and can be confidently recommended to all prospective wireless telegraphists."

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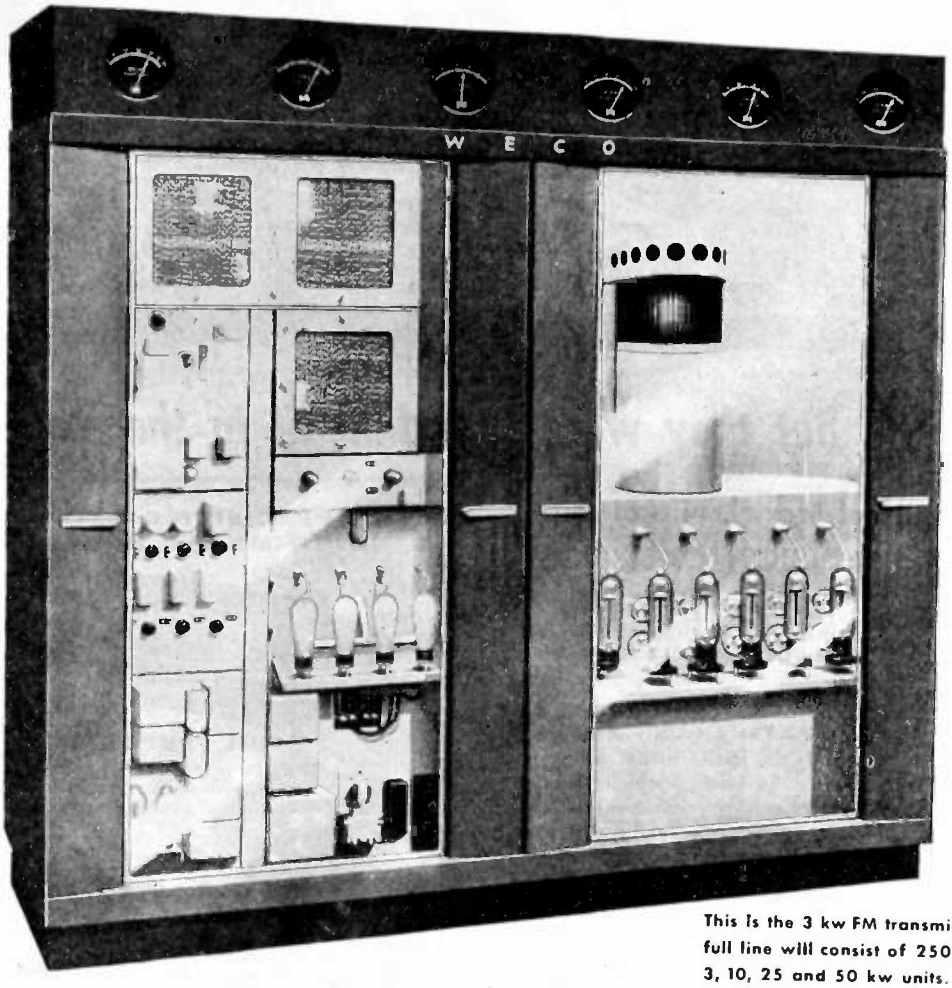

W E C O

SECTION MEETINGS

ATLANTA

"Results of FM field strength survey of Georgia Tech Station W4XAG," by M. A. Honnell, Georgia School of Technology; October 25, 1946.

BOSTON

"Charting the Course of the Institute of Radio Engineers," by F. B. Llewellyn, President of the Institute of Radio Engineers; November 21, 1946.

BUFFALO-NIAGARA

"Charting the course of the Institute of Radio Engineers," by F. B. Llewellyn, President of the Institute of Radio Engineers; November 14, 1946.

CEDAR RAPIDS

"K Carrier in the Bell System," by F. Willie, American Telephone and Telegraph Company; November 13, 1946.

CHICAGO

"The Connsonata," by E. L. Kent, C. G. Conn Ltd.; November 15, 1946.

CINCINNATI

"German Infra-Red Detection Equipment," by R. A. Underhill, Wright Field; November 19, 1946.

COLUMBUS

"The Traveling Wave Tube," by J. R. Pierce, Bell Telephone Laboratories; November 15, 1946.

DALLAS—FT. WORTH

"Tuning Up and Adjusting the Franklin Antenna," by H. J. Lovell, WKY, Oklahoma City; October 22, 1946.

"New Methods in Air Navigation," by H. I. Metz, C. A. A. Experimental Station, Indianapolis; November 5, 1946.

"An Operators' FM Transmitter," by J. R. Boykin, Westinghouse Electric Corporation; November 19, 1946.

DAYTON

"Magnetrons and Klystrons," by B. L. Griffing, Wright Field; November 14, 1946.

DETROIT

"The Detroit News Television Station WWDT," by L. A. Spragg, Station WWDT; November 22, 1946.

"Television Receiver Servicing," by C. E. Quinn, Lawrence Tech Television Club; November 22, 1946.

HOUSTON

"The Psychological Selection of Aircrews for the Army Air Forces and the Application of Experimental Psychology to the Formulation of Engineering Design Principles for Aircraft Instruments and Controls," by J. Buel, Vector Manufacturing Company; November 13, 1946.

INDIANAPOLIS

"Radar-Air and Marine Navigation," by H. J. Geist, Raytheon Manufacturing Company; October 25, 1946.

"Industrial Research Today," by J. E. Hobson, Armour Research Foundation; November 22, 1946.

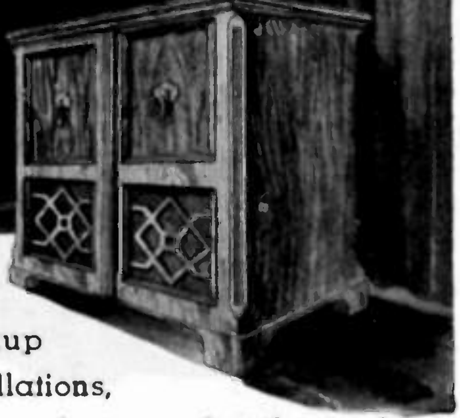
LONDON

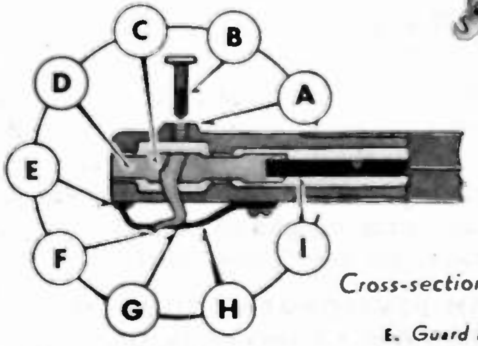
"The Design and Application of Small Panel Instruments," by J. R. Bach, Bach-Simpson Company; November 22, 1946.

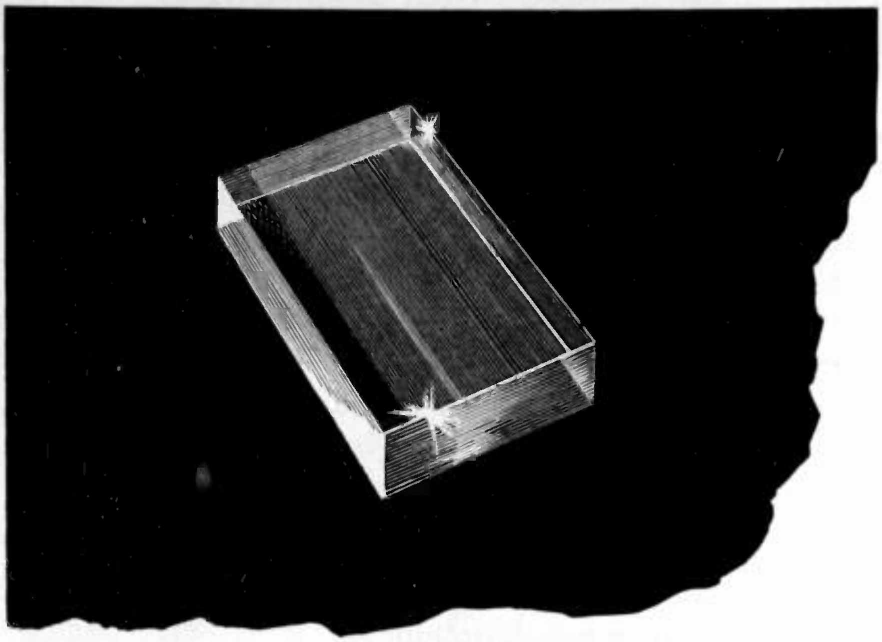
MILWAUKEE

"Microwaves and Radar," by J. O. Perrine, American Telephone Company; October 11, 1946.

(Continued on page 36A)







SECTION MEETINGS

(Continued from page 35A)

MONTREAL

"Electronic Developments in Australia," by J. G. Reed, Amalgamated Wireless (Australasia) Ltd.; November 6, 1946.

"Demonstration Talk on Canadian FM and Television," by J. A. Ouimet, Canadian Broadcasting Corporation; November 14, 1946.

NEW YORK

"Electronics at Bikini—Symposium on Operation Crossroads," by C. L. Engleman, United States Navy; November 6, 1946.

"Instrumentation for Physical Measurements," by T. D. Hanscome, United States Navy; November 6, 1946.

"Techniques for Remote Control and Measurement," by D. G. Fink, Executive Editor, Electronics; November 6, 1946.

"Bikini Observations and Their Significance," by H. Pratt and F. VanDyck, United States Navy; November 6, 1946.

OTTAWA

"Television," by K. R. Patrick, RCA-Victor Company; November 19, 1946.

PHILADELPHIA

"A Single Stage FM Detector," by W. E. Bradley, Philco Corporation; November 7, 1946.

PITTSBURGH

"The Bikini Tests," by J. A. Hutcheson and A. O. McCoubrey, Westinghouse Research Laboratory; October 14, 1946.

PORTLAND

"The Behavior of Dielectrics over Wide Ranges of Frequency, Temperature and Humidity," by R. F. Field, General Radio Company; October 11, 1946.

ST. LOUIS

"Radio Aids to Air Traffic Control," by R. C. Jones, Civil Aeronautics Authority; November 22, 1946.

SAN DIEGO

"The Behavior of Dielectrics over Wide Ranges of Frequency, Temperature and Humidity," by R. F. Field, General Radio Company; October 22, 1946.

TORONTO

"Electronics in the Navy," by H. G. Burchell, Royal Canadian Navy; November 4, 1946.

TWIN CITIES

"Microwaves and Radar," by J. O. Perrine, American Telephone and Telegraph Company; October 17, 1946.

WASHINGTON

"Enemy Electronic and Optical Equipment," by N. J. Granger, War Department; September 10, 1946.

"Synchronization of Oscillators," by R. D. Huntoon, National Bureau of Standards; October 14, 1946.

"German Communication Techniques and Color Photography Developments," by R. H. Ranger, War Department; November 5, 1946.

WILLIAMSPORT

"Color Television," by C. G. Hylkema, Columbia Broadcasting System, Inc., November 6, 1946.

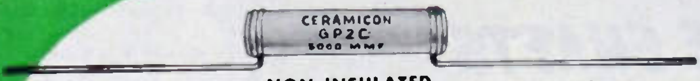
SUBSECTIONS

PRINCETON

"The Traveling Wave Tube," by J. R. Pierce, Bell Telephone Laboratories; November 6, 1946.

ERIE "GP" CERAMICONS*

The Complete, Long-Established Line of General Purpose Ceramic Condensers

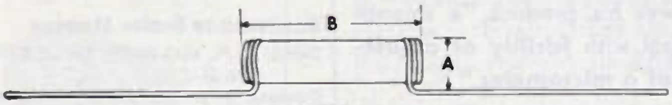


NON-INSULATED

INSULATED

THIS chart gives a complete selection of dependable, compact Erie general-purpose Ceramicons. Specially priced, these time-proven units are made for such applications as coupling and by-passing where temperature coefficient is not important—in other words for all receiver applications except in frequency determining circuits. For the latter type of applications the regular line of temperature-compensating Ceramicons is available.

Ceramics have been proven to be the best dielectrics—Erie Ceramicons have been proven to be the best in ceramic condensers. *Use them to your advantage.*



NON-INSULATED

INSULATED

UNITS STAMPED WITH CAPACITY AND TOLERANCE			
		Max. Overall Dimensions	
		A	B
Insulated	GP1K. GP2K	.250"	.562"
	GP1L. GP2L	.250"	.812"
	GP1M. GP2M	.340"	1.328"
Non-Insulated	GP1A, GP2A	.200"	.400"
	GP1B. GP2B	.200"	.656"
	GP2C	.265"	1.125"
	GP2D	.360"	1.110"
	GP2E	.360"	1.560"
	GP2S	.230"	.860"

**Ceramicon is the registered trade name of silvered ceramic condensers made by Erie Resistor Corporation.*

CAPACITY MMF	ERIE STYLE	TYPE* I=Insulated N=Non-Insulated	WORKING VOLTAGE	MAX. POWER FACTOR 1MC
10	GP1K	I	500	0.17%
	GP1A	N	500	0.17%
12	GP1K	I	500	0.16%
	GP1A	N	500	0.16%
15	GP1K	I	500	0.15%
	GP1A	N	500	0.15%
18	GP1K	I	500	0.14%
	GP1A	N	500	0.14%
20	GP1K	I	500	0.13%
	GP1A	N	500	0.13%
22	GP1K	I	500	0.12%
	GP1A	N	500	0.12%
24	GP1K	I	500	0.12%
	GP1A	N	500	0.12%
25	GP1K	I	500	0.12%
	GP1A	N	500	0.12%
27	GP1K	I	500	0.11%
	GP1A	N	500	0.11%
30	GP1K	I	500	0.1%
	GP1A	N	500	0.1%
33	GP1K	I	500	0.1%
	GP1A	N	500	0.1%
39	GP1K	I	500	0.1%
	GP1A	N	500	0.1%
47	GP1K	I	500	0.1%
	GP1A	N	500	0.1%
50	GP1K	I	500	0.1%
	GP1A	N	500	0.1%
51	GP1K	I	500	0.1%
	GP1A	N	500	0.1%
68	GP1K	I	500	0.1%
	GP1A	N	500	0.1%
75	GP1K	I	500	0.1%
	GP1A	N	500	0.1%
100	GP1K	I	500	0.1%
	GP1A	N	500	0.1%
	GP1L	I	500	0.1%
120	GP2K	I	500	3.0%
	GP2A	N	500	3.0%
150	GP2K	I	500	3.0%
	GP2A	N	500	3.0%
	GP1L	I	500	0.1%
	GP1B	N	500	0.1%
180	GP2K	I	500	3.0%
	GP2A	N	500	3.0%
200	GP2K	I	500	3.0%
	GP2A	N	500	3.0%
220	GP2K	I	500	3.0%
	GP2A	N	500	3.0%
240	GP2K	I	500	3.0%
	GP2A	N	500	3.0%
250	GP2K	I	500	3.0%
	GP2A	N	500	3.0%
270	GP2K	I	500	3.0%
	GP2A	N	500	3.0%
300	GP2K	I	500	3.0%
	GP2A	N	500	3.0%
330	GP2K	I	350	3.0%
	GP2A	N	350	3.0%
390	GP2K	I	350	3.0%
	GP2A	N	350	3.0%
470	GP2K	I	350	3.0%
	GP2A	N	350	3.0%
500	GP1M	I	500	0.1%
	GP2K	I	350	3.0%
	GP2A	N	350	3.0%
510	GP1M	I	500	0.1%
	GP2K	I	350	3.0%
	GP2A	N	350	3.0%
680	GP1M	I	500	0.1%
	GP2K	I	350	3.0%
	GP2A	N	350	3.0%
750	GP2L	I	350	3.0%
	GP2B	N	350	3.0%
1000	GP2L	I	350	3.0%
	GP2B	N	350	3.0%
1200	GP2L	I	350	3.0%
	GP2B	N	350	3.0%
1500	GP2L	I	350	3.0%
	GP2B	N	350	3.0%
1800	GP2M	I	350	3.0%
	GP2S	N	350	3.0%
2000	GP2M	I	350	3.0%
	GP2S	N	350	3.0%
2200	GP2M	I	350	3.0%
	GP2S	N	350	3.0%
2400	GP2M	I	350	3.0%
	GP2S	N	350	3.0%
2500	GP2M	I	350	3.0%
	GP2S	N	350	3.0%
2700	GP2M	I	350	3.0%
	GP2S	N	350	3.0%
3000	GP2M	I	350	3.0%
	GP2S	N	350	3.0%
3300	GP2M	I	350	3.0%
	GP2S	N	350	3.0%
4000	GP2M	I	350	3.0%
	GP2S	N	350	3.0%
4700	GP2M	I	350	3.0%
	GP2C	N	350	3.0%
5000	GP2M	I	350	3.0%
	GP2C	N	350	3.0%
5100	GP2M	I	350	3.0%
	GP2C	N	350	3.0%
6000	GP2D	N	350	3.0%
6800	GP2D	N	350	3.0%
7500	GP2E	N	350	3.0%
10000	GP2E	N	350	3.0%

Electronics Division

ERIE RESISTOR CORP., ERIE, PA.

LONDON, ENGLAND • • TORONTO, CANADA

MEMBERSHIP

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Brereton, C. H., RCA Victor Company, Ltd., 1001 Lenoir St., Montreal, Que., Canada
Brown, G. M., Rm. 634, 466 Lexington Ave., New York 17, N. Y.
Buss, R. R., Department of Electrical Engineering, Northwestern Technological Institute, Evanston, Ill.
Cox, R. T., 127—36 St. Dr., S.E., Cedar Rapids, Iowa
Dimmer, R. P., 1033 W. Van Buren St., Chicago 7, Ill.
Fiedler, G. J., 4389 Harvard Ave., La Mesa, Calif.
Greig, J., Department of Electrical Engineering, University of London, London, England
Hayes, E. A., Hughes Tool Company, 7000 Romaine St., Hollywood 38, Calif.
Hegbar, H. R., RFD 2, Princeton, N. J.
Kenefake, E. W., 525 Robineau Rd., Syracuse, N. Y.
McCoy, C. T., 428 Anthwyn Rd., Narberth, Pa.
Parsons, D. R., R. M. Electric Ltd., Kingsway, Team Valley Estate, Gateshead-on-Tyne, England
Rao, V. V. L., Government of Madras, Kilpauk P. O., Madras, South India
Richmond, L. P., 144 Forrer Blvd., Dayton 9, Ohio
Rudnick, P., RFD 1, Box 1-509, Princeton, N. J.
Sigmon, L. C., KMPC, 5939 Sunset Blvd., Los Angeles 28, Calif.
Smith, M. S., 2042 Linn Blvd., S.E., Cedar Rapids, Iowa
Smith, P. T., RCA Laboratories, Princeton, N. J.
Soderman, R. A., General Radio Company, 275 Massachusetts Ave., Cambridge 39, Mass.
Sturley, K. R., Broadcasting House, London W. 1, England
Wainwright, R. M., Montana-Dakota Utilities Company, Glendive, Mont.

Admission to Senior Member

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Downing, J. J., 532—20 St., N.W., Washington, D. C.
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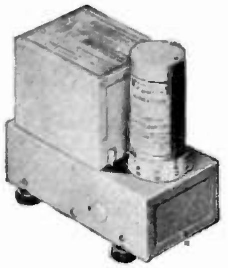
Hanson, F. E., Western Electric Company, Inc., 403 Hudson St., New York 14, N. Y.
Herr, D. L., Control Instrument Company, Inc., 67—35 St., Brooklyn 32, N. Y.
Macdonald, P. A., 221 Memorial Blvd., Winnipeg, Man., Canada
McIntosh, R. O., 263 Cascade Rd., Pittsburgh 21, Pa.
Murray, P. R., 135 Oakview Dr., Dayton 9, Ohio
Noble, A. E., Electronic Signal Engineering Company, 110 N. Franklin St., Chicago 6, Ill.
Romnes, H. I., American Telephone and Telegraph Company, 195 Broadway, New York 7, N. Y.
Sanial, A. J., 168-14—32 Ave., Flushing, L. I., N. Y.
Slattery, J. J., 1313 Third Ave., Spring Lake, N. J.
Taft, C. R., 59 Fairview Ave., West Orange, N. J.
Weitmann, O., 61 Cedar St., Cedar Grove, N. J.

Transfer to Membe.

Adkins, J. M., 135 Park Dr., Country Club Estates, Point Pleasant, W. Va.
Barnes, R., 12, Coledale Dr., Stanmore, Middlesex, England
Buckler, D. R., National Airlines, Inc., Municipal Airport, Jacksonville, Fla.
Chase, R. F., 2338—42 Ave., N., Seattle 2, Wash.
Choate, R. S., Group 22, RAAF, Darwin, N.T., Australia

(Continued on page 40A)

Save Space . . . Improve Efficiency with a MALLORY *Self-Rectifying* VIBRATOR



MALLORY VIBRAPACK*

Mallory Vibrapacks deliver voltages from 125 to 400 from low voltage sources . . . with high efficiency, low battery drain, ease of installation, long life.

**Reg. U. S. Pat. Off. for vibrator power supplies.*

If you're a manufacturer of automobile radios, mobile communication systems or of any other equipment using vibrator power supplies, now is the time to consider the Mallory *self-rectifying* vibrator.

Mallory self-rectifying vibrators are identical in size with the interrupter type. But—and here's the important point—they save space and wiring, and give you less components to worry about.

That's not the whole story either. You get top performance with Mallory self-rectifying vibrators. They assure high over-all power supply efficiency and eliminate the voltage drop within the rectifier tube plus the heater load.

Mallory engineers, who have produced more than 15,000,000 vibrators, invite your vibrator problems—will be glad to give you specific information about the self-rectifying type for your particular application. Write for copies of the comprehensive vibrator questionnaire.

More Mallory Vibrators are in use today than all other makes combined

P.R. MALLORY & CO. Inc
MALLORY VIBRATORS
AND VIBRATOR POWER SUPPLIES

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

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(Continued from page 38A)

Dausman, R. H., 3703 Jardin, Houston 5, Texas
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Dougherty, R. G., 158 Market St., Newark 2, N. J.
Dunn, R. F., 501 W. 135 St., New York 31, N. Y.
Foster, G. L., 161 High St., London, Ont., Canada
Franklin, R. A., 3367 Alabatross St., San Diego 3, Calif.
Harden, P. A., Suite 205, 529 E. Colorado St., Pasadena 1, Calif.
Fricke, R. E., Sylvania Electric Products, Inc., 70 Forsyth St., Boston 15, Mass.
Hodson, W. G., 524 Hampton Rd., Burbank, Calif.
Hoglund, R. H., Radio Research Laboratory, Cambridge 38, Mass.
Hudek, V. R., 905 G Ave., N.W., Cedar Rapids, Iowa
Ishler, H. K., 78 Clover Ave., Floral Park, N. Y.
Jacob, M. I., Apt. 9, Oaklee Village, Baltimore 29, Md.
Jones, O. J., Jr., 2808-A Yorkway, Baltimore 22, Md.
Kadak, E., 1801 Winans Ave., Halethorpe, Md.
Kerl, R. C., Pawnee City, Neb.
Kershner, R. H., 5521 Ashbourne Rd., Baltimore 27, Md.
Kriz, A., 7150 S. Mozart St., Chicago 29, Ill.
Lark, E. J., 811 Tenth St., Alexandria, Va.
Lebenbaum, M. T., Radio Research Laboratory, Divinity Ave., Harvard University, Cambridge, Mass.
Longman, E. R., 2230 Mount View Place, S.E. Washington 20, D. C.
Morey, A. M., 9525 N. Kellogg St., Portland 3, Ore.
Orner, R. J., 2643 Broadway Ave., Evanston, Ill.
Pearson, C. A., Naval Research Laboratory, North Beach, Md.
Persio, L. N., 244 W. Fourth St., Williamsport 1, Pa.
Rao, M. V. S., c/o Rev. G. Koppole, British Bible Society, Box 502, Madras, South India
Rennenkampf, L. J., 30-16—84 St., Jackson Heights L. I., N. Y.
Richey, F. R., 1235 E. Orange Grove Ave., Glendale 5, Calif.
Robertson, W. W., 1311 Ninth St., Wichita Falls, Texas
Rowe, R. G., Carborundum Company, Niagara Falls, N. Y.
Senn, P. J., 139 E. Court, Dayton 3, Ohio
Sigman, H. M., 6345 Sutherland, St. Louis 9, Mo.
Smith, J. W., 3734 C Ave., N.E., Cedar Rapids, Iowa
Swan, R. A., Jr., 4 Orne Sq., Salem, Mass.
Urquhart, M. S., Box 49, Merredin, Western Australia
Weiler, P. R., 6338 S. Lorel Ave., Chicago 38, Ill.
Younson, E. J., Bede Burn Rd., Jarrow Co., Durham, England
Zimmerman, R. F., 181 W. Hudson Ave., Dayton 5, Ohio

Admission to Member

Abel, H. L., c/o 157 Chesterton Rd., Cambridge England
Balton, I. A., 126-L Quentin St., Brooklyn 29, N. Y.
Blake, J. B., 10013 S. Manhattan Place, Los Angeles 44, Calif.
Briggs, R. F., Perry-Briggs Company, 1220 Huron Rd., Cleveland 15, Ohio
Charchian, B. M., 10 Mitchell Place, New York, N. Y.
Chernosky, A. A., 254 Humble Bldg., Houston, Texas
Cripe, J. H., 628 N. Brandywine Ave., Schenectady 8, N. Y.
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Erf, H. A., 17355 S. Woodland Rd., Shaker Heights, Cleveland 20, Ohio

(Continued on page 42A)

new source of wave shapes...

Model SE-512

SHERRON *Multi - wave shape Generator*

This latest Sherron development incorporates into a single instrument the source of several wave shapes. Because it supplies these fundamental wave shapes, it serves in testing amplifiers and related equipment at audio and video frequencies. This advancement is a further demonstration of the electronics know-how which has made the Sherron name a reliable guide to dependable laboratory control.

SPECIFICATIONS

OUTPUTS: *Sine waves, square waves, positive pulses, negative pulses and a trigger pulse. The impedance (output) is 250 ohms for all voltages.*

FREQUENCY RANGE: *50 cycles to 50,000 cycles for all voltages continuously variable with a direct reading dial.*

SQUARE WAVE: *Rise time is less than 3 of a micro-second at the highest frequency and about .7 of a micro-second at the lowest frequency.*

PULSES: *Pulse width of both positive and negative outputs is variable from about 1 to 75 micro-seconds.*

POWER REQUIREMENTS: *115 volts, 60 cycle, 300 watts.*

Sherron Electronics
W2XQK

SHERRON ELECTRONICS COMPANY
Division of Sherron Metallic Corporation
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West Coast Sales Office: Mechanics Institute Building • 57 Post Street • San Francisco, Calif.

MEMBERSHIP

(Continued from page 40A)

Esteve, J. A., Box 986, Havana, Cuba
Flatow, T., 1404 Michigan, Houston 6, Texas
Fleming, L., 510 N. West St., Falls Church, Va.
Frankenstein, G., 36 Seminary Ave., Dayton 3, Ohio
Gamble, E. H., 160 E. 48 St., New York 17, N. Y.
Hall, W. H., 804 Madison St., Syracuse 10, N. Y.
Heston, W. S., 121 Hallock St., Jamestown, N. Y.
Huebner, D. L., Nut Swamp Rd., RFD 1, Box 2, Red Bank, N. J.
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Jerrett, A. C., 196 Water St., St. John's Newfoundland
Kovski, J. J., 1802 Wayne Ave., Dayton 10, Ohio
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Livenick, C. E., 7553 S. Hamilton, Chicago, Ill.
Lynch, D. W., 142 Ivanhoe St., S.W., Washington 20, D. C.
McCharles, J. R., 1561 Euclid Ave., Berkeley, Calif.
McRae, F. B., 1106 W. 87 St., Los Angeles 44, Calif.
McReynolds, G. F., 3106 Plumb St., Houston 5, Texas
Michalowicz, J. C., One Noyes Ct., Silver Spring. Md.
Morgan, G. M., 1205 W. Sherwin Ave., Chicago. Ill.
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Nelson, V. G., 27 Schneider Dr., Osborn, Ohio
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Reed, D. M., 823 N. Eighth St., Manitowoc, Wis.
Reichman, B. J., 88 Rugby Rd., Brighton, Sussex, England
Rennaker, R. B., 722 N. Lombard Ave., Oak Park. Ill.
Ryesky, S., Mountain Side Colony, Dillsburg, Pa.
Saviers, G. B., 713 S. Dorchester Rd., Baltimore 29, Md.
Segal. R. J., National Bureau of Standards, Radio Division, Washington 25, D. C.
Simmons, C. A., RFD 5, Box 159-D, Shreveport, La.
Smith, W. R., 315 Bloor St., W., Toronto 5, Ont., Canada
Solley, B. J., 95 Station Rd., Hendon, N.W., 4, England
Shen, T. C., c/o Mr. A. J. Mundt, Personnel Director, Engineering Department, Western Union, 60 Hudson St., New York, N. Y.
Stern, H., 700 Washington St., Cumberland, Md.
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Taylor, W. R. C., 95 Niagara St., Winnipeg, Man., Canada
Thorp, R., "Comox," Westbourne Rd., Hornsea, Yorkshire, England
Todd, F. J., 1807 N. Nagle Ave., Chicago 35, Ill.
Todd, M. C., Wendell, N. C.
Trueblood, R. O., 505-A S. Tenth St., Laramie, Wyo.
Utz, R. H., RFD 1, Sterling, Va.
Venaglia, E. J., 27 Clarendon Rd., East Hempstead, N. Y.
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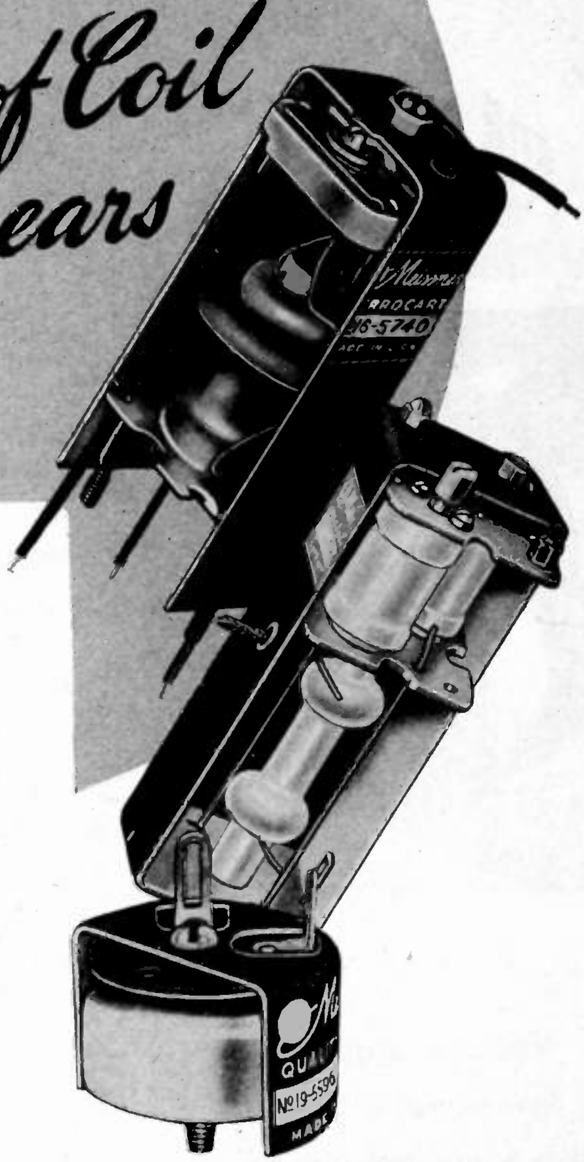
Watson, R. F., 849 E. 17 St., Brooklyn 30, N. Y.
Weil, F. M., 629 Raleigh St., Glendale 5, Calif.
Werner, G. L., 80 Clarkson Ave., Brooklyn 26, N. Y.
Whitacre, D. P., Sr., 2104 W. Commonwealth St., Alhambra, Calif.
Wiewara, E. J., 4021 Ninth St., N.W., Washington 11, D. C.
Willis, A. H., 3628 Hilmar Rd., Baltimore 7, Md.

(Continued on page 44A)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№ 19-5596
MADE

MEMBERSHIP

(Continued from page 42A)

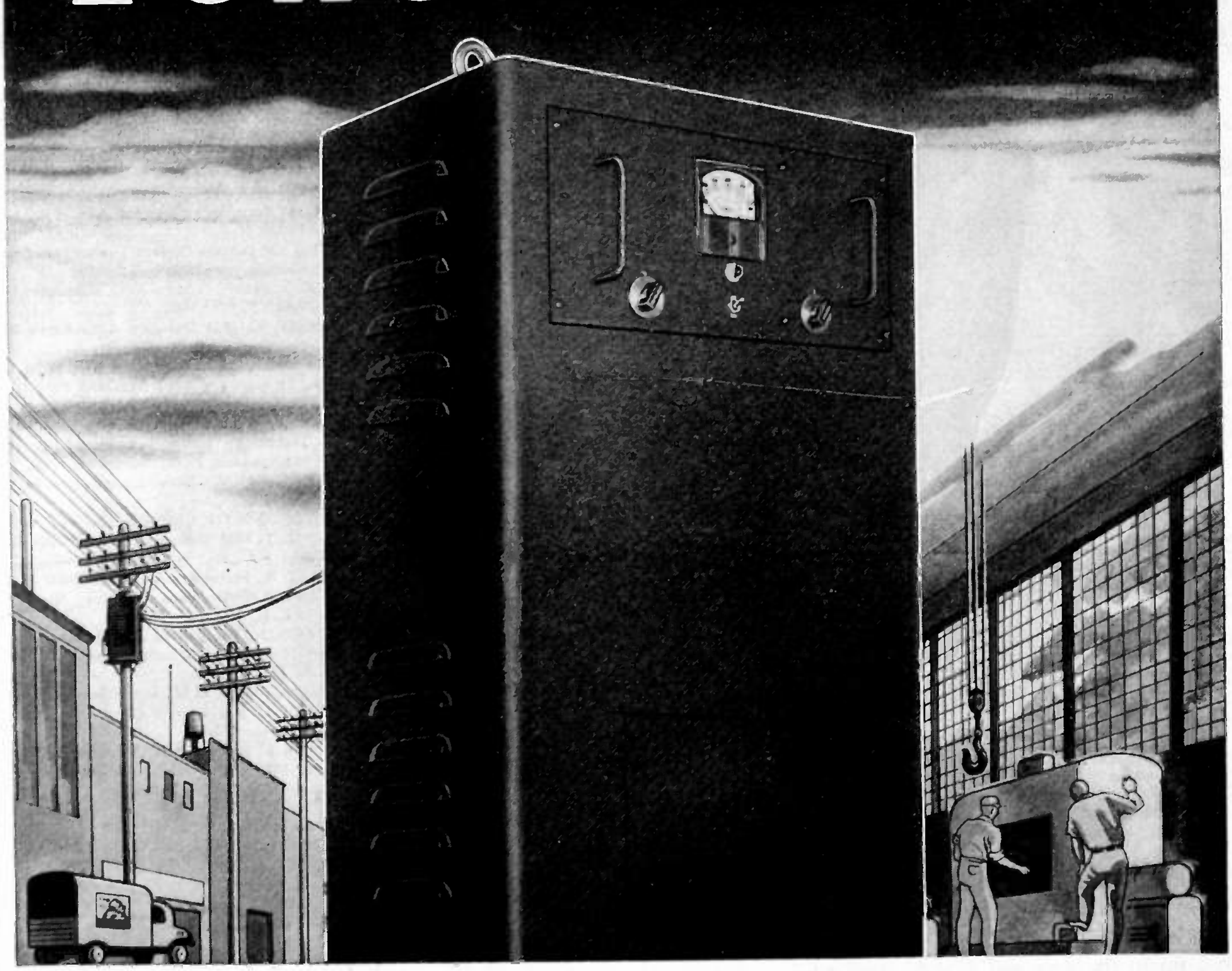
Wintle, H. C., 710 Matty Ave., East Syracuse, N. Y.
Wolff, R. L., 4424 N. Cramer St., Milwaukee 11, Wis.
Wormcke, W., 110 Rutgers Ave., Jersey City 5, N. J.
Wightman, R. R., 3109E Berkeley Dr., Philadelphia 29, Pa.
Young, W. G., 6350 McCommas, Dallas 14, Texas

Admission to Associate

Angevine, R. A., 143-19—38 Ave., Apt. 1-D, Flushing, L. I., N. Y.
Bahr, R., Jr., 88-15—73 Ave., Glendale, Brooklyn 27, N. Y.
Bath, C. C., Bendix Radio, Department 75, Towson 4, Md.
Behrens, H. G., 92-23—212 St., Queens Village 8, L. I., N. Y.
Bennett, R. M., 432 Nelson St., Ottawa, Ont., Canada
Berger, L., 1426 Franklin Ave., New York 56, N. Y.
Bergman, S., 6169 Liebig Ave., New York 63, N. Y.
Beser, J., Box 5100, Albuquerque, N. M.
Bliss, D. C., 142 W. Central Blvd., Palisades Park, N. J.
Bradford, L. C., 3 Castlewood Dr., Shooters Hill, London S.E. 9, England
Bradley, C. E., 129 Gale Dr. N., Beverly Hills, Calif.
Brunette, G. E., WLMDA, Watson Laboratories, Red Bank, N. J.
Carland, E., Apt. 325, 1630 R St., N.W., Washington 9, D. C.
Chadek, T. J., 3300 India St., San Diego 1, Calif.
Chandler, A. M., RFD 2, Tippecanoe, Ohio
Cheifetz, R. M., 432 Christopher Ave., Brooklyn 12, N. Y.
Clark, J. C., Jr., 628—60 St., Oakland 9, Calif.
Cohen, L. R., 115 Clarke St., Syracuse 10, N. Y.
Cope, G. W., 1817 P St., S.E., Washington 20, D. C.
Costanzo, P. A., 1933 Stanley Ave., Niagara Falls, Ont., Canada
Darne, F. R., 1420 Tuckerman St., Washington 11, D. C.
Davey, J. R., Bell Telephone Laboratories, 180 Varick St., New York 14, N. Y.
Delaney, J. N., "Carmody," 8 Kent Rd., Rose Bay, Sydney, N.S.W., Australia
Dewar, W., 2 Third St., North Arlington, N. J.
Dover, J., 5973 N. Clark St., Chicago, Ill.
Ebner, W. H., 3616 N.W. 12 St., Oklahoma City 7, Okla.
Edson, M., 409 Manor Rd., E., Toronto, Ont., Canada
Eskin, H. M., 183 Sellinger St., Rochester 5, N. Y.
Ehrhardt, R., 1920 Bailey Ave., Buffalo 11, N. Y.
Fickett, M. M., 2923 Bell Blvd., Bayside, N. Y.
Finke, R. S., 114-06 Queens Blvd., Forest Hills, L. I., N. Y.
Finlay, G. E., Chicago Laboratories, Section 8-1, Montgomery Ward and Company, 618 W. Chicago Ave., Chicago 7, Ill.
Finn, P. L., 229 W. 97 St., New York 25, N. Y.
Fontanellaz, G., Hallerstrasse 49, Bern, Switzerland
Foran, F. M., Box 434, Salamanca, N. Y.
Fox, W. J., 481 Laurier, W., Ottawa, Ont., Canada
Fullerton, R. D., 329 E. Jefferson St., Fort Wayne 2, Ind.
Fye, D. L., 141 Ivanhoe St., S.W., Washington 20, D. C.
Gerrand, J. H., 15 McLeod Ave., Roseville, N.S.W., Australia
Greenwood, D. T., 2149 Sanford Ave, St. Paul 5, Minn.
Hanselman, L. C., 83 Grant St., East Rockaway, L. I., N. Y.
Hartwigsen, C. K., 718 San Carlos Ave., Albany 6, Calif.

(Continued on page 46A)

Power Behaves



...SECO AUTOMATIC VOLTAGE REGULATORS

CONSTANT VOLTAGE — a necessity in the operation of electrical apparatus — can be best obtained by SECO Automatic Voltage Regulators.

Installed in a convenient corner in the plant, a SECO regulator will maintain constant voltage to entire laboratories, machine rooms, and other factory equipment sensitive to line voltage variations.

A SECO regulator offers for large industrial power users the same voltage control generally found only in instruments for laboratory use. Some inherent characteristics are:

- Rapid correction of line voltage variations.
- Zero wave-form distortion.
- Not affected by changes in the power factor or magnitude of the load.
- Does not affect power factor of the system.
- No critical mechanical adjustments.

Incidentally, the distinctive design of SECO Automatic Voltage Regulators adds much to the appearance of any plant interior. Investigate now!

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Superior SECO Electric
Company

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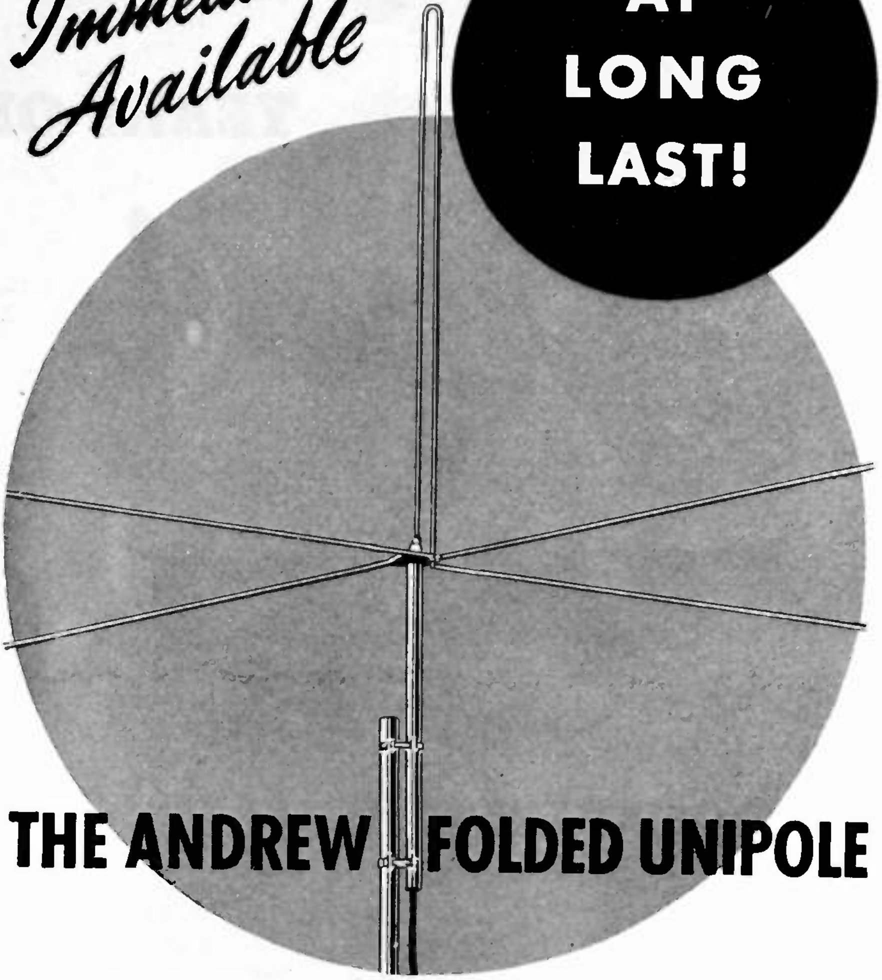
BRISTOL, CONNECTICUT, U. S. A.

MEMBERSHIP

(Continued from page 44A)

Heritage, D. P., 112 Irvington St., S.W., Washington 20, D. C.
Hites, C. E., 2308 Troy St., Dayton 3, Ohio
Hoehne, V. M., 1203 S. 121 St., West Allis 14, Wis.
Hoffmann, H. T., 1 Rue Saint-James, Neuilly s/Seine, France
Horton, W. A., Jr., 1368 Fry Ave., Lakewood, Ohio
Hoyles, H. A., 89 Grenadier Rd., Toronto, Ont., Canada
Israel, L. J., Turnpike at Sherman Ave., Pompton Plains, N. J.
Julien, G., 109 Jacques-Cartier, Donnacona, Que. Canada
Kapke, K. O., 2425 N. Oakland Ave., Milwaukee 11, Wis.
Karpeles, M. A., 1953 Davidson Ave., Bronx 53, N. Y.
Kiger, L. E., 281 Carlton Ave., Brooklyn 5, N. Y.
Kitai, R., 19 Streatley Ave., Auckland Park, Johannesburg, Transvaal, South Africa
Krishnaswamy, K. R. I., Department of Electrical, Technology, Indian Institute of Science, Malleswaram P. O., Bangalore, India
Malone, P. P., 937 N. Columbian Ave., Oak Park, Ill.
Martin, E. E., Box 279, Glen Head, N. Y.
Mattingly, L. J., 3657 N. Bosworth St., Chicago 13, Ill.
Meinheit, C. E., Sylvania Electric Products, Inc., 208-20 Cross Island Parkway, Bayside, L. I., N. Y.
Mereen, D. E., 1017 S. Sixth St., Milwaukee 4, Wis.
Messinger, H. P., 5125 S. Dorchester Ave., Chicago 15, Ill.
Myers, K. M., 2344 E. 41 St., Des Moines 17, Iowa
Mobley, M. P., Jr., 4610 Laurel Grove Ave., North Hollywood, Calif.
Montalvo, E., 5 Cyrus Place, San Francisco 9, Calif.
Pace, E. C., 111 E. Fifth St., Odessa, Texas
Pales, D., Box 804, Ponce, Puerto Rico
Paraska, T. F., 1361 Meridian Pl., N.W., Washington 10, D. C.
Perley, A. D., 1301 Bryn Mawr Dr., Dayton 7, Ohio
Pouliot, J. A., 54 Langevin, Apt. 2, Hull, Que., Canada
Ramakrishtna, V., "Vepa House," Main Rd., Vizagahatam, India
Resler, C., 309 W. 99 St., New York 25, N. Y.
Riesel, H., 9, George Elliot St., Tel-Aviv, Palestine
Ritz, E. A., 57 Heins Ave., Kitchener, Ont., Canada
Roderick, E. R., 323—37 St., N.W., Canton 7, Ohio
Rodriguez, R. R., 421 N. Ditman, Los Angeles, Calif.
Sander, C. M. R., Box 92, Bremerton, Wash.
Scheiner, M. L., 51 E. 78 St., New York 21, N. Y.
Schwarz, E. F., 512 Roslyn Ave., C. H., Petersburg, Va.
Sheets, C. C., 1410 Girard St., N.W., Apt. 2, Washington 9, D. C.
Speidel, F. X., 5468 Maple, St. Louis 12, Mo.
Smith, D. L., 208 Daly Ave., Ottawa, Ont., Canada
Smith, G. F., 1902 W. Minnehaha Ave., St. Paul 4, Minn.
Teich, P. J., 153 Mosley Dr., Syracuse 6, N. Y.
Thomas, A., 161 S. Mitton St., Sarnia, Ont., Canada
Thompson, J. L., Jr., United States Merchant Marine Academy, Kings Point, N. Y.
Tosch, R. L., Federal Communications Commission, Rm. 1215, New Post Office Bldg., 12 St. and Pennsylvania Ave., N.W., Washington 25, D. C.
Tufts, R. M., 617 E. Maumee, Angola, Ind.
Vandal, W. L., Jr., 5229—28 Ave., S., Minneapolis 6, Minn.
van Olst, M. D. A., 345 Dey St., New Muckeeneuk, Pretoria, South Africa
Vlok, T. C. B., 36 High St., Berea, Johannesburg, South Africa
Ward, R. O., 730 Rose, Burbank, Calif.

(Continued on page 48A)



MEMBERSHIP

(Continued from page 46A)

Watts, R. J., Rm. 105, War Research Bldg., Ohio State University, Columbus 10, Ohio
Weintraub, W. S., 1975 Beachwood Dr., Hollywood 28, Calif.
White, W. H., 1454 S. Gordon St., S.W., Atlanta, Ga.
Whiting, R. J., 3505 A.A.F.B.U., Scott Field, Ill.
Wilcomb, E. F., 1533 Elm St., Youngstown, Ohio
Winick, A. B., 47 Forrester St., S.W., Washington 20, D. C.
Winters, D. E., 6115 Weber Rd., Affton 23, Mo.
Wong, W. E., 823 Grant Ave., Apt. 1, San Francisco 8, Calif.
Wu, H. S., Department 110, RCA Victor Division, Lancaster, Pa.
Yuran, T., 177 Northern Pkwy., Hempstead, L. I., N. Y.
Zeller, H. R., 1434 N. Bosworth Ave., Chicago 22, 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.

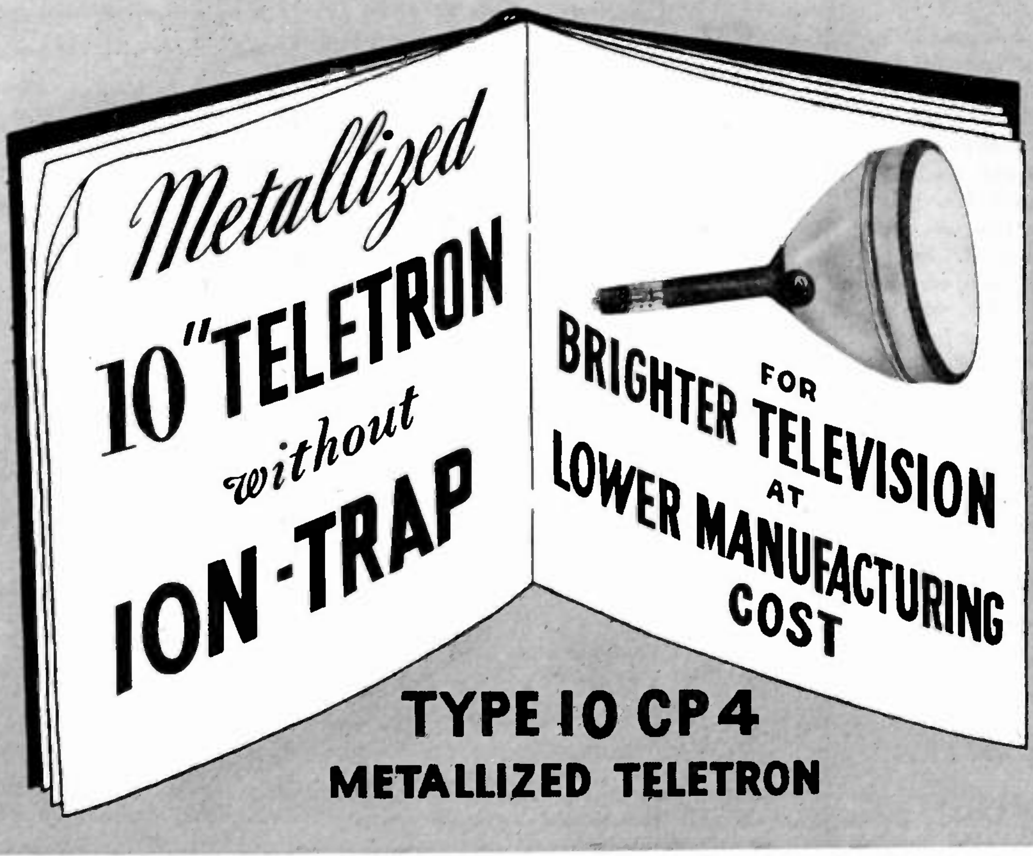
(Continued from page 8A)

Radioactivity Probe

A portable, battery-operated, beta-gamma count rate meter, for general radioactivity surveys, has been developed by Instrument Development Laboratories, 817 E. 55 St., Chicago 15, Ill. The new model 2610 meter has three ranges—0.2, 2, and 20 milliroentgens per hour full scale—which gives the instrument a range below the cosmic ray background and above the tolerance level.

The meter has a detachable probe with 4-foot cable for holding the Geiger-Mueller tube. Locations, not easily accessible to larger instruments, are readily explored with this probe. To distinguish between beta and gamma rays, the window of the probe has an adjustable shield which can be set to prevent beta particles from affecting the Geiger-Mueller tube. A headset is furnished for rapid surveying operations.

(Continued on page 56A)


Metallized
10" TELETRON
without
ION-TRAP
FOR
BRIGHTER TELEVISION
AT
LOWER MANUFACTURING
COST

BELMONT RADIO CORPORATION of Chicago

NEEDS SEVERAL ELECTRONIC ENGINEERS

●

Development and Research Engineers, Seniors and Juniors, well versed in all phases of RF circuits wanted. VHF or Microwave experience desirable. Top-notch applicants with Engineering degree or equivalent background for this type of work only will be considered. Chicago Area residents preferred, top salary, steady position, 40-hour week, occasional field trips. Appointment by letter only. Give background experience, educational and employment history. Address letter to Research Division, Belmont Radio Corporation, 5921 W. Dickens Ave., Chicago 39, Illinois.

WANTED

Electronic Engineer

By development laboratory of fifty-year-old manufacturing concern located in western New York. Graduate with several years experience in radar, navigation, fire control, servo or similar systems. Must have extensive knowledge of circuit fundamentals and be able to design circuit and apparatus for implementing broad new systems and to carry the work through prototype and pilot model stages.

Box No. 448.

The Institute of Radio Engineers
1 East 79 Street, New York 21, N.Y.

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . .

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.
1 East 79th St., New York 21, N.Y.

ELECTRICAL AND ELECTRONIC ENGINEERS

You are invited to investigate opportunities in electronics and electrical engineering at this laboratory. We are organized as a department in a large college located in a Middle Atlantic state, and are engaged in research and development. Members of our research staff are appointed with academic rank. Some are teaching, although this is not a condition of employment. Box 449.

TELEVISION ENGINEER

Wanted by well-known manufacturer of television radio receivers. To be service manager for New York area. Write fully, giving experience and salary requirements. Box 450.

INSTRUCTORS—ASSISTANT PROFESSORS

The electrical engineering department of a large southwestern college has staff openings in grades of instructor and assistant professor in communications and electronics. For details, write to Box 451.

MATHEMATICIAN

With knowledge of electrical and mechanical means of solving equations, to work on computer development with New York City concern. Additional knowledge of servo systems desirable. Box 452.

CIVIL ENGINEERS

The College of Applied Science, Syracuse University, is seeking candidates for two positions in its Department of Civil Engineering. It is desired that each applicant have the following qualifications:

1. At least a Master's degree in Civil Engineering.
2. Some experience as a teacher in the field of civil engineering, including courses such as mathematics, engineering mechanics, strength of materials, and hydraulics.
3. Two or more years' experience with civil engineering organizations such as consulting firms, Public Works departments, etc., with some degree of responsible charge.
4. Age range preferred, 28 to 40 years.

Address correspondence to: Donald E. Stearns, Professor of Civil Engineering, College of Applied Science, Syracuse University, Syracuse 10, N.Y.

PHYSICIST

With M.S. or Ph.D. degree, and training in both nuclear physics and electronics. Write in detail, stating qualifications, to Personnel Office, RCA Laboratories, Princeton, N.J.

WANTED PHYSICISTS ENGINEERS

Engineering laboratory of precision instrument manufacturer has interesting opportunities for graduate engineers with research, design and/or development experience on radio communications systems, electronic & mechanical aeronautical navigation instruments and ultra-high frequency & microwave technique.

●

WRITE TO BOX 1278
EQUITY ADVERTISING AGENCY
113 W. 42nd St., New York 18, N.Y.

INSTRUCTORS

To teach radio engineering. Must have B.S. in E.E. or Physics, or equivalent; 5 years experience in commercial electronics work. Knowledge of microwaves desirable. Write, giving full details regarding education and experience, to Personnel Department, Spartan School of Aeronautics, Tulsa, Oklahoma.

PHYSICISTS AND ELECTRICAL ENGINEERS

Unusual opportunities in recently organized research group for work in broad fields of radio, radar, telemetering, servomechanisms, and gyroscopes. Excellent openings for experienced physicists with good training in fundamentals. Please reply, giving experience and education, including transcript of college record, to: Engineering Personnel Office, North American Aviation, Inc., Municipal Airport, Los Angeles 45, California.

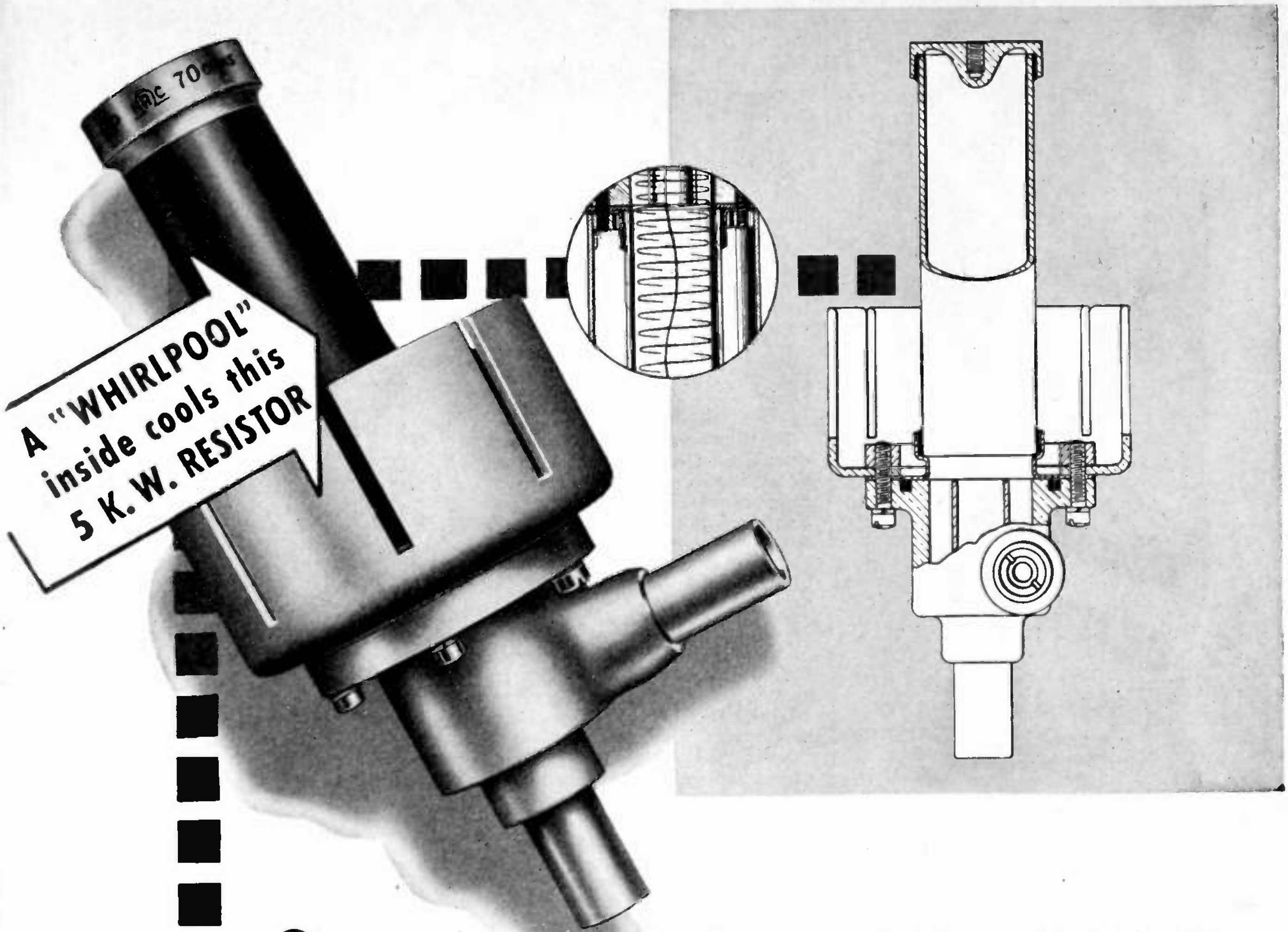
TRANSLATORS—PART-TIME

Foreign language articles, papers, patents, etc. Qualified readers state languages, technical subjects. Accurate Translation Service, 711 Woodward Building, Washington 5, D.C.

ENGINEERS—PHYSICISTS

Prominent communications development organization in Metropolitan area has openings for men experienced in microwave systems development; ultra-high-frequency and very-high-frequency transmission, receiver and antenna circuits; and electromechanical design. Write details. Box 444.

(Continued on page 52A)



Unique IRC High Frequency-High Power Resistor for Television, FM, and Dielectric Heating Applications

Inside IRC's new Type LP resistor a high velocity stream of water flows in a spiral path against the *metallized* resistance film and, through centrifugal force, maintains intimate thermal contact with the entire surface. Interchangeable intake nozzles permit adjusting the rate of water flow and therefore the cooling action to suit local water pressure and power dissipation up to 5 K.W.

A resistance film less than 0.001" thick, with an active length considerably less than ¼ wave length at FM and television frequencies, gives good inherent frequency characteristics.

The mechanical design permits direct mounting on the end of a coaxial line with both water intake and outlet connections at R.F. ground potential. Resistor elements are interchangeable. Different values or service replacements can be readily installed in the field.

The IRC Type LP Liquid Cooled High Frequency High Power Unit is the latest in IRC's continuing development of resistors. It is available in resistance values of 35 ohms to 1500 ohms. Resistance tolerance: ±15% standard. Tolerances of ±10% and ±5% can be supplied at increased cost.

PREFERRED FOR PERFORMANCE
IRC
FIXED & VARIABLE RESISTORS

SHORT DELIVERY CYCLE

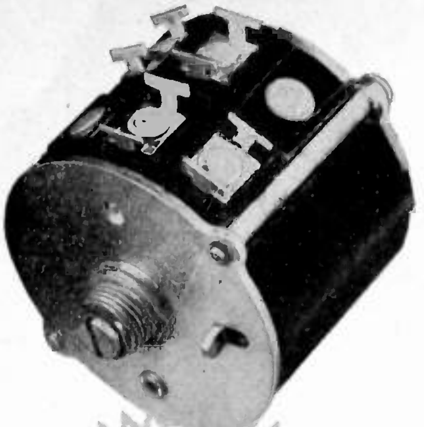
Fittings are now carried in stock. Resistance elements are made to order for range and tolerance.

For specific engineering information contact your IRC Sales Engineer or write Dept. 2A.

INTERNATIONAL RESISTANCE CO.

401 N. BROAD ST., PHILADELPHIA 8, PA.

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



POSITIONS OPEN

(Continued from page 50A)

PATENT ENGINEER

Good pay while you learn. An unusual opportunity for a young electrical engineer or physicist to receive intensive training as a patent solicitor for a substantial Texas corporation. Give details of experience, education, draft status, salary expected. Box 445.

PHYSICIST-METALLURGIST

For fundamental development and research in metals used in electron tubes. Work requires a high degree of theoretical background as well as practical experience. Location near Metropolitan area with ample facilities available for extensive research. State complete résumé of experience and background. Box 440.

ENGINEERS

Prominent aircraft company, Eastern United States, needs men with following qualifications: Experienced engineers with Bachelor's, Master's, Doctor's degrees in electrical engineering, physics, mathematics. At least two years experience required in design and development of radar and television systems, automatic computers, servo-mechanisms, target seekers, etc. Positions open for preliminary and detail design, research, and development of guided missiles under Army and Navy contracts. Starting salary commensurate with experience. Address Box 441.

COIL DEPARTMENT HEAD

Experienced in set-ups, winding, impregnation and testing home-receiver type radio-frequency and intermediate-frequency coils and chokes, wanted by television and radio manufacturer in New York area. Give experience and salary expected. Box 442.

RADIO ENGINEERS

Air King has openings for graduate engineers with senior experience on amplitude-modulation and frequency-modulation broadcast receiver design. Applicants must be schooled in measuring techniques and capable of establishing competent specifications for Vendors' guidance on a quantitative basis. Right men should be sufficiently capable to work on their own. Write or wire, giving full particulars of experience, salary desired, etc., to Mr. Frank A. Hinners, Vice-President in Charge of Engineering, Air King Radio Co., 1523-63rd St., Brooklyn 19, N.Y.

ENGINEERS (SENIORS)

ELECTRICAL. Several positions open for men with ultra-high-frequency experience.

MECHANICAL. Several positions open for men with radar design experience. Allen B. DuMont Laboratories, Inc., 2 Main Avenue, Passaic, N.J.

ENGINEER—EXECUTIVE

We need a top-flight executive to direct the engineering efforts of our young, expanding electronic manufacturing organization. Box 447.

1000 Watts of FM Broadcast Satisfaction

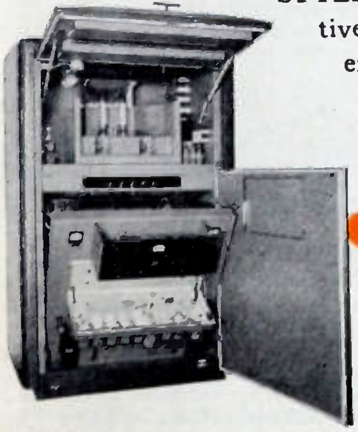
Collins FM transmitters are fully engineered in every detail. They reflect many years of successful experience in designing and manufacturing broadcast transmitters unexcelled in performance and reliability. Persons who attended the NAB convention in Chicago were noticeably impressed with the 732A on exhibition there.

The Collins 732A 1kw
FM Transmitter

What they saw:

RELIABILITY: They saw thorough design in every part of the equipment—Oversized components in all circuits—Personnel protection by means of electrical and mechanical interlocks—Overload protection—Proper ventilation. This transmitter is as substantial as it looks. Our engineers have the experience and know-how to design long and trouble-free life into radio equipment.

STYLE: The modern yet conservative exterior, with its three-tone gray finish, is attractive today and will be ten years from now. The beauty of Collins FM transmitters extends throughout the equipment. Chassis layout is symmetrical, roomy, and functional. Vertical construction and hinged chassis design provide utmost accessibility.



What they didn't see:

PERFORMANCE: They couldn't see the performance characteristics as measured in actual operation. Measurements show a carrier stability within ± 200 cps. Distortion is less than 1.0%. The frequency response is within 1.0 db total variation from 50 to 15000 cps. The noise level is at least 65 db below 100% modulation.

Write for an illustrated bulletin. And remember that we can supply your entire equipment requirements.

FOR BROADCAST QUALITY, IT'S . . . COLLINS

COLLINS RADIO COMPANY, CEDAR RAPIDS, IOWA

11 West 42nd Street, New York 18, N. Y.

458 South Spring Street, Los Angeles 13, California

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 within a period of one year. 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.

ELECTRONICS ENGINEER

B.E.E. 1944, Northeastern, Tufts. Age, 24; married. Annapolis Reserve Midshipmen School, Navy officer radar course at Bowdoin and M.I.T., sea duty on destroyer. Employed nearly one year by Army Air Forces as electronic engineer. Member Tau Beta Pi. Box 62W.

ENGINEER

B.S.E.E. 1938. Now studying management engineering. 3 years Navy electronic officer and radar instructor. 3 years X-ray field, 1 year electronic design. Desires position in decent climate at sufficient salary to properly raise a family. Prefers managerial duties. Box 63W.

JUNIOR ENGINEER

B.E.E. 1945. Age, 21; single. Six months experience with Army Air Forces on very-high-frequency transceiver design at Wright Field, Ohio. Interested in communications research (very-high-frequency or all other frequency ranges). Available December, 1946. Location N.Y., N.J., Conn. New York City preferable. Box 64W.

ENGINEER

B.S.E.E. University of Pittsburgh. Age, 25; married. Officer Signal Corps, installing transmitters, RTTY, and navigational aids for the Air Forces. Two years experience in electronics research laboratory. Desires position near Pittsburgh, Pa. Box 65W.

RADIO ENGINEER

B.S.E.E. Purdue, 1943. Two years civilian experience in development and production of radio equipment; also, one year with Signal Corps Engineering Laboratories. Desires development work on receivers or cathode-ray equipment with Midwestern concern. Available January. Box 66W.

ELECTRICAL ENGINEER

B.E.E. 1945. College of the City of New York. Age, 21; single. Worked several months in radio development. Desires development work in radio or electronics in vicinity of New York City. Box 67W.

ELECTRONIC ENGINEER

B.S. in E.E., 1937. Experience in communications-receiver design, and in design, testing, and production of ultra-high-frequency transmitters and receivers. Ex-Marine Corps radar officer. Desires design or research position, preferably on West Coast. C. B. Barnes, 4448 Union St., La Canada, Calif.

ENGINEER

B.E.E. by February, 1947. Senior work in E.E. completed. Requires two non-electrical subjects for degree now taking in evening school. Age, 27. Desires start in electronics field. Army meteorologist. Box 47W.

INDUSTRIAL ELECTRONICS

B.S.E.E. Wisconsin, some graduate work at Northwestern. Eta Kappa Nu and Tau Beta Pi. 2½ years broadcast station control engineer; 1 year electronics instructor; 2½ years development and production airborne radar equipment; 1½ years in service with Office of Scientific Research and Development new-developments division. Desires position involving design and development of industrial electronics applications. Will consider foreign service. Box 48W.

COMMUNICATIONS ENGINEER

B.S. Engineering, Physics, Toronto. Age, 26. Three years naval officer in charge of operating and maintaining all radar equipment aboard cruiser. One year field engineering commercial frequency-modulation radio. Seeks opportunity for original work in electronics, U.S. or Canada. Box 49W.

ENGINEER

B.S.E.E. Age, 33; married. M.I.T.-Harvard trained electronics-radar officer, Signal Corps, desires position of sales engineer or development of radar and allied fields. Sales, administrative, and manufacturing experience. Box 50W.

ELECTRICAL ENGINEER

B.E.E. Industrial experience in electrical test planning. Radio and radar experience in Army Signal Corps. Interested in radio and electronics. Age, 23. Will work within 300 miles of New York City. Résumé upon request. Box 51W.

ENGINEER

B.S.E.E. Purdue University and Massachusetts Institute of Technology trained electronics officer. Age, 28; married. Experience with naval radar, sonar, and loran while serving three years afloat. Civilian experience in transformer design and with utility. Desires permanent position in Midwest. Box 52W.

RADIO TECHNICIAN

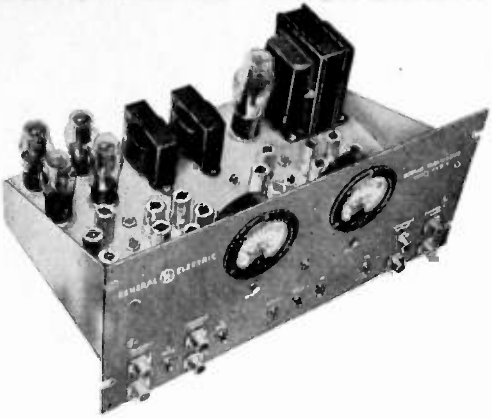
Signal Corps Officer. Age, 35. Graduate RCA, Press Wireless, and A.T.&T. courses. Five years extensive installation, operation, and maintenance of high-power stations. First-class phone license. Will travel. M. G. Gerstein, 1304 Grant Ave., New York 56, N.Y.

ELECTRONICS TECHNICIAN

Three years teaching, total experience, 13 years. Desires opportunity in research laboratory, teaching, frequency-modulation or television radio station work, theater sound maintenance, or commercial sound work. Box 53W.

(Continued on page 56A)


FOR A CONTROLLED BROADCASTING PATTERN



Positions Wanted

(Continued from page 54A)

SALES ENGINEER

Seventeen years experience in all phases broadcast engineering. Four years Army Air Forces as Commanding Officer communications group. Extensive experience aeronautical radio and navigational aids. Age, 36. Desires permanent responsible position with progressive company. Will consider foreign assignment. Box 34W.

ENGINEER

U.S.N.A. B.S., Harvard M.S. radio engineering. Commander U.S.N. 13 years naval electrical, engineering, and command experience. Age, 43. Desires executive or administrative position. Boston, Mass. Available October, 1946. Box 36W.

ENGINEER

B.S. in E.E. Vermont, 1941. Age, 29. Radar training M.I.T. 4 years naval officer, specializing in maintenance and installation of radar and radio equipment aboard aircraft. Interested in research and development. Box 38W.

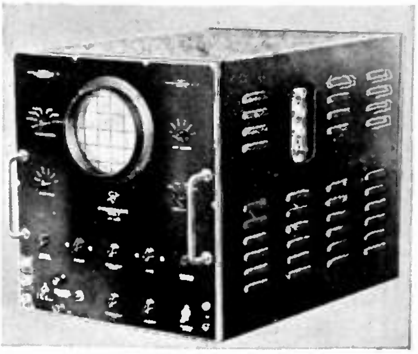
News—New Products

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

(Continued from page 48A)

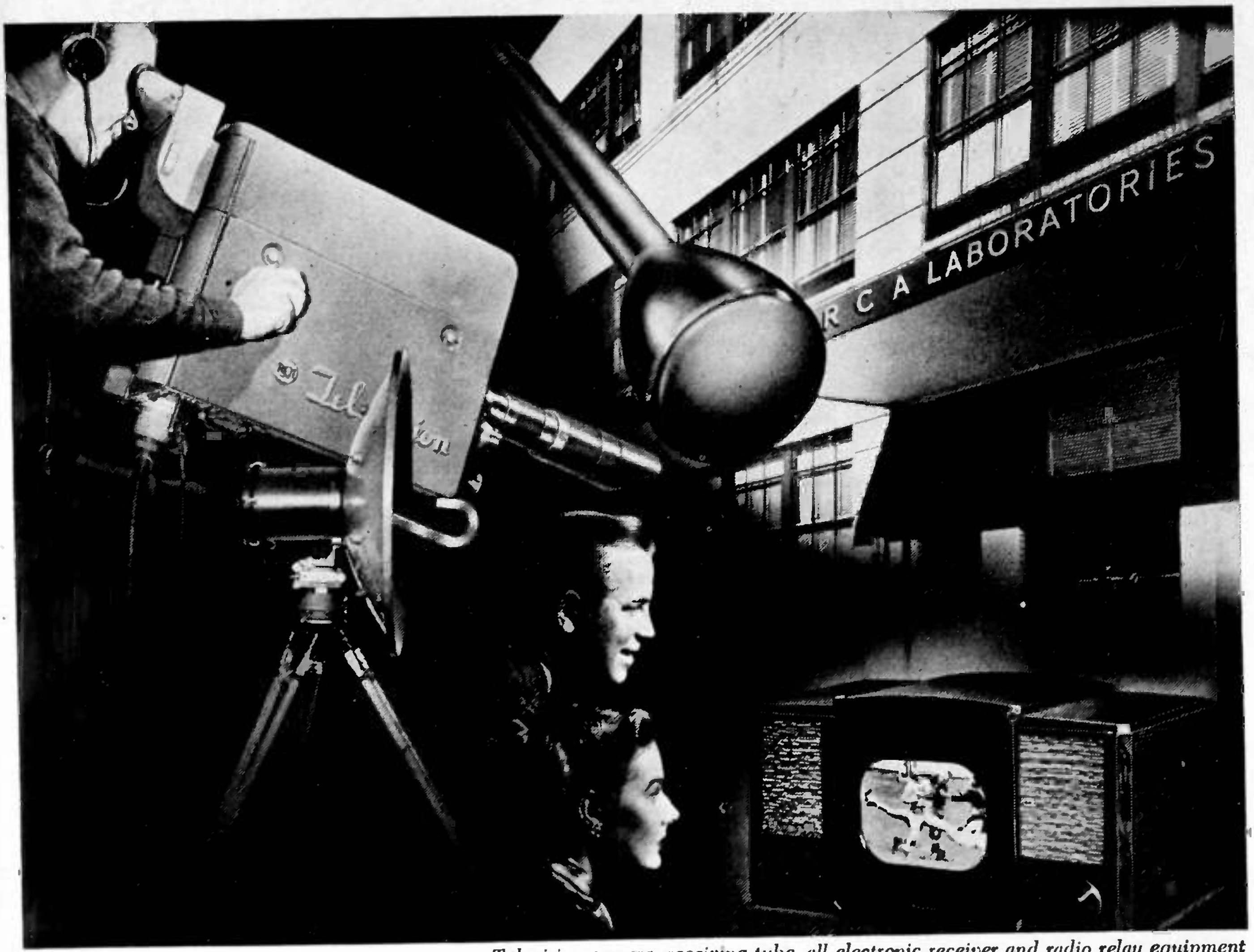
Synchroscope

A synchronoscope designed for the visual examination of the fine structure of periodic waveforms in television, pulse time modulation, sonic depthfinders, geophysical exploration and loran equipments has been placed on the market by the Electronics Division, Sylvania Electric Products, Inc., 500 Fifth Avenue, New York 18, N. Y.



This instrument includes a five inch cathode-ray oscilloscope; trigger generator for synchronization; space for the addition of a video amplifier and r-f envelope viewer; adjustable time-delay phasing circuits; and seven input connectors and selector switch for rapid viewing of separate external circuits. Many combinations of trigger and sweep pulses may be produced to feed synchronizing pulses to a variety of external equipments at pulse repetition frequencies up to several thousand per second.

(Continued on page 58A)



Television camera, receiving tube, all-electronic receiver and radio relay equipment —are the result of pioneering and research at RCA Laboratories.

Behind every big stride in Television—RCA Laboratories!

From the scene of action—to your own living room—these RCA developments based upon research at RCA Laboratories mean *television at its finest:*

RCA Image Orthicon Camera sees whatever the human eye sees, even in the light of a match! Sports events on cloudy days or in twilight do not fade because this super-sensitive camera eliminates the need for strong lighting.

RCA Mirror-backed Kinescope—searchlight brilliance for home television. All the lifelike realism and detail caught by the RCA Image Orthicon Camera is reproduced by this new receiving tube that loses none of the original brilliance.

RCA Victor Television Receiver—with the new RCA exclusive "Eye Witness" feature that "locks" the picture, keeps it bright, clear—as steady as a picture on the wall.

RCA Radio Relay equipment enables television stations to broadcast events taking place far from the studio, and eventually may link television networks. In television, as in radio, Victrola* radio-phonographs, records, or tubes, if it bears the name RCA or RCA Victor, it is one of the finest instruments of its kind science has achieved.

. . .

Radio Corporation of America, RCA Building, Radio City, New York 20 . . . Listen to The RCA Victor Show, Sundays, 2:00 P. M., Eastern Time, over NBC. *"Victrola" T. M. Reg. U. S. Pat. Off.

RCA VICTOR table model television receiver with the exclusive "Eye Witness Picture Synchronizer" that assures you *brighter, clearer, steadier* pictures. It is now available in some areas—see your local RCA Victor dealer.

RCA

RADIO CORPORATION of AMERICA

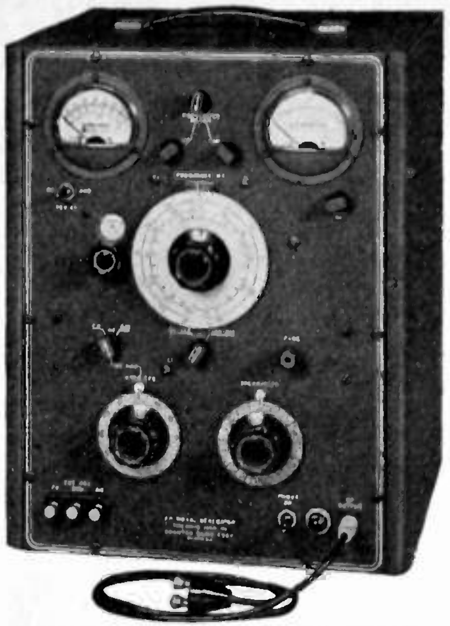
NEW!

FM SIGNAL GENERATOR

MODEL 202-B

FREQUENCY RANGE

54 to 216 MEGACYCLES



The model 202-B is specifically designed to meet the needs of television and FM engineers working in the frequency range from 54-216 mc. Following are some of the outstanding features of this instrument:

RF RANGES—54-108, 108-216 mc. ± 0.5% accuracy.

VERNIER DIAL—24:1 gear ratio with main frequency dial.

FREQUENCY DEVIATION RANGES—0-80 kc; 0-240 kc.

AMPLITUDE MODULATION—Continuously variable 0-50%; calibrated at 30% and 50% points.

MODULATING OSCILLATOR—Eight internal modulating frequencies from 50 cycles to 15 kc., available for FM or AM.

RF OUTPUT VOLTAGE—0.2 volt to 0.1 microvolt. Output impedance 26.5 ohms.

FM DISTORTION—Less than 2% at 75 kc deviation.

SPURIOUS RF OUTPUT—All spurious RF voltages 30 db or more below fundamental.

Write for Catalog D

This instrument was described editorially in November ELECTRONICS—reprints available on request

BOONTON RADIO Corporation

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BEAT FREQUENCY GENERATOR
AND OTHER DIRECT READING INSTRUMENTS

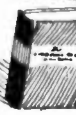
PRACTICAL WORKING TOOLS FOR THE ENGINEER

UP TO THE MINUTE!

ELECTRONIC ENGINEERING HANDBOOK

For radio-electronic specialists this Caldwell-Clements book provides a convenient, authentic source of formulas and principles, as well as the latest in electronic applications. Electrical operating and executive engineers will find here the solutions to many production problems. Easily understood by anyone with a knowledge of basic electrical principles and simple circuits. Free from involved mathematical explanations. Section I covers Vacuum Tube Fundamentals, Section II, Electronic Circuit Fundamentals; Section III, Electronic Applications; Section IV, Vacuum Tube Data.

456 Pages · $4.50
560 Illustrations



ELECTRONIC CONTROL HANDBOOK

Here are all the essential data necessary to determine the worth of an electronic control device; a dependable guide toward your taking advantage of the cost-cutting, production-speeding, quality-control possibilities of electronic devices. Gives you facts to intelligently balance the advantages of electronics against mechanical and other methods of control. Easily understood without advanced knowledge of electronics. Section I, Basic Elements of Control; Section II, Conversion Elements; Section III, Electronic Modification Circuits; Section IV, Activation Elements; Section V, Control Applications.

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Name..
Address..
City and State..
Company Name..

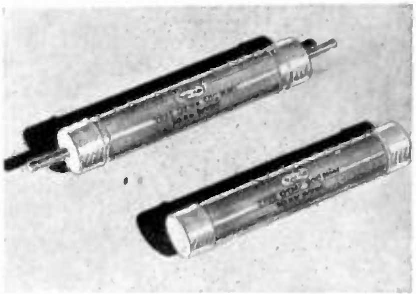
News—New Products

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

(Continued from page 56A)

Television Capacitors

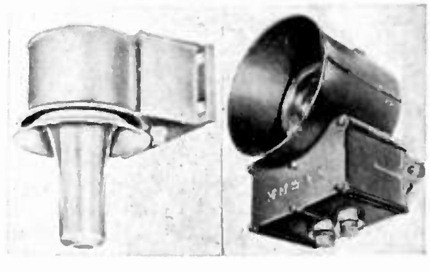
Hermetically-sealed high-voltage capacitors especially designed for use in rectified radio-frequency type power supplies for television receivers and other cathode-ray tube applications are being manufactured by Solar Manufacturing Corp., 285 Madison Ave., New York 17, N. Y. Excellent performance, under the high operating temperature found in the confined space of compact new set designs, is stated to be assured by utilizing selected, low power-factor paper and impregnating it with mineral oil under extra high vacuum.



These capacitors are available in voltage ratings up to 30 kilovolts working voltage, direct current, and in capacitances up to 500 mmf. They are housed in solder-sealed glass bushings with either ferrule or screw terminals. These constructions are identified as Solar types QTMF and QTMH, respectively.

Submergence Loudspeakers

University Loudspeakers, Inc., 225 Varick Street, New York 14, N. Y., announce two new loudspeakers which are designed to operate in explosive atmospheres and even under water. The manufacturer states that consistent operation is assured regardless of salt spray or extreme weather conditions.



Models MSR (left) and MM-2TC (right) are equipped with Alnico V permanent magnets. The first mentioned model has a radial deflector for 360° dispersion and the other unit is directional through 120°. Both units are designed for 15 watts output and have an impedance of 16 ohms.

(Continued on page 60A)


ELECTRONIC
VOLT
OHM
MA
MFD
MODEL 2450
OHMS ADJ.
SELECTOR
+DC VOLTS
OHMS
-DC VOLTS
OFF
DC MA
AC VOLTS
MFD
RANGE
CIRCUIT BALANCE
ACV-MFD
MFD
COMMON
DCV-OHMS-MA

News—New Products

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

(Continued from page 58A)

Tunable Dipole

A new type of dipole for television and FM reception has been announced by Kings Electronics, 372 Classon Avenue, Brooklyn 5, N. Y. Its efficiency, it is claimed, is considerably higher than other types due to the adjustable arms which can be harmonized with the wave length of weak stations.

The adjustable feature consists of an U.H.F. element which is calibrated from 1.0 to 21.5 in half-steps. This element can be moved in or out according to a calculated table to reinforce signals from weak stations. Provision is made for locking the arm after adjustment.

Ferrograph

The Ferrograph, developed to provide a simple and convenient method of comparing ferrous materials as to their chemical analysis and heat treatment, is now being released in its latest refined form by the Allen B. DuMont Laboratories, Inc., Passaic, N. J. This instrument is based on the correlation between magnetic properties, particularly remanent magnetism which predominates at the low frequency which is used, and metallurgical properties. It makes possible the determination of composition and condition of ferromagnetic materials by magnetic testing, with the cathode-ray tube as the instantaneous indicator.

The basic principle of operation of the Ferrograph is the harmonic analysis of the induced voltage in the secondary of a test transformer in which the sample to be tested is made the core. By indicating the amplitudes and phase angles of the fundamental and third harmonic components, the Ferrograph serves for many practical problems in industry and provides an instrument not too involved for use by unskilled industrial personnel.

(Continued on page 62A)


RHEOSTATS
TAP SWITCHES
RESISTORS
CHOKES
If it's made by OHMITE
it's DEPENDABLE

Announcing

The 1st volume in the long-awaited Massachusetts Institute of Technology RADIATION LABORATORY SERIES

Bringing you

the engineering data you need to design RADAR SYSTEMS

Here is an important book which presents the general principles of the design of various radar systems. From the standpoint of the designer the book discusses the basic considerations which underlie and are particular to systems design. After a general approach to problems encountered, it takes up the leading design considerations for the important components that make up a radar set. Two new and important auxiliary techniques—moving target indication and the transmission of radar displays to a remote indicator by radio means—are fully treated. Detailed examples of actual systems are included. Anyone interested in the varied applications of radar will find this new volume of immense value as a basic, useful reference.

Just Out

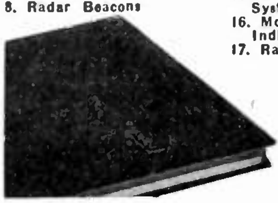
RADAR SYSTEM ENGINEERING

Edited by Louis N. Ridenour, Editor-in-Chief, Radiation Laboratory Series; Associate Professor of Physics, University of Pennsylvania. Approximately 900 pages, 6 x 9, $7.50.

This is the first of twenty-eight volumes prepared principally by members of the Radiation Laboratory maintained during the war at the Massachusetts Institute of Technology under contract with the National Defense Research Committee of the Office of Scientific Research and Development. The Laboratory was the foremost U.S. research and development institution in the field of microwave radar. The accuracy and usefulness of the material made available in these volumes is attested by their authoritative background.

Contents

1. Introduction
2. The Radar Equation
3. Properties of Radar Targets
4. Limitations of Pulse Radar
5. C-W Radar Systems
6. The Gathering and Presentation of Radar Oata
7. The Employment of Radar Data
8. Radar Beacons
9. Antennas, Scanners, and Stabilization
10. The Magnetron and the Pulser
11. R-F Components
12. The Receiving System—Radar Receivers
13. The Receiving System—Indicators
14. Prime Power Supplies for Radar
15. Examples of Radar System Oesign
16. Moving-target Indication
17. Radar Relay



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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 60A)

High Frequency Voltohmyst

The servicing of industrial and radio equipment using the very-high frequencies up to 250 megacycles will be facilitated by a new model Voltohmyst which is now in production and will be available shortly, it was announced by the Test and Measuring Equipment Section of the Radio Corporation of America, Camden, N. J.

Employing a newly developed diode probe, and capable of measuring peak-to-peak voltages at very high frequencies, the new meter, designated as RCA Type WV-75A, has several innovations which make it particularly suitable for high-frequency work. When used for FM and television testing, the meter can be used to make all measurements in the receiver up to 1000 volts, as well as bias-cell voltages and the values of the Automatic Frequency Control, Automatic Volume Control, and FM discriminator voltages.

AC-DC Amplifier

A high quality, inexpensive AC-DC, type A-319 amplifier designed primarily for use in commercial wired-music systems is being manufactured by Altec Lansing Corp., 1161 N. Vine St., Hollywood 28, Calif. It can be used also as a terminal amplifier for paging systems, and elsewhere requiring a medium gain low power amplifier.

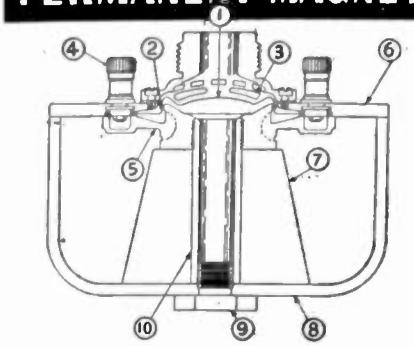
Water Cooled Resistor

International Resistance Company, 401 North Broad St., Philadelphia 8, Pa., recently announced their new Type LP, liquid cooled, high-frequency, high-power resistor. This development is for television, FM and dielectric heating applications.

Inside the resistor a high velocity stream of water flows in a spiral path against the resistance film. Interchangeable intake nozzles permit adjusting the rate of water flow and therefore, the cooling action, to suit local water pressure and power dissipation up to 5 kw.

(Continued on page 64A)

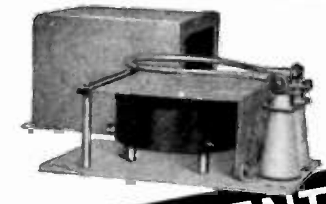
Inside Facts About ROWE *No. 7* PERMANENT MAGNET



1—One-piece metal diaphragm. Voice coil wound on diaphragm.
2—Horn mounting accurately centered.
3—Perforated die-cast palate.
4—Cadmium plated heavy-duty binding posts.
5—Flange molded to inner pole piece.
6—Outer pole piece of special alloy.
7—Heavy ALNICO permanent magnet.
8—Bowl of heavy gauge steel.
9—Brass assembly nut binds all parts.
10—Inner pole piece made of brass.

Write for circular No. 64 giving complete details.

ROWE *Industries*
1702 WAYNE ST., TOLEDO 9, OHIO



RF CURRENT TRANSFORMER

R. F. Current Transformers provide remote metering and antenna current readings at the antenna for power up to 50 KW.

R. F. Sampling Transformers (same in appearance) are highly recommended for phase sampling antenna current in directional systems with shorter towers. They're totally shielded from external stray fields and free of electrostatic coupling. Complete data and prices on request from department W

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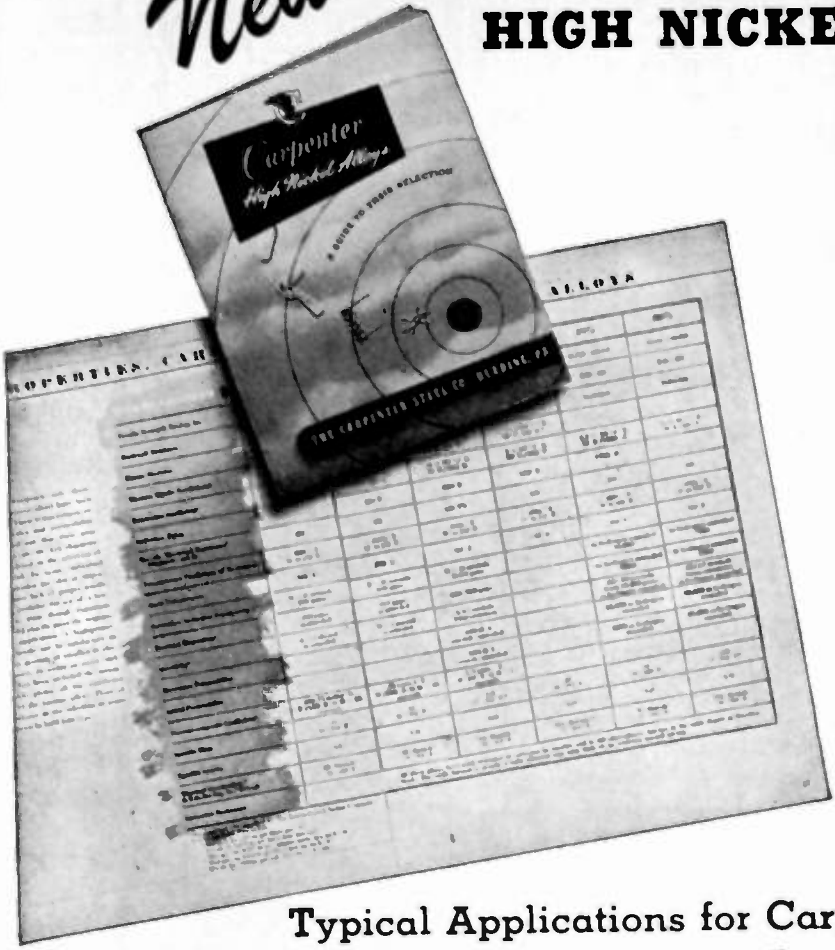
New

TECHNICAL INFORMATION ON HIGH NICKEL ALLOYS FOR . . .



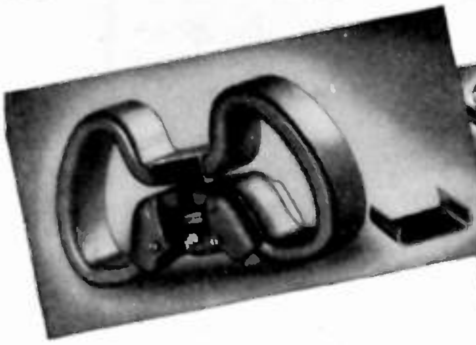
Low Expansion
Temperature Compensation
Glass Sealing
Magnetic Permeability

This information, much of it never before published, will help you find new ways to improve product performance with iron-nickel alloys. Carpenter's new 22-page bulletin includes a complete chart showing the general effect of various percentages of nickel on the magnetic and expansion properties of iron.

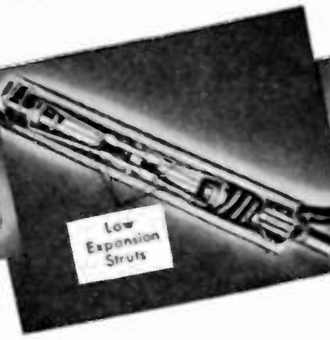
Typical applications for High Nickel Alloys are described and illustrated, and a Table of Physical Properties gives basic data on these alloys.

For a copy of the new Carpenter High Nickel Alloy Bulletin, drop us a note on your company letterhead, indicating your title. Your Bulletin will be sent promptly.

Typical Applications for Carpenter High Nickel Alloys



Watthour meter permanent magnet assembly with shunt of Carpenter Temperature Compensator "30".



Thermal switch using Carpenter Low Expansion "42" for low expansion struts as indicated.

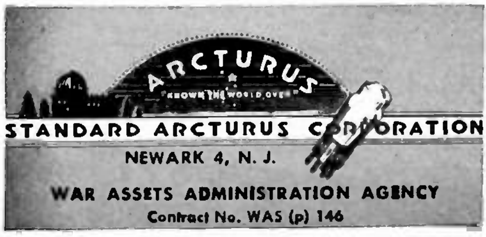
Water heater thermostat employing Carpenter Free-Cut Invar "36" together with a high expansion alloy.

High wattage lamp with ferrules of Carpenter Glass Sealing "42" used to bring out leads.

THE CARPENTER STEEL COMPANY—171 W. Bern St., Reading, Pa.

Carpenter
High Nickel Alloys

Temperature Compensator "30"
Carpenter Invar "36"
Carpenter Free-Cut Invar "36"
Glass Sealing "42"
High Permeability "49"



News—New Products

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

(Continued from page 62A)

60-Watt Transmitter

A new 60-watt radio transmitter is now in production by the John Meck Industries of Plymouth, Ind. This transmitter, which is a compact single unit, has a 6L6G in a regenerative oscillator circuit, operating with the output at the crystal frequency on all bands except 10 meters. For operation on 10 meters, the plate of the oscillator is tuned to the second harmonic of the crystal.

Other features of the transmitter are: built-in antenna changeover relay; openings for plug-in battery or "vibropac" operation; meter switching to final amplifier grid or plate circuits for tuning; crystal socket on front panel for rapid-frequency changing; and a send-receive switch for receiver standby.

Portable Bench Rectifier

A new portable rectifier unit, model 725S1C has been announced by W. Green Electric Co., 130 Cedar St., New York 6, N. Y. This rectifier, a complete DC power supply source with a capacity of 150 watts, utilizes a quadruple protected selenium rectifier element.

Flameproof Aircraft Wire

A new aircraft wire which reduces the fire hazard and lightens the weight of planes was recently announced by United States Rubber Company. Known as Neolay, the new product is 30% lighter than conventional aircraft wire. If used in a plane the size of the B-29 Superfortress, it would save nearly 300 pounds. In addition to being flameproof, tests indicate it to be resistant to oil, chemicals, mildew and fungus.

(Continued on page 66A)

when specifications say...

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specify BLILEY CRYSTALS

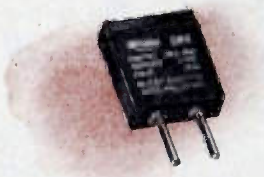

AMERICAN AIRLINES PHOTO

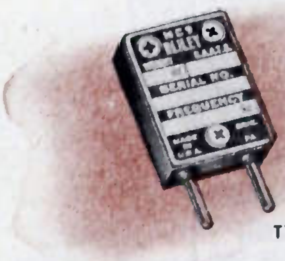
In air carrier aircraft the seal of airworthiness is CAATC . . . Civil Aeronautics Administration Type Certification. The Bliley crystal units, shown on this page, are available with CAATC when specified.

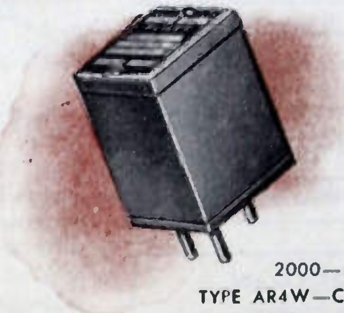
The prime requisite for airworthiness is reliability. Each Bliley crystal, whether standard or CAATC, is rigidly tested for reliable performance. Frequency stability, precision and activity are proven for all conditions of temperature, moisture and vibration covered by the specifications that govern.

Engineers everywhere rely on Bliley "techniquality" for the answer to their frequency control problems. When you specify Bliley crystals you automatically include the creative engineering and production talent that has pioneered in frequency control for over fifteen years.

Write for Bulletin P-27


2000—11000KC
TYPE SR5—CAATC No. 363


3000—11000KC
TYPE MC9—CAATC NO. 362


2000—5000KC
TYPE AR4W—CAATC NO. 360
TYPE AR5W—CAATC NO. 361

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Great Neck, New York Imperial 645

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Industrial Electronics
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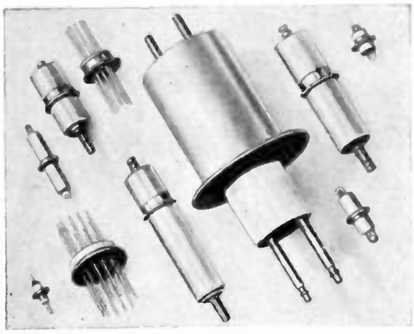
272 Centre St., Newton, Mass. BIG-9240

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 64A)

Sealed Bushings



General Ceramics & Steatite Corp., Keasbey, N. J., offers an expanded line of sealed bushings suitable for practically all applications. These terminals provide permanent hermetic sealing capable of withstanding unusually severe mechanical and thermal shock. The steatite dielectric is chemically bonded to the metal collar and electrode. The coefficients of expansion of the "Steatite" and metal are matched to insure a bond which, it is claimed, is uneffected by frequent or even violent temperature fluctuation.

Recent Catalogs

• • •On Flat Wire Wound Resistors, by **International Resistance Co.**, 401 N. Broad St., Philadelphia 8, Pa. Catalog No. C-1.

• • •A Reference Guide for Miniature Electron Tubes, by **Hytron Radio & Electronics Corp.**, 76 Lafayette St., Salem, Mass.

• • •On Electronic Motor Controls, by **Electron Equipment Corp.**, 917 Meridian Ave., So. Pasadena, Calif. Bulletin No. 178

• • •On Solder, Fluxes, Wire Strippers, and Flux Removers, by **Division Lead Co.**, 836 W. Kinzie St., Chicago 22, Ill., Bulletin Nos. 1A through 4A.

• • •On "Lectrofilm" Capacitors for Radio-Frequency Blocking, by **General Electric Co.**, Schenectady, N. Y. Bulletin No. GEA 4295A.

• • •On Heavy-Duty Multiple-Arm Relays, by **Signal Engineering & Mfg. Co.**, 152 West 14 St., New York 11, N. Y. Bulletin No. 30.

• • •On Speech Equipment, Amplifiers, Turntables, and Monitoring Units, by **Collins Radio Co.**, Cedar Rapids, Iowa. Bulletin No. N8314.

• • •On Electronically Regulated Power Supplies, by **Furst Electronics**, 800 W. North Ave., Chicago 22, Ill. Bulletin No. 310.

• • •On Sound Recording and Playback Equipment, by **The Fairchild Camera & Instrument Corp.**, 88-06 Van Wyck, Blvd., Jamaica 1, N. Y. Catalog No. 1246.

• • •On Type PRT Power Rheostats by **International Resistance Co.**, 401 No. Broad St., Philadelphia 8, Pa. Bulletin No. E-1.

• • •Technical Property Chart of "AlSiMag" Ceramics, by **American Lava Corp.**, Chattanooga 5, Tenn. Chart No. 346.

• • •On Ship-to-Shore Radio Equipment, by **Radiomarine Corporation of America**, 75 Varick St., New York 13, N. Y. Bulletin No. PC102.

• • •On Capacitor Analyzers, by **Solar Manufacturing Corp.**, 285 Madison Ave., New York 17, N. Y. Catalog No. IN-2.

• • •Electronic Tube Data Book, by **Westinghouse Electric Corp.**, P. O. Box 868, Pittsburgh, Pa. Booklet No. 86-020.

• • •On Electronic Parts and Components, by **Radionic Equipment Co.**, 170 Nassau St., New York 7, N. Y. Catalog No. 47.

• • •On Physiological Pulse Stimulators, by **The Electrodyne Co.**, 899 Boylston St., Boston 15, Mass. Bulletin No. 461.

Interesting Abstracts

• • •On "Theory, Construction and Applications of the Crystal Diode," printed in October 1946 issue of The Aerovox published by **Aerovox Corporation**, New Bedford, Mass.

• • •On "New Midget Radio Tubes," printed in October 1946 issue of the Philips Technical Review published by **Philips Laboratories, Inc.**, 100 East 42 St., New York 17, N. Y.

(Continued on page 68A)

THE NEW, COMPACT SUPER-SENSITIVE *Analyzer*

Weston

MODEL 779

Extreme compactness and lightweight—dual DC voltage sensitivity of either 1000 or 20,000 ohms per volt – five AC and DC voltage ranges, seven DC current ranges, four DC resistance ranges, and five decibel ranges – all carefully selected to meet the broadest requirements of testing and maintenance —precision WESTON resistors throughout—large 50 microampere WESTON meter – temperature compensated including AC ranges – size only 6⅜" x 9⅛" x 4⅞" – furnished in rugged, solid oak carrying case.

NOW AVAILABLE . . . see Model 779 at the Radio Parts and Electronic Show . . . Stevens Hotel . . . Booth No. 75. Weston Electrical Instrument Corporation, 589 Frelinghuysen Avenue, Newark 5, New Jersey.

Model 779 is designed for use with WESTON Socket Selectors which facilitate checking tube circuit conditions – and with WESTON Televerters for DC voltage measurements up to 10,000 volts.

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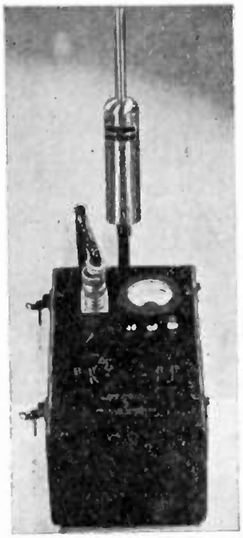
NEWS and NEW PRODUCTS

January, 1947

Sound Pressure Meter

A precision electro-acoustic instrument for conveniently making absolute sound-pressure measurements over the entire audible and low-supersonic frequency range to about 40 kilocycles, has been announced by **Massa Laboratories, Inc.**, 3868 Carnegie Ave., Cleveland 15, Ohio. This Model GA-1002 Sound Pressure Measurement System is intended as a reliable standard, but so designed that it may be used as an everyday test instrument without change in calibration. The complete system includes a Model M-101 standard microphone, a shock-mounted preamplifier with 15-foot cable, and an auxiliary amplifier complete with dry batteries. A built-in calibrating circuit permits setting the gain to produce an output of 1 millivolt per dyne/cm² sound pressure so that a conventional electronic voltmeter can be employed for the direct reading of sound pressure.

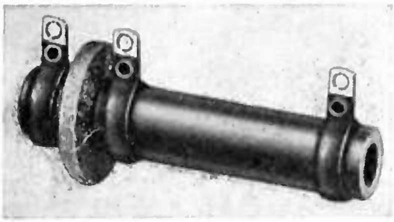


A wide dynamic range permits measurements from less than 1 dyne/cm² to 20,000 dynes/cm² (160 db level) without distortion. For measuring pressures up to several million dynes/cm², a non-frequency discriminating attenuator can be provided to prevent overloading of the input circuit. It is claimed that while shock pressures generated by explosions or engine exhausts may be measured by this system, it also meets the exacting requirements for a stable and accurate laboratory instrument. It is further stated that due to the extremely high acoustic impedance and small size of the microphone, free-field measurements may be more accurately made at the higher frequencies or inside confined spaces than has heretofore been possible with other types of available instruments.

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)

Videocoupler



A three-terminal network, combining two peaking inductances and the load resistor in one assembly, known as Type VC-1 Videocoupler, designed to couple the video amplifier to the picture tube of television equipment, has been placed on the market by P. R. Mallory & Co., 3029 E. Washington St., Indianapolis 6, Indiana.

Electronic Counter

An electronic counter offering great accuracy and convenience for nuclear measurement, incorporating a precision pulse amplitude discriminator in the input circuit which allows only pulses greater than a predetermined amplitude to operate the scaler, has been recently announced by the Atomic Instrument Co., 154 Charles St., Boston 14, Mass.

By means of a direct reading dial, on the Model 101 Scaler, the discrimination level may be accurately set to within 1% for pulses between 50 and 100 volts. It is claimed that the resolution of the scaler is such that pulses occurring as close together as 5 microseconds will be individually recorded. A counting rate of 100,000 counts per minute at only 1½% loss is obtained with this unit.

Hyflux Recording Tape

The use of a magnetic powder, having magnetic characteristics comparable to those of well-known grades of Alnico, when coated on a paper tape, produces a high fidelity tape recording medium, according to The Indiana Steel Products Co., 6 North Michigan Avenue, Chicago 2, Illinois, whose Research Department and the Battelle Memorial Institute of Columbus, Ohio, have cooperated on this development.

Considerable research and experimentation have been accomplished in the perfection of a sound recorder which will best utilize the qualities of this new Hyflux magnetic tape. It will be simple to operate, and economical to produce.

Voltage Regulator

Designed specifically for aircraft use and general industrial applications, model D-500, AC voltage regulator employs lightweight "fosterited" transformers and is hermetically sealed to withstand climatic conditions. A regulation accuracy of 0.5% is obtained through a load range of 50 to 500 VA, at 360–500 cycles with an output voltage of 110–120, according to the manufacturer. Additional models with greater power ratings can also be provided by Sorenson & Co., 375 Fairfield Avenue, Stamford, Conn.

Mobile Telvision Unit

Development of a lightweight, self-contained mobile television unit which will greatly facilitate news coverage and other remote pickup operations has been recently announced by the Engineering Products Department, **Radio Corporation of America**, Camden, N. J. This new Mobile Television Unit, mounted on a standard 1½ ton truck chassis, can be used to transport all the equipment required for picking up, monitoring, and relaying to the studio remote-television events, and provides for mounting the "cameras" on the roof of the truck body or placing them as much as 500 feet away from the mobile unit.

(Continued on page 70A)

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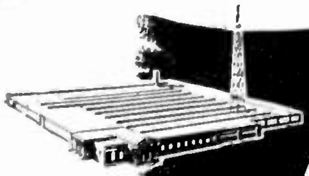
This is the long-awaited Hallicrafters SX-42, a truly great communications receiver. The tremendous frequency range of the SX-42, *greater than ever before available in a receiver of this type*, is made possible by the development of a new "split-stator" tuning system and the use of dual intermediate frequency transformers. Packed with advance features that every ham and every other radio enthusiast desires, the SX-42 clearly lives up to the Hallicrafters ideal of "the radio man's radio."

From now on watch Hallicrafters — the name that's remembered by the veteran, preferred by the radio amateur. See your distributor for demonstration of the SX-42 and for colorful literature describing this great set in complete technical detail.

Because of the precise and thorough engineering that must be done on the SX-42 and because the parts supply has not been continuous, top production peaks have not yet been reached. In the immediate future deliveries will necessarily run behind the demand, but the SX-42 is definitely worth waiting for.



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BUILDERS OF Skyfone AVIATION RADIOTELEPHONE

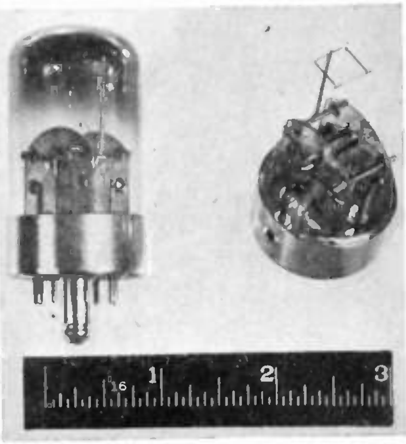
News—New Products

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)

Acceleration Pickup

A new electronic tube, appearing much as an ordinary radio tube but capable of measuring accurately the rapidly changing accelerations to which various parts of an airplane are subjected in flight, has been developed at the **National Bureau of Standards,** Washington 25, D. C. This new tube, known as the Vacuum Tube Acceleration Pickup, is also proving useful in such applications as the measuring of accelerations in portions of the body of "dummy" pilots when subjected to critical acceleration during crash landings or seat-ejections from jet-propelled airplanes.



The pickup takes advantage of the effect of acceleration on the relative position of the electrodes in the tube. This is in contrast to the design of conventional tubes where this effect is suppressed as much as possible. The tube contains a fixed, indirectly-heated cathode with two plates, one on either side. The plates are elastically mounted to deflect in response to acceleration normal to the plane of the plates. This deflection causes a change in plate current proportional to the acceleration and such changes in current are recorded on a standard oscillograph.

Sound Reproducer

The Allen D. Cardwell Mfg. Corp., 97 Whiting St., Plainville, Conn., announce their new Model CE-26 Sound Reproducer which contains, in addition to a ten-inch heavy-duty permanent magnet speaker, an amplifier with 8 watts undistorted output and 11 watts maximum output. The manufacturer states that the frequency response is flat within 2 decibels from 60 to 8000 cycles and that four additional speakers may be tapped onto the output transformer.

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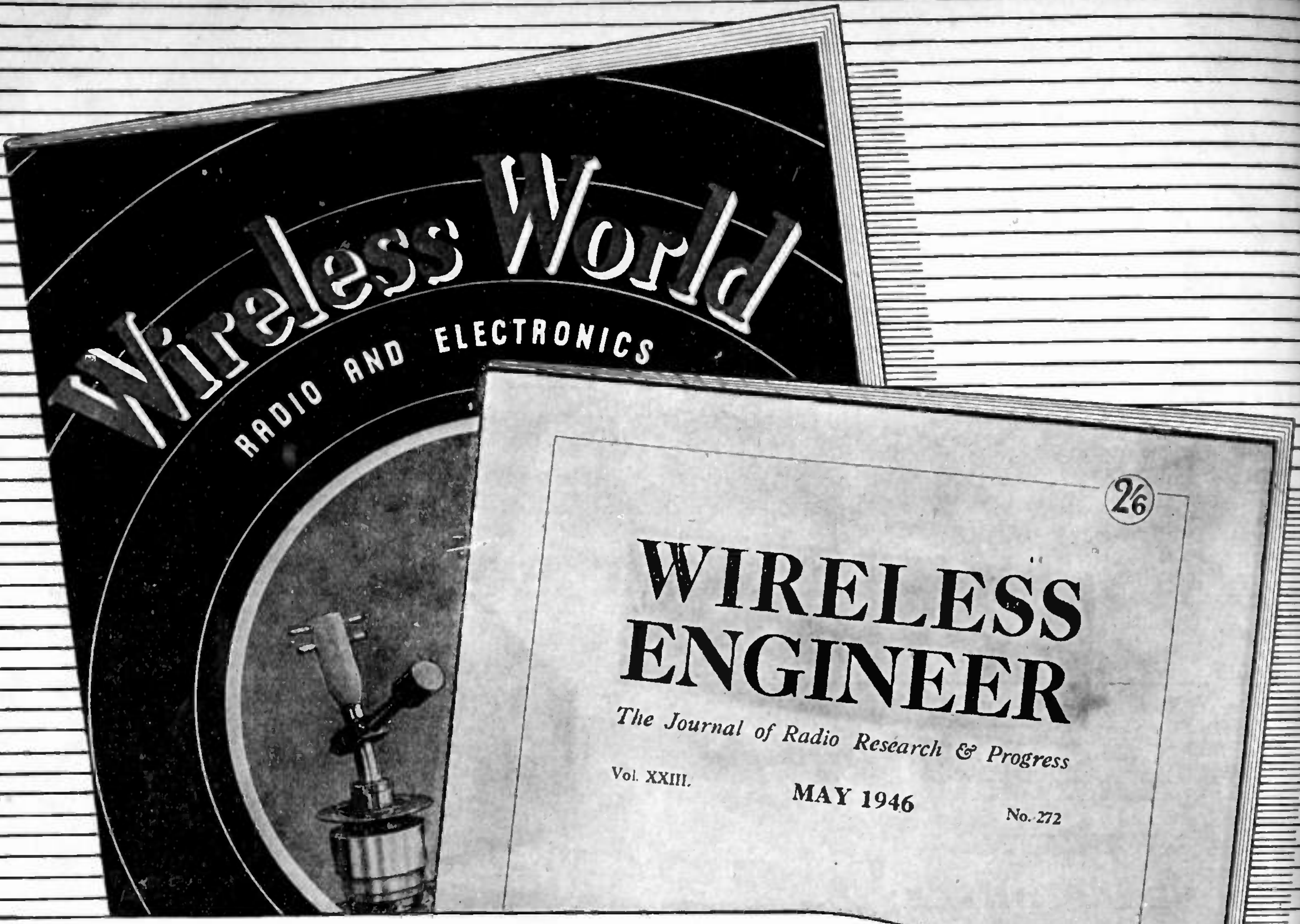

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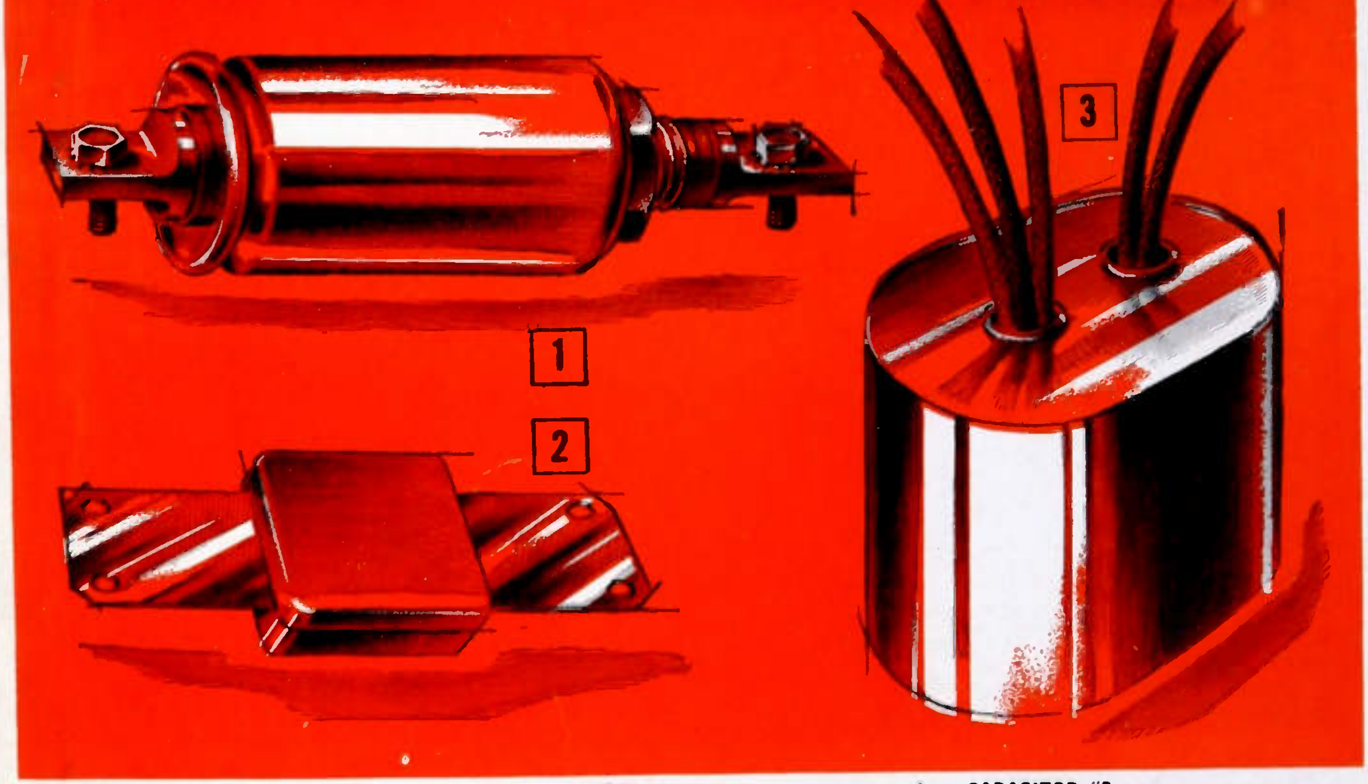

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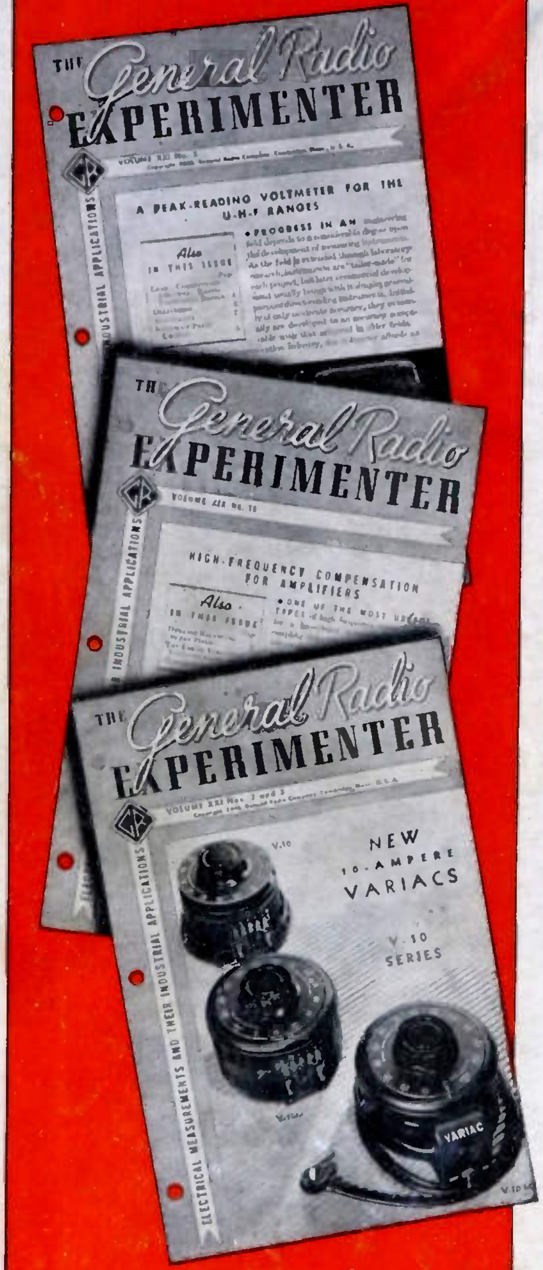

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