EXPERIMENTAL UHF TRANSMITTER
At 850 megacycles, an output power of 400 watts is obtained by multiple operation of tube groups.
FOR COMPACT HIGH FIDELITY EQUIPMENT

Ultra compact, lightweight, these UTC audio units are ideal for remote control amplifier and similar small equipment. New design methods provide high fidelity in all individual units, the frequency response being =2DB from 30 to 20,000 cycles. There is no need to resonate one unit in an amplifier to compensate for the drop of another unit. All units, except those carrying DC in Primary, employ a true hum balancing coil structure which, combined with a high conductivity outer case, effects good inductive shielding. Maximum operating level + 10 DB. Weight—8 ounces. Dimensions—1 1/2" wide x

HERMETICALLY SEALED

On special order, we can supply any of the Ultra Compacts hermetically sealed per Jan T-27 Grade 1 Class A in our RC 50 case as illustrated. Dimensions: Height 2 1/4", Base 1 3/8" x 1 9/16.

ULTRA COMPACT HIGH FIDELITY AUDIO UNITS

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

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

United Transformer Co.
150 Varick Street, New York 13, N.Y.
Export Division: 13 East 40th Street, New York 16, N.Y. Cables: "ARLAB"
The annual Southwestern I.R.E. Conference is again being sponsored by the Dallas-Fort Worth Section. This Conference is of outstanding importance to all members in Region 6. Under the theme of "New Developments in Radio Electronics," it is dedicated to satisfying the informative needs of radio and electronics engineers in the Southwest.

CHECK THESE FEATURES

1. Manufacturers Exhibit Show: The latest in radio engineering equipment will be shown throughout the two day session.

2. Technical Sessions: Selected papers in the fields of Radio Broadcast, Television, Measurements, and Geophysics will be presented by nationally known men in these fields.

3. Banquet: President Stuart L. Bailey will be the speaker.


For Further Information Write

Registration
G. B. Loper
2621 Easter Drive
Dallas 8, Texas

Exhibits
John F. Klutitz
Radio Station KRLD
Dallas 1, Texas

Technical Program
John K. Godbey
3120 Dutton Drive
Dallas 11, Texas

The Skyline of Dallas, Texas

IRE Regional Meetings Accelerate Electronic Progress!
**NOTE THE SIZES AND RATINGS!**

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

<table>
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</tbody>
</table>

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

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

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

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

Write for Sprague Engineering Bulletin 203.
NEW - hp - ACCESSORIES INCREASE
SCOPE OF YOUR - hp - VOLTMETERS

- hp - 452A Capacitive Voltage Divider
For - hp - 400A, 400C and 410A Voltmeters. Safely measure power, supersonic and dielectric heating voltages to 25 kv. Accuracy ± 3%. Frequency range, 25 cps to 20 mc. Division ratio 1,000:1. Input capacity 15 μf. Price $75.00.

Extend the usefulness of your present - hp - voltmeters with these new precision-built - hp - accessories. Save time and work. Simplify tedious jobs. Make fast, accurate measurements far beyond the original range of your instruments.

- hp - 453A Capacitive Voltage Divider
For - hp - 410A Voltmeter. Increases range so transmitter voltages can be measured quickly, easily. Accuracy ± 1%. Division ratio, 100:1. Input impedance 50 megs. Max. voltage 2,000 v. For frequencies 10 kc and above. Price $20.00.

- hp - 454A Capacitive Voltage Divider
For - hp - 400C Voltmeters. Safely measure power, audio, supersonic and rf voltages. Accuracy ± 3%. Division ratio, 100:1. Input impedance 50 megs. resistance shunted with 2.75 μf. capacity. Max. voltage 1,500 v. Price $20.00.

- hp - 455A Probe Coaxial "T" Connector
For - hp - 410A Voltmeter. Measures voltages between center conductor and sheath of 50 ohm transmission line. Maximum standing wave ratio 1 to 1.1 at 500 mc; 1 to 1.2 at 1,000 mc. Male and female Type "N" fittings. Price $35.00.

Write for details or see your - hp - Representative.

HEWLETT-PACKARD CO.
19350 PAGE MILL ROAD - PALO ALTO, CALIFORNIA
Export Agents: Frazar & Hansen, Ltd.
301 Clay Street, San Francisco 11, California, U.S.A.

- hp - 470A - 470F Shunt Resistors
For - hp - 400A or 400C Voltmeters, to measure currents as small as 1 μa full scale. Accuracy, ± 1% to 100 kc, ± 5% to 2 mc. Max. power dissipation 1 watt.

- hp - 458A Probe Coaxial "N" Connector

All prices and data subject to change
Prices F. O. B. Palo Alto

- hp - laboratory instruments
FOR SPEED AND ACCURACY
He finds trouble by ear

As this cableman runs his pickup coil along the cable, his ear tells him when he has hit the *exact spot* where unseen trouble is interfering with somebody's telephone service.

Trouble develops when water enters a cable sheath cracked perhaps by a bullet or a flying stone. With insulation damaged, currents stray from one wire to another or to the sheath. At the telephone office, electrical tests on the faulty wires tell a repairman approximately where to look for the damage.

A special “tracer” current, sent over the faulty wires, generates a magnetic field. Held against the sheath, an exploring coil picks up the distinctive tracer signal and sends it through an amplifier on the man’s belt to headphones. A change in signal strength along the cable tells the exact location of the “fault.”

Compact, light, simple to use, this test set makes it easier for repairmen to keep your line in order. It is another example of how Bell Laboratories research helps make Bell Telephone service the most dependable in the world.

**BELL TELEPHONE LABORATORIES**

Exploring and inventing, devising and perfecting, for continued improvements and economies in telephone service.
Hi-Q engineers were recently asked to design a component which would replace the 4 standard components called for in the schematic drawing illustrated at left. The problem was one of space saving without affecting the operation of the circuit.

**The Solution**

Hi-Q engineers designed a printed circuit known as the Hi-Q P.C. 100. This component replaced all 4 of the standard sized units formerly used, thus reducing the physical proportions of the space formerly required. In addition, this new component eliminated 25% of soldering time as well as eliminating 75% of the unit handling cost. The result of this customer’s foresight in placing his problem before Hi-Q engineers is that a new component was designed which saved our customer space, labor and time.

**What’s Your Problem?**

Our engineering department will gladly work with you on any problems you might have. Consult with us and ask for our suggestions regarding your specifications before your design has gone too far. Perhaps we can work out savings in space, time and labor for you.

*difference in space, time, labor, cost*

**WRITE FOR FREE CATALOG**

**Hi-Q Components**

**Better 4 Ways**

- Precision: Tested step by step from raw material to finished product. Accuracy guaranteed to your specified tolerance.
- Uniformity: Constancy of quality is maintained over entire production through continuous manufacturing controls.
- Dependability: Interpret this factor in terms of your customer's satisfaction. Our Hi-Q makes your product better.
- Miniaturization: The smallest BIG VALUE components in the business make possible space saving factors which reduce your production costs — increase your profits.
a simplified, outstandingly dependable LINE SWITCH for Stackpole Controls

Only .888" in diameter by .312" thick, this Type A-10 double-pole, single-throw line switch fits even the smallest Stackpole controls. Rated 1 ampere at 250 volts AC-DC or 3 amperes at 125 volts AC-DC, it combines outstanding ruggedness of design with ample-sized contacts and positive contact wiping action. Stationary contacts are mounted on a fiber surfaced Bakelite base to reduce arc tracing. The base is held securely in the can. Throughout, the switch is constructed for long, trouble-free service and in suitable ratings for portable and auto radios and numerous other applications. A similar single-pole design (Type A-11) with dummy terminal is also available.

Write for Stackpole Bulletin RC-7

ELECTRONIC COMPONENTS DIVISION
STACKPOLE CARBON COMPANY, ST. MARYS, PA.

STACKPOLE VARIABLE RESISTORS FOR MODERN RADIO AND TELEVISION NEEDS
YOU GET A SQUARE DEAL
AT OUR ROUND TABLE

When you bring your sheet metal fabrication problems to KARP, you immediately set in motion a "round table" board of experts whose combined specialized skill and experience is without an equal in the field. This group includes the president, chief engineer, chief draftsman-designer, chief toolmaker, plant superintendent, production manager and cost accountant.

These men make a detailed study of your special requirements. They plan, design and engineer the job with your needs and uses in mind. They determine the best manner of producing it, utilizing KARP'S superior equipment and facilities to your greatest advantage.

When your job is finished, it will be correctly designed for its application, handsome, rugged and built for long service life. You will have no costly problem of assembly ... no need to spend additional time and labor on finishing touches. The job will be COMPLETE, ready for the installation of your electrical or mechanical operating parts with ease and simplicity. No matter how many units you order, every last detail will be absolutely uniform.

This custom service not only gives your product added value, but under KARP methods may often save you money.

Consult us for cabinets, housings, chassis, racks, boxes, enclosures or any type of sheet metal fabrication.

Karp METAL PRODUCTS CO., INC.
Custom Craftsmen in Sheet Metal
223 — 63rd STREET, BROOKLYN 20, NEW YORK
Heat dissipation can be mighty tough ... but not for IRC resistors. They are universally engineered for the lowest possible operating temperatures and maximum power dissipation within the smallest size units consistent with good engineering practice.

Long experience with the widest line of resistor types in the industry has provided IRC with a wealth of "know-how" on resistor heat dissipation. In Power Wire Wound Resistors for example, the complete range of tubular and flat types manufactured by IRC utilizes a special cement coating to attain rapid heat dissipation. This dark rough surface does double duty by effectively guarding the windings against harmful atmospheric moisture and corrosion. Use the handy coupon to get complete data on proven advantages of IRC Power Wire Wounds.
Heat dissipation properties of aluminum are used to full advantage in housing and winding core of IRC Power Rheostats, 25 and 50 watts. Type PR Rheostats operate at full rating at about half temperature rise of equivalent units. Can be operated at full power in as low as 25% of rotation without appreciable difference in temperature rise. Direct contact between rheostat and mounting panel allows rapid conduction to panel of a portion of heat dissipated. Send for Bulletin E-2.

New, ADVANCED BT Resistors obsolete present performance standards for fixed composition resistors. Extremely low operating temperature and excellent power dissipation in compact, light weight, fully insulated units at 1/5, 1/2, 1 and 2 watts. These ADVANCED resistors meet JAN-R-11 specifications. All the facts are included in 12-page technical data Bulletin B-1.

Water-cooled LP Resistors utilize high velocity water stream flowing in spiral path against thin resistance film. High power dissipation is made possible by centrifugal force holding water in thermal contact with resistance surface. Resistance film less than 0.001” thick with active length much less than 1/4 wave length at FM and television frequencies, gives excellent frequency characteristics. Resistance values 35 to 1500 ohms; 15% tolerance standard; power dissipation up to 5 K.W. ac. Bulletin F-2 gives all the facts.

If you have the heat put to you for speedy service on small order resistor requirements for experimental work, pilot runs, etc., you'll appreciate the advantages of IRC's Industrial Service Plan. This enables you to get 'round-the-corner service from the local stocks of your IRC Distributor. He's a good man to know . . . we'll gladly send you his name and address.

INTERNATIONAL RESISTANCE COMPANY
405 N. BROAD ST., PHILA. 8, PA.
Send me additional data on items checked below:
- Power Wire Wounds (tubular)
- Flat Power Wire Wounds
- Advanced BT Resistors
- Power Rheostats
- Water-Cooled Resistors
- Name and address of our local IRC Distributor

NAME: __________________________
TITLE: __________________________
COMPANY: _______________________
ADDRESS: ________________________

END
Hundreds of thousands are now enjoying RCA's thrilling new way of playing records... they marvel at its wonderful tone... and the speed with which it changes records.

Prolonged research is behind this achievement, research which sought—for the first time in 80 years of phonograph history—a record and automatic player designed for each other.

Revolutionary is its record-changing principle, with mechanism inside the central spindle post on which records are so easily stacked. Result: a simplified machine, that changes records in 5 seconds.

Remarkable, too, are the new records—only 6½ inches in diameter—yet giving as much playing time as conventional 12-inch records. Unbreakable, these compact vinyl plastic discs use only the distortion-free "quality zone"... for unbelievable beauty of tone.

Value of the research behind RCA's 45 rpm system—which was started 11 years ago at RCA Laboratories—is seen in the instant acceptance, by the public, of this better way of playing records. Music lovers may now have both the 45 rpm system, and the conventional "78."

Development of an entirely new record-playing principle is just one of hundreds of ways in which RCA research works for you. Leadership in science and engineering adds value beyond price to any product of RCA, or RCA Victor.
TORTURE TESTS PROVE SWITCH PERFORMANCE

Turn it on. Turn it off. Do this 25,000 times or more and you’ll get a good idea of the terrific punishment Centralab switches must be able to withstand. Day and night, skilled CRL engineers and specially designed testing machines put Centralab switches through torture tests no switch is ever asked to undergo in ordinary operation. What does this mean to you? Just this. You can be sure that Centralab gives you the smooth operation, positive indexing and accurate positioning you want in the switches you buy. What’s more, you can be sure CRL switches will continue to provide these advantages for a long, long time.

Constant checking makes sure CRL switches give you desirable uniform low contact resistance. Here an engineer tests resistance by running 1 ampere through contacts.

Accelerated life test machine rotates through fixed number of positions at 1000 cycles per hour proves switch springs, clips and contacts stand up under long, hard use.

Resistance of switch insulation to atmospheric change is tested in controlled temperature and humidity chamber. Test helps avoid breakdown or leakage.
Centralab reports to

1. Up to 24 insulated clips on each section — an exclusive CRL feature — assures great variety of switching combinations... cuts size and cost of units. • No rotor rivets used. Where it's not necessary to connect contacts on opposite side of rotor, contacts are held by legs formed on contact. • Stator and rotor constructed in only the highest grades of laminated phenolic... clips are of silver-plated brass or silver alloy for better contact. • Offset inter-element construction in both rotating contacts and back-to-back terminals provides lower electrostatic capacities. • Choose from many types in this double-wipe style switch. One or more sections—a versatile multiple-section switch built to your specifications. • Ratings: 7½ amperes at 115 volts. Used up to 20 megacycles.

2. Great step forward in switching is CRL's New Rotary, Coil, Spring and Cam Index Switch. It gives you smoother action, longer life.

3. Centralab's development of a revolutionary, new Slide Switch vastly facilitates AM and FM set design! Flat, horizontal design saves valuable space, allows short leads, convenient location to coils, reduced lead inductances for increased efficiency in low and high frequencies. CRL Slide Switches are rugged and dependable.
For by-pass or coupling applications, check Centralab's original line of ceramic disc Hi-Kaps. Disc Hi-Kaps are smaller than a dime.

Hi-Vo-Kaps are filter and by-pass capacitors combining high voltage, small size and variety of terminal connections to fit most TV needs.

Ceramic Trimmers are made in five basic types. Full capacity change within 180° rotation. Spring pressure maintains constant rotor balance.

CRL's new high quality Model 2 Radiohm Controls specifically designed for TV, radio, other electronic equipment. Lower noise level, longer life.

Let Centralab's complete Radiohm line take care of your special needs. Wide range of variations: Model "R" — wire wound, 3 watts; or composition type, 1 watt. Model "E" — composition type, 1/4 watt. Direct contact, 6 resistance tapers. Model "M" — composition type, 1/2 watt.

Centralab's Ampec, above, is an integral assembly of tube sockets, capacitors, resistors and wiring combined into one miniature amplifier unit.

Couplate consists of plate and grid resistors, plate by-pass and coupling capacitors. Minimum soldered connections speed production.

This is the new CRL Vertical Integrator Network used in TV sets. Variations of this Centralab Network are available on special order.
IMPORTANT BULLETINS FOR YOUR TECHNICAL LIBRARY!

They’re factual!

Choose From This List!

Centralab Printed Electronic Circuits
999 — PENTODE COUPLATE — specialized P. E. C. coupling plate.
42-9 — FILPEC — Printed Electronic Circuit filter.

Centralab Capacitors
42-3 — BC TUBULAR HI-KAPS — capacitors for use where temperature compensation is unimportant.
42-4 — BC DISC HI-KAPS — miniature ceramic BC capacitors.
42-10 — HI-VO-KAPS — high voltage capacitors for TV application.
695 — CERAMIC TRIMMERS — CRL trimmer catalog.
981 — HI-VO-KAPS — capacitors for TV application. For jobbers.
42-18 — TC CAPACITORS — temperature compensating capacitors.
814 — CAPACITORS — high-voltage capacitors.
975 — FT HI-KAPS — feed-thru capacitors.

Centralab Switches
953 — SLIDE SWITCH — applies to AM and FM switching circuits.
970 — LEVER SWITCH — shows indexing combinations.
995 — ROTARY SWITCH — schematic application diagrams.
722 — SWITCH CATALOG — facts on CRL’s complete line of switches.

Centralab Controls
42-7 — MODEL “J” RADIOHM — world’s smallest commercially produced control.
697 — VARIABLE RESISTORS — full data on CRL Variable Resistors.

Centralab Ceramics
967 — CERAMIC CAPACITOR DIELECTRIC MATERIALS.
720 — CERAMIC CATALOG — CRL steatite, ceramic products.

General
26 — GENERAL CATALOG — Combines Centralab’s line of products for jobber, ham, experimenter, serviceman or industrial user.

Look to CENTRALAB in 1949! First in component research that means lower costs for the electronic industry. If you’re planning new equipment, let Centralab’s sales and engineering service work with you. For complete information on all CRL products, get in touch with your Centralab Representative. Or write direct.

CENTRALAB
Division of Globe-Union Inc.
900 East Keefe Avenue, Milwaukee, Wisconsin

Yes—I would like to have the CRL bulletins, checked below, for my technical library!

☐ 973  ☐ 42-9  ☐ 42-18  ☐ 953  ☐ 42-10  ☐ 722  ☐ 970  ☐ 42-7  ☐ 697  ☐ 995
☐ 42-6  ☐ 42-3  ☐ 695  ☐ 814  ☐ 970  ☐ 42-7  ☐ 26
☐ 999  ☐ 42-4  ☐ 981  ☐ 975  ☐ 995  ☐ 697  ☐ 967

Name

Address

City  State

TEAR OUT COUPON

for the Bulletins you want
"An excellent job well done"
says KOCY-FM, Oklahoma City

"Yesterday's 85-mile-an-hour wind speaks well of the ruggedness of our new Truscon Tower", continues a letter from M. H. Bonebrake, general manager of this important Mutual Network member, to Truscon's Oklahoma City District Manager. "Your design is serving our purpose excellently and also makes a beautiful tower."

Including the General Electric 8-bay circular FM antenna and its beacon, this Truscon Guyed Radio Tower rises 938 feet above the Oklahoma plain. Yet it stands strong, slender and sure in the face of high velocity winds, and delivers the KOCY-FM 176 kilowatt signal on a frequency of 94.7 megacycles without interruption.

This sincere tribute is evidence of Truscon engineering and construction skills in assuring AM, FM and TV Tower dependability. Whether your operations call for tall or small towers . . . guyed or self-supporting . . . tapered or uniform cross-section . . . contact your nearby Truscon District Office . . . or our home office in Youngstown—for expert assistance without obligation.

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

TRUSCON
SELF-SUPPORTING
AND UNIFORM CROSS SECTION GUYED TOWERS
Can you afford a switch that isn’t dependable?

No?

Then Specify MALLORY!

ENGINEERING DATA SHEETS
Send for the Mallory Engineering Data Sheets on the RS series. They contain complete specifications for available circuit combinations with respective terminal locations, dimensional drawings — everything the engineer needs.

SPECIFICATION SHEETS
Specification sheets for all RS switches have also been prepared. These sheets are printed on thin paper to permit blueprinting. The sectional drawings indicate standard and optional dimensions — make it easy for you to order production samples built to your requirements.

There is a Mallory switch to fit your design — write for further details.

Design engineers who specify Mallory RS switches know they are getting the best that substantial construction and precision manufacturing can produce. They know that Mallory RS switches protect their good name because they provide maximum long-life and efficient dependable service.

Mallory RS switches are available with cam and ball type index assembly, or with positive indexing hill-and-valley double roller type index assembly. These are the features that make Mallory switches famous for dependability and quality. All are advantages of extreme importance in television and high frequency applications where stability is essential.

- Insulation of high-grade, low-loss laminated phenolic.
- Terminals and contacts of special Mallory spring alloy, heavily silver-plated to insure long life at low contact resistance.
- Terminals held securely by exclusive Mallory two-point fastening — heavy staples prevent loosening or twisting.
- Double wiping action on contacts with an inherent flexing feature — insures good electrical contact with the rotor shoes throughout rotation.
- Six rotor supports on the stator — insures accurate alignment.
- Brass rotor shoes, heavily silver-plated — insures low contact resistance.
- All shoes held flat and securely to phenolic rotor by rivets — prevents stubbing — insures smooth rotation — minimum of noise in critical circuits.

Precision Electronic Parts — Switches, Controls, Resistors

SERVING INDUSTRY WITH
- Capacitors
- Rectifiers
- Contacts
- Switches
- Controls
- Vibrators
- Power Supplies
- Resistance Welding Materials

P. R. MALLORY & Co., Inc.
P. R. MALLORY & Co., Inc., INDIANAPOLIS 6, INDIANA
The Successor to the World-Famous Type 208...

The New TYPE 304 and TYPE 304-H Cathode-ray OSCILLOGRAPHS

Allen B. Du Mont Laboratories, Inc.
Instrument Division
The New TYPES 304 and 304-H Have All the Features That Made the Type 208 the Most Popular Oscillograph in the World... PLUS MANY MORE... AT NO EXTRA COST!

HIGH-GAIN A-C & D-C AMPLIFIERS FOR BOTH X- & Y-AXES

- Sensitivity:
  - Y-Axis — 10 millivolts rms per inch (ac and dc)
  - X-Axis — 50 millivolts rms per inch (ac and dc)

- Frequency Response:
  - D-C amp. X and Y axes — 0.100.000 cps within ±10%
  - A-C amp. X and Y axes — 20.100.000 cps within ±10%

- No pattern "pop" even with high-gain amplifiers.
- Excellent stability and minimum microphonics and drift.
- Provision for applying signals directly to deflection plates.

EXPANSION OF DETAILS

- Due to the fact that deflection of over 5 times full screen diameter is available on both the X and Y Axes, performance equivalent to that of a 25-inch cathode-ray tube is possible, with the high resolution of d-C amplifiers.
- X-axis positioning is available over this entire range on both axes, with no on-screen distortion present. (See Figures 3 and 4).

RECURRENT AND DRIVEN SWEEPS

- Recurrent and driven sweeps variable from 9 to 30,000 cps.
- Sweep speeds faster than 0.75 inch/sec. with fully expanded time base.
- Provision incorporated for sweeps of 10 seconds and slower through the use of external capacitors at front-panel terminals. (See Figure 7).
- Sync amplifier with sync-polarity selection is provided.

STABILIZED SYNCHRONIZATION

- Sync limiting provided so that sweep length and synchronization are maintained as signal level varies. (See Figure 6).

INTENSITY MODULATION

- 2-Axis input terminal on the front panel capacitively coupled to the grid of the cathode-ray tube. (See Figure 5).
- 15 volts peak will blank trace fully at normal intensity.

INCREASED ACCELERATING POTENTIAL

- Du Mont Type 5CP-A Cathode-ray Tube operated at an accelerating potential of 1760 volts, 380 volts higher than that of the Type 208.
- Type 304-H also available with an overall accelerating potential of 3000 volts. This higher accelerating potential facilitates the use of long-persistence screens so that fullest possible advantage may be taken of the low-frequency sweeps, the fast driven sweeps, and the d-C amplifiers. (See Figures 8 and 9).

ADDITIONAL FEATURES

- An engraved, permanently mounted, calibrated scale greatly facilitates quantitative measurements.
- Metal magnetic shield affords maximum protection of the cathode-ray tube from the effects of external magnetic fields.
- Du Mont Type 2501 Bezel incorporated for attaching such accessories as the Du Mont Type 271-A or 314-A Oscillograph-record Camera.

MECHANICAL CHARACTERISTICS

- In all dimensions, the Type 304 is smaller than the Type 208.
- Height — 13½ inches
- Width — 8½ inches
- Weight — 50 pounds
- Panel is reverse-etched, white on gray.

TRIED AND PROVED

- The NEW Du Mont Type 304 Cathode-ray Oscillograph has undergone a most rigid field test, both in Du Mont Laboratories and in selected laboratories and institutions throughout the country. Here in a great variety of applications, every feature of the Type 304 was given a thorough workout. Thus, Du Mont presents the Type 304, not as a new instrument of unknown quality, but as an oscillograph of TRIED AND PROVED EXCELLENCE!

<table>
<thead>
<tr>
<th>Type</th>
<th>Price</th>
</tr>
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<tbody>
<tr>
<td>304</td>
<td>$285.00</td>
</tr>
<tr>
<td>304-H</td>
<td>$307.50</td>
</tr>
</tbody>
</table>

For additional information concerning the Du Mont Types 304 and 304-H, request Bulletin 304 from the Instrument Division.
ample kiln capacity

safeguards AlSiMag quality and helps keep deliveries on schedule

Side view of one of AlSiMag's tunnel kilns. All kilns, both circular and tunnel, are handled from one centralized control panel.

- Completely automatic controls hold firing temperatures within $\pm 2^\circ$ C in AlSiMag's kilns. As an extra safeguard, highly trained and skilled kiln operators are on duty every minute of the day and night. Recording instruments plus operator's hourly checks and records assure that all AlSiMag material is accurately fired.

AMERICAN LAVA CORPORATION

48TH YEAR OF CERAMIC LEADERSHIP

CHATTANOOGA 5, TENNESSEE
New Audio Oscillator

The new Model 50 audio oscillator, intended for use as a secondary standard for low frequency application, has been designed and developed by C. G. S. Laboratories, Inc. 36 Ludlow St. Stamford, Conn.

This oscillator operates in the range from 2,500 to 25,000 cps, accuracy ±0.1 per cent. Tube change and ±10 per cent line voltage change affect the frequency less than ±0.03 per cent. Temperature coefficient is of the order of 0.002 per cent per degree centigrade.

Detailed information is available from H. Langstroth at C. G. S.

Polinear Recorder

The model PFR Polinear Recorder, which offers the combined ability of polar and rectilinear movement permitting the recording of angular patterns, frequency response characteristics, and other measurements, has been developed by Sound Apparatus Co. Stirling, N. J.

New Sweep Generator

An entirely electronic sweep generator, Type ST-4A, using a variable-permeability type sweep and having no moving components, has been developed by the Specialty Div., General Electric Co., Electronics Park, Syracuse, N. Y.

Frequency is continuously variable from 4 to 110 Mc and from 170 to 220 Mc with a linear sweep width of from 500 kc to greater than 15 Mc. High output voltage is available over the entire range.

Attenuation is continuously variable from maximum output down to 20 microvolts. Leakage is low, it is claimed, with stray fields of less than 10 microvolts, induced in a 2-inch loop 6 inches from the case in any direction.

Further information on this sweep generator is available from the Specialty Division.

Rhombic Antenna Terminating Resistor

A new rhombic antenna terminating resistor #079 which is two noninductive Ayrton-Perry wound 362.5 ohm resistors enclosed in glazed ceramic shell and vacuum sealed, is available to the industry from Shallcross Mfg. Co., 520 Pusey Ave., Collingdale, Pa.

Leads are brought out to terminal eyes which are designed for 7-strand #16 antenna wire. These resistors are rated at 25 watts dissipation and are designed to operate between -20°F to +110°F. Over-all length 6 1/4", diameter 2 1/2".

Nuclear Radiation Counter

A beta-gamma counting rate meter, with the counter located in the probe, has been developed by the Instrument Div., Kelly-Koet Mfg. Co., 12 E. Sixth St., Covington, Ky.

Described as Kelkelet Model K-800, this meter is housed in a cast magnesium-aluminum case, and is waterproof. The probe has a movable shield to differentiate between beta and gamma radiation. This has been accomplished without conventional interdispersed foils. The shield is stainless steel, 2 millimeters thick, and will exclude all but very high beta rays, the manufacturer claims.

Three ranges of gamma activity may be measured, 0.2, 2.0, and 20.0 mr/HR. The scale is also calibrated in 360, 3,600 and 36,000 counts per minute.

Small Disk Ceramic Capacitors

New thin disk ceramic capacitors are the latest addition to the line of fixed capacitors manufactured by Sprague Electric Co., North Adams, Mass.

The manufacturer asserts that these capacitors consist of a dime-sized ceramic plate of high dielectric constant with silvered electrodes fired on both faces of the disk. The leads are soldered to the silvering and the capacitors are coated with resin.

These components are available in ratings up to 0.01 or 2 X 0.004uf, 500 volts dc working.

Engineering Bulletin 601A supplies complete data.
The MOST VERSATILE AND SENSITIVE Oscilloscope EVER Built!

Some of the outstanding advantages of the...

NEW LAVOIE LA-239A OSCILLOSCOPE

1. Takes the guesswork out of pulse techniques.
2. Accurately measures amplitude, width, separation, repetition rate and rise time without the need of additional equipment.
3. Accurate timing markers provide means of calibrating the linear time base.
4. Internal trigger generator permits pulse generator and oscilloscope to be triggered simultaneously, while sweep delay circuit allows a small portion of image to be expanded TEN TIMES normal size.

INCREASED PRODUCTION NOW PERMITS A REDUCTION OVER FORMER LIST PRICE WITH SPECIAL REDUCTIONS TO TECHNICAL SCHOOLS AND NON-PROFIT ORGANIZATIONS

Write for Technical Bulletin LA-239A giving complete detailed information.

LaVoie Laboratories
Radio Engineers and Manufacturers
Morganville, N. J.

Specialists in the Development and Manufacture of UHF Equipment
Now you can work with

REMALLOY
Permanent Magnet Material
(Manufactured under license from Western Electric Company)

It’s fully available for the first time

How you can get it!

ARNOLD can supply REMALLOY in the form of BARS and CASTINGS or SINTERED TO SPECIAL SHAPES

How you can use it!

REMALLOY generally may be used instead of 36-41% Cobalt Permanent Magnet Steel—replacing it without design changes, and at a cost saving.

In addition to our customary production of all types of ALNICO and other permanent magnet materials, we now produce REMALLOY. The various forms in which it is available—bars, castings or sintered shapes—are all produced under the Arnold methods of 100% quality control; and can be supplied to you either in rough or semi-finished condition, or as completely finished units ready for assembly. Let us help you secure the cost-saving advantages of REMALLOY in your designs. Call or write for further data, or for engineering assistance.

The Arnold Engineering Company
SUBSIDIARY OF ALLEGHENY LUDLUM STEEL CORPORATION
147 East Ontario Street, Chicago 11, Illinois

Specialists and Leaders in the Design, Engineering and Manufacture of Permanent Magnets

ARNOLD
GENERAL ELECTRIC COMPANY pioneered the broad-band gas switching tube for microwave applications. From G-E research laboratories and drawing-boards came the original plans for these r-f "traffic sentinels" whose instant and automatic operation makes possible modern radar for military purposes—for electronic navigation in fog and darkness—for airway scanning, airport traffic control, and cloud and weather study.

Now G.E. offers to equipment designers and users a group of highly developed TR, ATR, and PRE-TR types which reflect intensive effort to achieve still more efficient tube-switching in microwave work.

Key ratings are given below. Complete characteristics and performance data gladly will be supplied at your request, covering any or all of the tubes listed. Announcement of still other types later, may be expected in view of General Electric's continuing program in the field.

For information, prices, and the help of specialist tube engineers who gladly will cooperate in choosing the right tubes for your microwave circuits, wire or write General Electric Company, Electronics Department, Schenectady 5, New York.

<table>
<thead>
<tr>
<th>Group</th>
<th>Type No.</th>
<th>Freq. range</th>
<th>Max peak power</th>
<th>Leakage power</th>
<th>Recovery time, max</th>
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<td>8490-9578 mc</td>
<td>250 kw</td>
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<td>4 mu sec @-3 db</td>
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<td>2700-2910 mc</td>
<td>1000 kw</td>
<td>100 kw</td>
<td>.0002 joules</td>
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</table>

GENERAL ELECTRIC

FIRST AND GREATEST NAME IN ELECTRONICS
REVERE FREE-CUTTING COPPER ROD

... INCREASES ELECTRONIC PRODUCTION

Since its introduction, Revere Free-Cutting Copper has decisively proved its great value for the precision manufacture of copper parts. Uses include certain tube elements requiring both great dimensional precision, and exceptional finish. It is also being used for switch gear, high-capacity plug connectors and in similar applications requiring copper to be machined with great accuracy and smoothness. This copper may also be cold-upset to a considerable deformation, and may be hot forged.

Revere Free-Cutting Copper is oxygen-free, high conductivity, and contains a small amount of tellurium, which, plus special processing in the Revere mills, greatly increases machining speeds, makes possible closer tolerances and much smoother finish.

Thus production is increased, costs are cut, rejects lessened. The material’s one important limitation is that it does not make a vacuum-tight seal with glass. In all other electronic applications this special-quality material offers great advantages. Write Revere for details.

REVERE COPPER AND BRASS INCORPORATED

Founded by Paul Revere in 1801

Executive Offices: 230 Park Avenue
New York 17, New York


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 3-1/2" 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.)
El-Menco Capacitors, mighty midgets of electronics, are so tiny you can hold a dozen of them in the hollow of your hand. Yet, their capacity, strength, long life and dependable performance under the most critical conditions are recognized throughout the industry.

The next time you need capacitors, call for El-Menco. Order the capacitors that are small in size but mighty in performance.

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

El-Menco Capacitors, mighty midgets of electronics, are so tiny you can hold a dozen of them in the hollow of your hand. Yet, their capacity, strength, long life and dependable performance under the most critical conditions are recognized throughout the industry.

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The next time you need capacitors, call for El-Menco. Order the capacitors that are small in size but mighty in performance.

THE ELECTRO MOTIVE MFG. CO., Inc.
WILLIMANTIC CONNECTICUT
**FIXED RESISTORS**

Bradley units will carry 100% load for 1,000 hours . . . at 70C ambient temperature with a resistance change of less than 5%. In standard R.M.A. values from 10 ohms to 22 megohms, except 1-watt unit available from 2.7 ohms to 22 megohms.

**ADJUSTABLE RESISTORS**

Type J Bradleyometers are rated at 2 watts with a big safety factor. The solid-molded resistor unit is not affected by heat, cold, moisture, or wear. Can be furnished with line switch. Available in single, dual, and triple-unit designs.

For circuits that require resistors of unsurpassed quality...

Specify Allen-Bradley

**BRADLEY UNITS** are available in ½, 1, and 2-watt ratings. They have high mechanical strength and permanent electrical characteristics.

The leads are differentially tempered to prevent sharp bends near the resistor. The leads are easily formed to fit any spot.

All Bradley units are packed in convenient honeycomb cartons that keep the leads straight. Send for Allen-Bradley resistor chart.

**TYPE J BRADLEYOMETERS** have solid-molded resistor elements. They are thick rings molded to provide any resistance-rotation curve. After molding, heat, cold, moisture, and hard use do not affect the resistor.

The resistor is molded as a single unit with insulation, terminals, face plate, and treadsed bushing in ONE piece. There are no rivets, nor welded or soldered connections.

Allen-Bradley Co., 114 W. Greenfield Ave., Milwaukee 4, Wis.
another **Dumont** first...

**19” Metal Cone TELETRON**

**Type 19AP4**

- FOR a direct-viewing 203 square-inch picture
- FOR superlative brilliance and definition
- FOR shorter tube length requiring smaller cabinet
- FOR exclusive bent-gun construction — sharper focus — no ion-spots

Literature and quotations on request.

© Allen B. Du Mont Laboratories, Inc.

**Dumont Teletrons**

FIRST WITH THE FINEST IN T-V TUBES

ALLEN B. DU MONT LABORATORIES, INC. • TUBE DIVISION • PASSAIC, NEW JERSEY

PROCEEDINGS OF THE I.R.E. October, 1949
Pioneer in the commercial development of crystal diodes, Sylvania Electric is today continuing its scientific research to advance the understanding and applications of semi-conductors.

An example of this research is taking place in the Sylvania physics laboratory at Bayside, N. Y. In the photo, a laboratory worker adjusts contacts on an experimental germanium tetrode for use as a mixer at input frequencies far beyond 5 mc. A typical mixer with an IF of 600 ke can show a conversion voltage gain of 10 or corresponding to a power gain of +1 db. at a frequency of 150 mc.

These are only experimental results, yet they promise that further investigation will be very much worthwhile. Such research is typical of the work Sylvania is doing and will continue to do, to produce new and better products.

SYLVANIA ELECTRIC

EXPERIMENTAL CRYSTAL TETRODE MIXER provides good conversion gain plus high conversion transconductance

S Y N N I A E L E C    I R I C

ELECTRONIC DEVICES; RADIO TUBES; CATHODE RAY TUBES; FLUORESCENT LAMPS, FIXTURES, WIRING DEVICES, SIGN TUBING: LIGHT BULBS; PHOTOLAMPS
Where high-current, non-shorting tap switches are required, scores of equipment manufacturers prefer Ohmite over all others...

Because Ohmite tap switches combine high-current capacity and a large number of taps with unusual compactness...

Because their sturdy one-piece ceramic bodies provide permanent non-arcing insulation...

Because their heavy silver-to-silver contacts have a self-cleaning action... and provide continuous, dependable contact with low resistance...

Because their cam-and-roller mechanism has a positive "slow-break, quick-make" action—particularly suited for alternating current.

That's why more Ohmite high-current tap switches are purchased than all other makes combined... and why it will pay you to standardize on Ohmite in your product.

In addition to the types and sizes illustrated, Ohmite tap switches are supplied in open, all-ceramic, shorting and non-shorting types. All

**Compact**

**Convenient**

**Dependable**

<table>
<thead>
<tr>
<th>AMPS.</th>
<th>MODEL No.</th>
<th>MAX. V. (A-C)</th>
<th>No. TAPS</th>
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<tr>
<td>10</td>
<td>111</td>
<td>150</td>
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<td>25</td>
<td>312</td>
<td>300*</td>
<td>2 to 12</td>
</tr>
<tr>
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<td>412</td>
<td>300*</td>
<td>2 to 12</td>
</tr>
<tr>
<td>100</td>
<td>608</td>
<td>300</td>
<td>2 to 8</td>
</tr>
</tbody>
</table>

*150 volts between taps

Ohmite tap switches can be mounted in tandem for multiple-pole operation.

Write on company letterhead for Ohmite Catalog and Engineering Manual No. 40.

**Be Right with...**

**Ohmite**

RHEOSTATS • RESISTORS

TAP SWITCHES

Industry's first choice
No Tube Trouble

If you are one of the many owners of FM transmitters where tube replacement cost has been heavily draining the reserve bank account, you will be particularly interested in the Gates BF-3D FM transmitter for 3000 watts power. The highly vulnerable power amplifier tubes which can be quickly damaged by changes in antenna characteristics, improper air circulation around the tubes and in some instances even low line voltage, have been engineered not only to good performance but to low maintenance cost.

On the attached brochure note the unique tank circuit design where the new 4-1000 power amplifier tubes are covered with a pyrex jacket which confines all of the air around the tube and finally concentrates it on the important end seal. Broadcasters are reporting from 2500 to 3000 hours of tube life and many purchasers of the BF-3D transmitter have the original set of tubes in the sockets after many months of use. To aid long tube life is a scientific air pressure control that immediately disconnects the plate voltage control that immediately disconnects the plate voltage control that terminates the problem of compressing an air flow. Also a direct reading power and standing wave ratio indicator which tells the operator instantly if the antenna characteristics have changed because of icing or other reasons and is placing a heavy load on the power amplifier tubes.

The BF-3D, like all Gates products, is engineered not only for fine performance, meeting rigid FM requirements, but having the practical touch added by such things as longer tube life that means dollars to the broadcaster.

Further information about this fine transmitter that cannot be found in the attached brochure will gladly be given upon request.

Yours very truly,
GATES RADIO COMPANY
Sales Department

Commercially proven... the Eimac 4-1000A is an outstanding high-power tetrode. Its rugged construction and stability of performance enable the country's leading transmitter manufacturers to enthusiastically expound the tubes' advantages in their key socket positions.

Consider the Eimac 4-1000A tetrode for your high-power equipment... frequency limits are well into the vhf. Complete data is available, please write direct.

EITEL-McCULLOUGH INC.
San Bruno, California

Export Agents: Fraser & Hansen, 301 Clay St., San Francisco, California

Follow the Leaders to

EIMAC TUBES

*This letter was distributed with a brochure on the popular Gates BF-3D, 3KW FM transmitter.
Here's why......

the new series 300

**AMPEX**

MAGNETIC TAPE RECORDER

answers industry need!

*Original program quality preserved*

Use of independent reproduction facilities allows instantaneous monitoring and makes possible the most stringent comparisons between recordings and originals.

*Tape and playback noise non-existent*

Use of special record and bias circuits has eliminated tape noise.*

*Editing made easy*

With Ampex editing is almost instantaneous. Single letters have been actually cut off the end of words. Scissors and Scotch tape are all the tools needed.

*You can depend on Ampex*

Read what Frank Marx, Vice President in charge of Engineering, American Broadcasting Company, says: "For the past two years A.B.C. has successfully used magnetic tape for rebroadcast purposes...A.B.C. recorded on AMPEX in Chicago...17 hours per day. For 2618 hours of playback time, the air time lost was less than 3 minutes: a truly remarkable record."

Console Model 300* $1,573.75

Portable Model 300 $1,594.41

Rack Mounted $1,491.75

*F. O. B. Factory, San Carlos, Calif.*

**SPECIFICATIONS**

**FREQUENCY RESPONSE:**
At 15" ± 2 db, 50—15,000 cycles
At 7.5" ± 2 db, 50—7,500 cycles

**SIGNAL-TO-NOISE RATIO:** The overall unweighted system noise is 70 db. below tape saturation, and over 60 db. below 3% total harmonic distortion at 400 cycles.

**STARTING TIME:** Instantaneous.
(When starting in the Normal Play mode of operation, the tape is up to full speed in less than .1 second.)

**FLUTTER AND WOOF:** At 15 inches per second, well under 0.1% r.m.s., measuring all flutter components from 0 to 300 cycles, using a tone of 3000 cycles. At 7.5 inches, under .2%.

Manufactured by Ampex Electric Corporation, San Carlos, Calif.

DISTRIBUTED BY

BING CROSBY ENTERPRISES

AUDI0 & VIDEO PRODUCTS CORPORATION

GRAYBAR ELECTRIC CO. INC.

9028 Sunset Blvd., Hollywood 46, Calif.

1650 Broadway, New York, New York

420 Lexington Ave., New York 17, N.Y.

**PROCEEDINGS OF THE I.R.E.** October, 1949

**29A**
Here are some of the many reasons why there are more Simpson 260 high sensitivity volt-ohm-milliammeters in use today than all others combined. The Simpson 260 has earned world-wide acceptance because it was the first tester of its kind with all these "Firsts":

Simpson 260 SET TESTER
WORLD FAMOUS FOR ALL THESE "FIRSTS"

- First high sensitivity instrument to use a metal armature frame.
- First to use fully enclosed dust proof rotary switch with all contacts molded in place accurately and firmly.
- First to do away with harness wiring.
- First to provide separate molded recesses for resistors, batteries, etc.
- First to cover all resistors to prevent shorts and accidental damage and to protect against dust and dirt.
- First with a sturdy movement adapted to the rugged requirements of a wide range of service work or laboratory testing.
- First to provide easy means of replacing batteries.
- First to use all bakelite case and panels in volt-ohm-milliammeters.
- First volt-ohm-milliammeter at 20,000 ohms per volt with large 4½" meter supplied in compact case (size 5½" x 7" x 3½").
- First and only one available with Simpson patented Roll Top Case.
- First to provide convenient compartment for test leads (Roll Top case).
- First to offer choice of colors.

The Model 260 also is available in the famous patented Roll Top safety case with built-in lead compartment. This sturdy, molded, bakelite case with Roll Top provides maximum protection for your 260 when used for servicing in the field or shop.

25,000 volt DC Probe for television servicing, complete, for use with 260, $12.85

SIMPSON ELECTRIC COMPANY • 5200-18 W. Kinzie St., Chicago 44, Ill. • In Canada: Bach-Simpson, Ltd., London, Ont.

RANGES
20,000 Ohms per Volt DC, 1,000 Ohms per Volt AC
VOLTS: AC & DC—2.5, 10, 50, 250, 1,000, 5,000
OUTPUT: 2.5, 10, 50, 250, 1,000
MILLIAMPERES, DC: 10, 100, 500
MICROAMPERES, DC: 100
AMPERES, DC: 10
DEcIBELS: (5 ranges)—12 to +55 DB
OHMS: 0-2,000 (12 ohms center), 0-200,000 (1200 ohms center), 0-20 megohms (120,000 ohms center).

Prices: $38.95 dealers net; Roll Top $45.95 dealers net.
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HARRISON, N. J.
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(Including the WAVES AND ELECTRONS Section)

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Donald G. Fink

DIRECTOR-AT-LARGE, 1949–1951

Donald Glen Fink was born on November 8, 1911, in Englewood, New Jersey. He was graduated in 1933 from Massachusetts Institute of Technology with the B.Sc. degree in electrical communications. After a year as a research assistant on the staff of the departments of geology and electrical engineering at MIT, Mr. Fink joined the staff of the journal Electronics, as editorial assistant. In 1937 he became managing editor; in 1945, executive editor; and in 1946, editor-in-chief. In 1943 he was awarded the degree of M.Sc. in electrical engineering by Columbia University.

Obtaining a leave of absence from his editorial duties in 1941, Mr. Fink became a member of the staff of the radiation laboratory at MIT where, in 1943, he headed the loran division. He then transferred to the Office of the Secretary of War as an expert consultant on radio navigation and radar. During his war service Mr. Fink travelled over 80,000 miles from Cairo, Egypt, to Darwin, Australia, siting loran stations and arranging for the use of the loran system by the allied forces. In 1946 he participated in the atom bomb tests at Bikini, as a civilian consultant on the staff of Admiral Blandy.

Mr. Fink is the author of numerous books, including “Engineering Electronics,” “Principles of Television Engineering,” “Television Standards and Practice,” “Microwave Radar,” and “Radar Engineering.” As editor of Proceedings of the National Television System Committee, member of the Television Panel of the Radio Technical Planning Board, and currently as chairman of the Joint Technical Advisory Committee, he is active in standardization work, particularly in the field of television. In 1948 he was chairman of the IRE Television System Committee. He is a member of the Committee on Navigation of the Research and Development Board. Mr. Fink is a Fellow of the Institute and a member of Tau Beta Pi, Sigma Xi and Eta Kappa Nu.
The technical and industrial potentials of a modern country are vital factors in its military strength. Accordingly, increasingly extensive and close co-ordination between engineers and the Armed Forces is a timely matter, and one of major importance. It is ably discussed in the following guest editorial by a radio development engineer of the Bell Telephone Laboratories, who is a Senior Member of The Institute of Radio Engineers.—The Editor.

The Radio Engineer and the Armed Services

W. C. TINUS

A great deal has been written during the past few years about the rapid growth of the radio and electronic engineering profession from adolescence to mature professional standing. This growth is particularly evident in the changed relation between radio engineers and the armed services. Radio engineers in industrial firms merit, and receive, a great deal more confidence than they did before the war from those in the services who are responsible for the development of new military electronic devices.

Not many years ago the first contact a typical industrial radio engineer had with a military electronic problem was a detailed equipment specification telling what was wanted and what the equipment was expected to do in its individual tests. The last contact the engineer had with the problem was when the equipment was designed, built, and made to pass its tests. What the military problem was and whether the specified equipment best solved it was the concern of others.

The situation today is very different. The military problems for which electronics may provide a solution are more complex. The engineer is on the job at a much earlier stage. His vision is broader and his opinion more respected. He frequently has the opportunity not only to become thoroughly acquainted with the basic military problem but also to help the military people define it. Many equipment design contracts are now preceded by study contracts in which the problem is defined and a solution is proposed. The solution may employ any of the multitude of things in electronics' bags of tricks.

These changes are all to the good. They are good for the military in getting more technical people studying their problems, and eventually in getting better solutions to them. They are good for radio engineers in enlarging their horizons of thought and their opportunities to apply their specialized knowledge. They are good for all of us in getting more value received for our money spent on national defense.
Audio-Frequency Measurements*

W. L. BLACK†, SENIOR MEMBER, IRE, AND H. H. SCOTT‡, SENIOR MEMBER, IRE

Summary—This paper indicates the theory involved in making measurements of gain, frequency response, distortion, and noise at audio frequencies, with particular emphasis on such measurements made on high-gain systems. There are also discussed techniques of measurement and factors affecting the accuracy of results. This subject is not new art but has not previously been published in correlated form, to the knowledge of the authors.

INTRODUCTION

During the last several years the Radio Manufacturers Association has given considerable attention to codifying minimum standards of performance for the major components of radio broadcasting systems. The engineering aspects of this subject have been considered by committees in the Transmitter Section of the Engineering Department of the Association.

Definitions and minimum standards for the audio facilities of a radio broadcasting system considered primarily as a complete electrical system have been issued as an RMA standard.1 (Subscripts refer to numbered Bibliography on pages 1114-1115.) As defined in this standard, audio facilities comprise "all audio facilities from the input terminals of the microphone preamplifier to the input terminals of the main transmitter, excluding the studio-transmitter link which may be either wire line or radio. No pre-emphasis is included in the audio facilities."

Subsequently the RMA undertook to outline methods of measurement of audio facilities to insure comparable results from measurements, made at different times and on different apparatus, consistent with practical instrumentation. It is the purpose of this paper to summarize the technical background which is the basis for the RMA standardization activity and to outline possible pitfalls in making measurements on complex, high-gain audio systems. Emphasis is placed on system measurements, as ordinarily a complex, high-gain system presents more of a measurement problem than do system components.

As a practical matter, the important characteristics of audio systems for radio broadcasting which have been agreed upon for RMA standardization are: gain, frequency response (relative gain over a frequency range), single frequency harmonic distortion, and noise (in the sense of noise being extraneous sound or corresponding electrical energy tending to interfere with the proper and easy perception of desired sounds or their equivalent electrical waves).

Gain

Gain for the determination of performance characteristics is measured between resistances equal to the rated source and load impedances. For the purpose of measuring such gain (or loss) the concept of the ideal transformer4,5,13 for impedance translation for optimum power transfer in the reference condition is useful. This is shown in Fig. 1. Then the gain is the ratio of the power in the load with the equipment under test in the circuit to the power delivered to the same load without the equipment under test, but including an ideal transformer. It is obvious that any further discussion of gain will refer with equal validity to frequency response, as the latter is relative gain over the specified frequency range.

Measurements using resistance terminations for the measurement of the component parts of a system may not agree in total with the results of over-all measurement of the system between resistances. Ordinarily the disagreement is, however, of relatively small magnitude.

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when the interrelations of the impedances of the components have been considered in conjunction with the system design. For large variations from rated impedances, distortion of gain-versus-frequency characteristics may be sufficient to require consideration of the coupling factors involved and may indicate corrective equalization.

**Practical Test Circuit**

For practical tests, Fig. 2 presents an essentially complete diagram showing all circuit elements which would ordinarily be required for audio-frequency measurements. In many cases, several of the units may be combined in one physical instrument. Separate consideration of each of the various elements follows.

**Oscillator**

The source of power for gain testing is ordinarily an audio-frequency oscillator. This oscillator should cover the range of frequencies involved in the tests. Ordinarily absolute frequency accuracy and stability with time are not practical limitations, except in the measurement of frequency selective devices such as filters, and in the use of some types of distortion measuring instruments containing sharply tuned rejection circuits. However, freedom from drift of output voltage with time and with variation in oscillator power supply voltage are of practical convenience. Except when distortion measurements are made, distortion and noise components ordinarily contributed by present-day oscillators are not limiting unless the equipment under test includes filters having sharp cutoff characteristics.

Some oscillators may be critical as to load impedance. Load impedance correction is ordinarily accomplished by a shunt or a series resistance, depending upon whether the optimum load for the oscillator is lower or higher respectively than that offered by the measuring circuit. It will also often be necessary to include a transformer having an electrostatic shield between windings between the oscillator output and the input of the testing circuit to control ground connections and parasitic coupling between the oscillator and the remainder of the testing system.

In any event, once the circuit at the output of the oscillator has been determined, the oscillator output circuit does not affect the measurements of gain or of response as a constant output voltage is maintained and the oscillator is then the equivalent of a zero internal impedance generator.

**Terminations**

The desired objective of comparable results indicates the use of pure resistances for terminations. Such terminations include that for the output of the equipment under test, as well as any used for input termination including that used for equivalent generator internal impedance. The impedance of the resistors used for such terminations may therefore warrant consideration. Other factors in this connection include the absolute value of output terminating resistors and the stability of such resistors with temperature, particularly when the output distortion of a power amplifier is being measured.

**Meters**

The input voltmeter may be of the vacuum tube, rectifier, or thermocouple type, or may be a standard volume indicator. This meter may be rms, average, or peak reading, except as noted in specific instances later. The stability of this meter must be such that its readings are not influenced by extraneous factors, such as power supply to a vacuum tube.

If a rectifier meter is used, its possible introduction of extraneous modulation products, particularly when making distortion measurements, as well as its possible variation in impedance with change in input are limiting factors. The loss introduced in the transmission circuit by the ordinary meters of this type (of which the "standard volume indicator" is a special form) must be taken into account, particularly if the meter is alternately connected and disconnected during a test.

A thermocouple meter may be used, provided it is appropriately connected to the circuit so that its internal resistance is taken into account. Suitably calibrated thermocouple meters have good scale spread, practical freedom from frequency discrimination at audio frequencies, freedom from modulation, and adequate accuracy. The disadvantages are relatively slow speed of operation and danger of damage due to excessive current.

In this general connection, the indication of a rectifier type meter, typified by one using a copper oxide rectifier, approximates the average value of a sine wave while a thermocouple meter indicates the effective or
The output meter may be similar to the input meter or may actually be the same meter alternately connected at the input and at the output. If the other components of the measuring system are so arranged that the input and output voltages are the same, the measurements are expedited and errors due to absolute calibration of the meter and to its variation in indication with frequency are eliminated in making response measurements. However, convenient switching on this basis involves bringing wires from the input and the output of the system in close proximity and may cause undesirable coupling, particularly when a high-gain system is being measured. This may cause an error in absolute gain indication or an error changing with frequency during response measurements, or both. This procedure is, of course, impractical if the input and output of the equipment under test are not in approximately the same vicinity. For example, the input of the system being tested might be in a studio control booth and its output in a master control room.

Other factors to consider are the frequency response of the meter, accuracy of reading possible due to instrument scale spread and pointer structure and the sensitivity of the output meter in the case of loss measurements. Finally, the factors such as pivot friction and uniformity of magnetic field ordinarily affecting the accuracy of electrical instruments should be considered in conjunction with the absolute accuracy desired.

Adjustable Attenuators

As a practical matter, greatest convenience of measurement is achieved if the meters used are held at constant readings, and the variations in gain are determined by calibrated adjustable attenuators at the input and at the output of the equipment under test as shown in Fig. 2. This method has several advantages. The meter scale range, absolute sensitivity, and possible variation in accuracy with indication are not factors in the choice of meters used. Possible observational errors due to the use of meter range switches are eliminated. The amount of calculation to obtain the desired result from observed readings is minimized. Further this arrangement has the merit that known input and output levels over a wide range of values can readily be provided. This is of particular practical importance if the same testing equipment is to be used to measure a wide range of gains and losses and various power levels.

Impedance Matching Networks

The calibrated adjustable attenuators will have fixed input and output impedances determined during design. If the impedances from and into which it is desired to measure the equipment under test are other than those of the attenuators, it is necessary to add impedance matching networks. These ordinarily can be fixed pads. The addition of such pads to the testing circuit is shown in Fig. 3. The determination of the values of such pads is well covered in extant literature. The increments of attenuation available and the accuracy of their calibration, together with the scale sensitivity and accuracy of the meters used, govern the accuracy with which absolute gain can be determined. Response (relative gain) measurements require absolute accuracy such that relative gains are indicated within the desired limits of precision. A total measuring error of ±0.2 db from all causes is the order of precision required for frequency response (relative gain) measurements on equipment for broadcasting transmission. This limit includes frequency response errors in the attenuators. These may ordinarily be avoided by careful attenuator design. Failing this, recourse to calibration is possible. In some instances, particularly in attenuator-network arms of low resistance values, contact resistance in switches may introduce errors. Careful design and proper maintenance of the switches will minimize this difficulty. The adjustable attenuators preferably should be mounted in a shielded enclosure. This precaution minimizes the possibility of errors due to parasitic capacities to external surroundings which become increasingly troublesome with increasing frequency.

![Fig. 3—Matching pads](image-url)

In Fig. 3 (B) and (C) are minimum loss “L” pads matching $R_A$ to $R_A$ and $R_B$ to $R_B$, respectively.

If $Z$ is the larger and $z$ is the smaller of the two impedances being matched by each matching pad, then

$$\text{loss (db)} = 20 \log_{10} \left( \sqrt{\frac{Z}{z}} + \sqrt{\frac{z}{Z}} \right)$$

The power handling capacity of the resistors used in pads may also require consideration on occasion. For example, when measurements are made on an amplifier for a high-powered sound system, an output pad might well be used to reduce the power to normal measuring level, but the difference between the power out of the amplifier and that in the meter circuit would be dissipated in the pad.
Ordinarily, one per cent accuracy will be adequate for the values of the resistors used for pads. However, the required accuracy (or alternative calibration of actual loss) will be governed by the accuracy desired in absolute gain indication. In addition, freedom from change of attenuation with frequency or calibration for such change of loss is also a necessary consideration.

In connection with the use of both adjustable attenuators and fixed pads, whether for impedance matching or to supplement limited range adjustable attenuators, it should be noted that Thevenin’s theorem may be applied to reduce the input and the output networks to the equivalent of a generator in series with a resistance.

**Balance and Grounding**

For measured results comparable to actual performance conditions, the equipment under test should be grounded in the same way for measurement as it is in actual use. In particular, the grounding or freedom from grounding of the input or output or both should be maintained. Among the practical problems which this introduces are: first, the balance of attenuators (both fixed and adjustable), and second, the control of oscillator grounding.

The attenuators, including those for impedance matching, should be of the “unbalanced” type for connection to a grounded circuit, and of the balanced type for connection to circuits intended for operation ungrounded or with a “center tap” ground. When using an unbalanced measuring circuit one side should always be grounded. The term “ground” is here used as meaning that zero potential plane used for reference. The impedance to ground for all points on the grounded side should be as low as possible, the interconnecting leads should be as short as possible, and no series resistors included in the ground circuits. For example, matching pads should have their series resistors in the ungrounded side of the circuit. A properly balanced attenuator requires both resistance balance between the two sides of the attenuator and capacity balance to ground on the two sides. The latter is ordinarily provided by the proper design of the calibrated adjustable attenuator switches. A ground at the center point of the shunt resistance element of the attenuator is usually necessary to insure balance in actual use. Tests to determine the adequacy of the balance (assuming freedom from parasitic coupling) may be made by reversing the connections either to the input or to the output of the attenuator in question and observing whether the output is affected due to such reversal, particularly at the higher audio frequencies. An alternative arrangement to the use of balanced attenuators is the use of an electrostatically shielded and balanced transformer between the attenuators and the equipment under test as shown in Fig. 4. This necessitates taking into account the loss and the possible frequency discrimination introduced by such a coil.

One practical check of the stability of the measuring circuit is to make a known change in the output level from the oscillator with a corresponding compensating change in the input attenuator setting. For example, a 10-db increase in test signal level and a 10-db increase in the input attenuator should result in the same input level to the equipment under test, and consequently no change in observed output level. This check should preferably be made at a frequency in the higher range of the frequencies to be used for testing.

The oscillator output circuit may be grounded, or even though not grounded, may have sufficient capacitance to ground to introduce errors (the usual practical case). Therefore, this circuit must also be considered in conjunction with the input ground. This condition can often be provided for by the use of an electrostatically shielded transformer at the oscillator output as shown in Fig. 4. As the reference input voltage is measured beyond this coil, its frequency response characteristic does not affect frequency response measurements. However, when making harmonic measurements, it should be borne in mind that such a coil may contribute undue harmonic distortion unless it is chosen to handle the desired oscillator output power adequately. The efficacy of an electrostatically shielded transformer is a function of the reduction of effective interwinding capacitance brought about by the shield which is a function of the transformer design.

In addition to the transmission circuit grounds, the shields ordinarily required on the interconnecting wiring must be grounded as shown in Fig. 4.
Bridging Gain

A special condition of gain measurement occurs when it is desired to determine bridging gain which has been defined as "the ratio, expressed in db, of the power delivered to the bridging amplifier load to the power in the load across which the input of the amplifier is bridged." For this measurement the output of the input attenuator is terminated by a resistance equal to the reference load, and the input of the equipment under test is connected in parallel with this load. Equipment intended for bridging operation will ordinarily have an input impedance high relative to that of the circuit on which it is bridged. However, it will still be necessary to take into account the reduction in power in the reference load due to the bridged input for accurate results. Otherwise, the measurement technique does not differ from that previously discussed.

Illustrative Example

A specific arrangement for measuring the gain and frequency response of a hypothetical system is shown in Fig. 5. When both meters are indicating 0.001 watt, no meter correction need be added. However, using thermocouple meters, it is probable that greater accuracy of indication can ordinarily be obtained by reading small variations on the thermocouple meter scales than by the use of 0.1-db steps on attenuators.

In this same connection if the desired input level is equivalent to -50 dbm (2.45 millivolts, rms, in series with 150 ohms), this level is achieved in the illustrative example with 0.001 watt indicated by $M_1$ and 39.5 db in the adjustable attenuator. This 39.5 db plus the 10.5 db of the matching network provides the desired equivalent of -60 dB referred to 0.001 watt.

The illustrative example of Fig. 5 is essentially similar to the general basic circuit of Fig. 2. However, it will be noted that the arrangement of the input meter circuit of Fig. 5 is such that the same input voltage appears across the input terminals of the input adjustable attenuator as that indicated by $M_1$, thus avoiding a 6-db correction required in the case in Fig. 2. With the arrangement of Fig. 5 the relative values of the series input resistances are, of course, factors in the accuracy of measurement.

Distortion Measurement

Distortion Measurements Using a Wave Analyzer

Wave analyzer determination of harmonic distortion at the output of the equipment under test involves separate measurement of the individual harmonics and the calculation of the square root of the sum of their squares. The wave analyzer is connected as shown in Fig. 2. The wave analyzer can be used to check the signal source for harmonic content by connecting it directly to the output of the oscillator circuit. Such a check should be made to insure that measurements are being made of harmonics generated in the equipment under test.

For such tests on balanced equipment the same precautions must be observed as outlined in the section on response and gain measurements. In the case of wave analyzer measurements, it probably will not be possible to connect directly to the output terminals of the equipment under test without producing unbalance as the usual wave analyzer has one terminal at ground potential, unless an isolating transformer is inserted in the input leads to the wave analyzer.

The measurement of distortion with a wave analyzer tends to be more precise than measurements with a noise and distortion meter under some conditions, as discussed later. However, it has the disadvantages that the actual observations are more time consuming, and that calculations are necessary to determine the total distortion percentages.

In making distortion measurements, the same absolute gain as well as the same relative distribution of gains and losses in the equipment under test as were used for the gain and frequency response measurements must be used to insure correlation of results.

It is also advisable to make a check at the output of the equipment under test with the signal removed to see if actual components of the signal are causing the indication; or if it is due to noise or hum generated in the equipment. If noise causes a reading approaching the same order of magnitude of the expected distortion readings, the selective method of measuring the harmonics with a wave analyzer will ordinarily result in more accurate determination of distortion than a non-selective method.
Distortion Measurements Using a Distortion Factor Meter

A distortion factor meter,\(^1\) when used for distortion measurements, attenuates the fundamental of the test signal output of the equipment under test and passes the remaining signal, which is composed of harmonics and noise. The rejection of the fundamental is achieved either by the use of a high-pass filter or by the use of a sharply selective circuit. Comparative measurements may thus be difficult to correlate, particularly if the amplitude of the remaining noise approximates that of the harmonics being measured. Such meters are sometimes so arranged that the frequency-selective circuits may be disconnected for use in measuring noise as discussed later. Such a device is ordinarily termed a distortion and noise meter.

A distortion factor meter may be substituted for the output meter as shown in Fig. 2. It will be necessary to preserve the desired grounding condition of the circuit under test and to terminate the output attenuator properly. This termination can be in part or in whole the input of the distortion and noise meter.

In addition to the errors caused by noise as already mentioned, this method of distortion measurement may be subject to other errors. The readings may be accentuated or reduced by phase relationships between different harmonics or between the same harmonics introduced by each of several elements in the equipment under test. Furthermore, when the magnitude of the harmonics exceeds a few per cent there is a metering error as the distortion factor meter indicates the ratio of the harmonics to the fundamental plus the harmonics rather than to the fundamental only.

Intermodulation Distortion

By substituting two signal sources at the input, two signals can be simultaneously applied to the equipment under test and, by means of an analyzer connected at the output, the intermodulation between the two signals can be measured. Different methods of making such measurements have been discussed in the literature.\(^1,7,8\) However, the Committee on Audio Facilities of the Transmitter Section of RMA has taken the position that “there is not sufficient data available at this time based on subjective observation to establish recommended conditions for test for intermodulation measurement.”\(^9\)

Noise

General

The measurement of electrical noise in a complex audio system such as the complete audio facilities for an elaborate broadcasting studio is very simple in theory. The reference output power at the reference frequency is first determined. The signal is then removed, the input is terminated with a resistance equal to the rated source impedance, and the residual output level is measured. The ratio of the two levels expressed in decibels is then the signal-to-noise ratio. In actuality there are two major complications in measuring the noise. These complications are the nature of the noise and the characteristics of the noise meter.

The nature of the noise is important basically because even though it is measured as electrical energy, in the actual application of the equipment under test it is ultimately converted to acoustical energy, and as such is a disturbance on a subjective basis to the listener. The noise energy present in audio facilities for radio broadcasting is ordinarily composed mainly of two types. The first is random energy distributed substantially uniformly over the frequency spectrum and of relatively uniform amplitude such as that caused by thermal noise in resistors. The second is energy concentrated at one or more frequencies. A typical cause of such noise is operation of equipment from an alternating-current power source. In addition to these two types ordinarily present, peaks of relatively high amplitude but of short duration may be experienced. Such peaks may occur at random intervals or may recur at repetitive time intervals.

The characteristics of the noise meter become a problem because of the possible variations in the nature of the noise energy. If all noises had the same wave shape, then any meter capable of indicating steady-state electrical energy and having the requisite sensitivity could be used. However, when noises have different frequency makeups, either due to various mixtures of single frequency and random noise or due to various mixtures of different single frequencies, then if the noise meter is to give equal readings on noises which are equally disturbing to a listener, the frequency response characteristic of the noise meter is important and should not be flat. Finally, the possible presence of random pulses emphasizes the time of integration of the meter and the ballistic properties of the indicating instrument.

As a temporary arrangement for measuring the noise on audio facilities for radio broadcasting (not including studio-transmitter links), the Committee on Audio Facilities has agreed upon making noise measurements with a device having a frequency response flat within plus or minus two decibels from 50 to 15,000 cps and having the ballistic characteristics of the “standard VU meter” but reading (responsive to) the rms value of a complex wave approximating a steady-state condition.\(^1\) The Committee also notes that measurement of pulse noise conditions has not been included because of lack of definition and equipment.

The results of the measurement of noise with a meter having a substantially flat frequency response are ordinarily reasonably adequate for comparative use in the field outlined herein, particularly on like equipment. For example, in audio facilities for radio broadcasting, the noise problem frequently simplifies to that of the control of hum from power supply when power is supplied as alternating current to the heaters of indirectly heated cathodes and as rectified and filtered alternating
current for the other elements. Where a diversity of conditions exists, it may be necessary to qualify the results of noise measurements to insure proper interpretation; this results largely from employing flat response measurement of electrical noise, as generally the ultimate criterion is disturbance on an acoustical basis. For this reason, frequency weighting in the meter has been generally resorted to in noise tests on telephone systems and in acoustical measurements, to simulate human ears at representative listening levels. Examples of this are the Bell System program noise measuring equipment and the American Standard Sound Level Meter.\textsuperscript{20,21,22} There are incorporated in the latter device two frequency weightings, corresponding to ear characteristics for 40 db and 70 db acoustical levels. Results of acoustical measurements with such a meter would of necessity require supplementary data for coordination with previous results of flat response measurements of the electrical noise in amplifiers intended as components of an electro-acoustical system. However, generally available data correlated with weighted electrical measurements made on amplifying systems do not appear to exist in sufficient volume to warrant abandonment at the present time of the existing use of flat weighting for electrical noise measurements on the audio systems under discussion. This decision is an expedient and is subject to review in the light of more adequate information which may become available later.

As noted above, the Committee on Audio Facilities has specified that the noise measuring device have the ballistic characteristics of a standard VU meter but that it read the rms value of a steady complex wave. The standard VU meter does not read this rms value as the exponent applying to the instantaneous value of voltage for it\textsuperscript{2} is 1.2 ± 0.2, while the corresponding exponent is 2.0 for an rms device. However, it seems likely that on ordinary steady noises found in audio systems the difference between a VU meter reading and an rms reading would not exceed about 1 db. For noises of markedly peaked wave shapes, the difference would be greater\textsuperscript{22}; but these are not ordinarily originated in audio facilities.

The Committee on Audio Facilities has not specified the frequency response of the noise measuring equipment above 15,000 cycles. It would seem desirable that this response cut off frequencies above 15,000 cycles, since the broadcast receiver and listener would ordinarily be insensitive to them. If measuring instruments of differing frequency response above 15,000 cycles are used to measure a noise having components above 15,000 cycles, different readings will be obtained, particularly if the equipment under test has a substantial frequency response above 15,000 cps. However, the difference between the readings of a standard volume indicator and an ordinary noise and distortion meter, contributed by the differences between their frequency response characteristics, will usually amount to only a few tenths of a decibel, though in particular cases (e.g., where the apparatus under test picks up the field from a long-wave radio-telegraph station) larger differences may be found.

**Noise Measurement Using a Distortion and Noise Meter**

To make noise measurements with the distortion and noise meter of the general type already described, its calibration is first adjusted so that it indicates zero with a test signal at a medium frequency such as 1,000 cps after the input level is established at the normal or operating level. The gain and control settings of the equipment under test should again be the same as for previous measurements. The output of the equipment under test should be terminated by resistance equal to the rated load impedance. Noise measurement is then made by removing the test circuit from the input to the equipment under test and substituting a resistor equal to the rated input source impedance. The resistor should be noninductive and of a type, such as a wire-wound resistor, which will not in itself generate appreciable noise. In high-impedance, low-level circuits it may be necessary to shield it carefully to avoid hum pickup trouble.

**Acknowledgements**

The authors wish to acknowledge the assistance of W. F. Byers of the General Radio Company for his help in the preparation of a preliminary draft. In addition, thanks are due to those who offered constructive comments to this presentation in manuscript form and to the many persons who, over the years, have published results of their work and thus made this compilation possible. The appended Bibliography is not intended to be all-inclusive, but it does indicate some of the sources of information available which bear on the subject matter.

**Bibliography**

Error-Actuated Power Filters

GORDON NEWSTEAD† AND D. L. H. GIBBINGS†

Summary—A filter is described which is capable of handling large amounts of power using components rated at a small fraction of that power. The filter employs negative feedback to cancel out unwanted frequencies. Design equations are given and the operation of a typical filter circuit is examined.

I. INTRODUCTION

A Passive filter achieves its results by presenting a frequency variable impedance to the supply which is to be filtered. An alternative method of filtering is to generate voltages to oppose and cancel those which it is desired to remove. In this paper we describe filters operating on this principle which may be likened to error-actuated servomechanisms, the error being the residual voltage at the unwanted frequencies. We have named such filters “error-actuated filters.”

Error-actuated filters have distinct advantages over classical types of filter in respect to power-handling capacity, and their filtering action is not significantly affected by large variations in load impedance. In one stage of such a filter, the error can be reduced to a value fixed only by stability requirements, and, if necessary, a number of stages can be used in cascade. The lower limit to which harmonics can be reduced in this way is fixed by the production of harmonics in the filter itself.

In this paper we are mainly concerned with the purification of a fundamental frequency, by the removal of unwanted harmonic frequencies. However, there appears to be no fundamental limitations on the method which would prevent its being extended to apply to a given set or band of frequencies, and we intend, subsequently, to investigate error-actuated filters with wider pass bands.

Although the power to be filtered does not pass through the amplifiers used in the filter, it is necessary, as discussed below, for it to have a power rating. However, in the case where the unwanted frequencies are a small fraction of the wanted frequencies, this rating is only a fraction of the power handled by the filter. For example, with receiving tubes and small transmitting tubes, we have found it possible to remove the harmonics from about \( \frac{1}{2} \) kw of alternating current containing 5 per cent harmonics. Larger powers could be handled using larger transmitting tubes.

II. NOTATION AND SYMBOLS

The subscripts \( i \) and \( o \) are used to denote input and output quantities, and the subscripts \( w \) and \( u \) are used to denote wanted and unwanted quantities, respectively. The symbols \( I \) and \( V \) are used for current and voltage, respectively; for example, \( V_{uo} \) means the wanted component of the voltage present in the input. The subscript \( H \) is used to denote the difference between input and output quantities, thus

\[ V_H = V_i - V_o. \]

In general, the \( V \) and \( I \) are complex numbers to take account of phase shifts, the modulus symbol being used for the rms value of their magnitudes. In the case where there is more than one wanted or unwanted component, symbols such as \( V_{iu} \) are to be interpreted as applying to typical components, but \( |V_{iu}| \) is to be interpreted as the rms value of all the \( u \) components. We also use:

- \( D \) = the maximum allowable dissipation in the anode of the final amplifier tube. With the suffix \( u \), \( D \) means the actual power supplied by or to the tube at the unwanted frequencies.
- \( T \) = the turns ratio of the output transformer of the amplifier (see Fig. 2).
- \( G \) = the gain of the amplifier.
- \( \beta \) = a factor to take account of the attenuation of the network.
- \( I_s \) = the steady dc current in the anode of the final amplifier tube.
- \( E_s \) = the steady dc voltage on the anode of the final amplifier tube.
- \( r_o \) = the plate resistance of the final amplifier tube.
- \( n \) = \( |V_{iu}|/|V_{iu}| = \) ratio of unwanted to wanted voltages present in the input.

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III. Theory of the Error-Actuated Filter

The general arrangement of this type of filter is shown in Fig. 1. The unwanted frequencies are taken through a feedback path of gain $G\beta$ and reinserted further back in the circuit, the connection being such that the feedback is negative. The voltage gain $G\beta$ is defined as the ratio $V_H/V_o$ and is a function of frequency. It is convenient to take the gain as the product of two factors, $G$ depending on the amplification of the amplifier, and $\beta$ depending on the passive network. This method and notation will be familiar to those accustomed to servomechanisms and inverse feedback amplifiers, although so far as the authors are aware this method of filtering has not been described before. Terman¹ and others have described amplifiers whose gain versus frequency characteristics are controlled in a manner similar to that of Fig. 1. However, if such devices were to be used as filters, the whole of the power would have to pass through, and be delivered by, the amplifier.

In an arrangement such as in Fig. 1, it is easy to show that

$$V_o = V_i/(1 + G\beta). \quad (1)$$

By using a null network in the feedback path, $\beta$ can be made zero at the wanted frequencies, and we have

$$V_{ou} = V_{iu}. \quad (2)$$

The wanted frequencies are thus transmitted without attenuation. At the unwanted frequencies, however, $\beta$ is different from zero and we have

$$V_{eu} = V_{iu}/(1 + G\beta), \quad (3)$$

where the factor $1 + G\beta$ can be made as large as stability considerations will allow. This aspect will be discussed later.


An important matter is the power-handling capacitance required of the amplifier. The volt-amperes to be supplied by the amplifier at the unwanted frequencies are equal to

$$|V_{iu}| \cdot |V_{ou}| / |Z_L|,$$

which, on substituting for $|V_{iu}|$ and $|V_{ou}|$ and taking $|1 + G\beta| \approx |G\beta|$, reduce to

$$|V_{iu}|^2 / |Z_L| \cdot |G\beta|.$$

The volt-amperes being supplied to the load at the wanted frequency are

$$|V_{iw}|^2 / |Z_L|.$$

Thus the volt-amperes rating of the filter is

$$D_u = |G\beta| / \mu^2. \quad (4)$$

Taking $|G\beta| = 50$ (a value typical of those we have achieved) and $\mu = 1/20$, e.g., 5 per cent unwanted voltages in the supply, we see that the amplifier is required to deliver only 1/20,000 of the volt-amperes supplied to the load. In practice the power-handling capacity of the amplifier at the unwanted frequencies is not a limiting factor on the power-handling capacity of the filter where the percentage of unwanted frequencies is not too large.

The load current $I_L$ flows through the output terminals of the amplifier and because of the impedance presented by these terminals, a voltage drop at the wanted frequencies occurs. Besides impeding the regulation of the supply, this voltage drop is impressed on the amplifier which must be capable of handling the requisite volt-amperes. The magnitude of the voltage drop is $|I_L| \cdot |Z_w|$, where $Z_w$ is the impedance looking into the amplifier terminals from the line side at the wanted frequency, and $|I_L|^2 \cdot |Z_w|$ volt-amperes are handled by the amplifier. If voltages at the wanted frequency are present or introduced further back in the amplifier, then $Z_w$ is not entirely a passive impedance and it can be controlled by controlling the magnitude and phase of these voltages. It has been found, for example, that by unbalancing the null network, the voltage drop at the wanted frequency can be reduced to zero. However, the amplifier is now supplying volt-amperes at the wanted frequency, and each case can only be examined in detail when its arrangement is completely known.

A typical arrangement is shown in Fig. 2. The final tube of the amplifier is shown there as a single-ended stage, whereas, in order to reduce the effect of the direct current through the transformer, push-pull arrangements are preferred in practice. However, the change in the analysis required is trivial, and Fig. 2 is discussed as having no irrelevant aspects which may tend to obscure the fundamental facts. In passing, it is important to note that any harmonics introduced by the amplifier or other components in the circuit are themselves divided by $|1 + G\beta|$. 

---

**Fig. 1—General arrangement of error-actuated filter.**
In order to select a tube which will be capable of filtering a given power and to fix its operating conditions at their optimum values, we require an analysis that will give the power-handling capacity of the filter in terms of the power-handling capacity of the tube. We will consider a tube restricted to operate class A on the linear portion of its characteristic and without voltages at the wanted frequencies applied to its grid. This last restriction can be removed, if desired, but as we have found that the voltage which can be applied to the grid is a limiting factor, it is not generally desirable to increase it by applying voltages at the wanted frequencies. There are four conditions to be fulfilled:

(i) The grid drive must be limited so that grid current does not flow during the positive half of the cycle, and plate current is not cut off during the negative half of the cycle. So far as the harmonics are concerned, the tube is working in a very large impedance, so that the load line is nearly horizontal and there is little possibility of driving the grid-to-plate current to cutoff. If \(E_r\) is the grid bias voltage, the grid drive must not exceed \(E_r/\sqrt{2}\), and, since the gain of the stage from tube grid to transformer secondary is \(\mu/T\), we have

\[
T \cdot \frac{|V_{iw}|}{\mu} \leq E_r/\sqrt{2}
\]

or

\[
T \leq \frac{\mu E_r}{\sqrt{2} \cdot |V_{iw}|}.
\]  

(5)

(ii) The plate voltage must not exceed the rated value on the positive half of the cycle or become negative on the negative half cycle. If the dc voltage \(E_o\) is adjusted halfway between these limits, we have

\[
|V_{iw}| \cdot T + r_o \cdot |I_w| / T \leq E_o/\sqrt{2}.
\]

(6)

(iii) Similar considerations as under (ii) above apply to the plate current \(I_o\), and we have

\[
|I_w| / T \leq E_o / \sqrt{2}.
\]

or

\[
|I_w| \leq T \cdot I_o / \sqrt{2}.
\]  

(7)

(iv) The dissipation in the tube at the wanted frequency must not be exceeded, and we have

\[
|I_w|^2 \cdot r_o \leq T^2(D - D_o).
\]  

(8)

With most tubes it is generally found that the restrictions due to (i) and (iii) are first reached and so the transformer ratio is chosen by making (5) an equation and limiting the load current in accordance with inequality (7). A check is then carried out to see that (6) and (8) are satisfied. In order to prevent distortion and to limit the dissipation it is necessary that these last inequalities are rather strongly fulfilled.

The following considerations show how to fix the operating conditions. Using (5) and (7) as equations, we obtain

\[
|I_w| = T \cdot I_o / \sqrt{2} \cdot |V_{iw}|.
\]

Multiplying through by \(|V_{iw}|\) we have for the volt-amperes rating of the filter

\[
|I_w| \cdot |V_{iw}| = \mu E_o I_o / 2n.
\]

We therefore desire to maximize \(\mu E_o I_o\). Assuming linear characteristics, we have

\[
E_o = \frac{1}{\mu} (E_o - I_o r_o)
\]

\[
\Rightarrow |I_w| \cdot |V_{iw}| = \frac{1}{2n} I_o (E_o - I_o r_o)
\]

\[
= \frac{1}{2n} (D_o - I_o^2 r_o).
\]  

(9)

Thus the greatest power can be handled when \(I_o\) is made as small as possible. In the limit, the volt-amperes that can be handled are given by

\[
|I_w| \cdot |V_{iw}| = D_o / 2n.
\]  

(10)

A suitable design procedure is, then, as follows:

(i) Select a tube or decide on the number of tubes in parallel required, using (10). It has been our experience that a factor of safety of about 2 should be allowed at this stage.

(ii) Select \(I_o\) as small as possible on the reasonably linear portion of the characteristics.

(iii) Work out \(E_o\) from \(E_o I_o = D_o\).

(iv) Find \(E_r\) from \(E_o\) and \(I_o\).

(v) Find \(T\) from equation (5).

(vi) Work out \(I_w\) from equation (7).

(vii) Check inequality (6). If it does not hold it can generally be fulfilled by reducing \(T\) and repeating (vi) and (vii).

It should be stressed that no great accuracy is claimed for the above analysis, but it does provide order of magnitude calculations useful for design purposes.

Investigations using the above methods show that in the receiving tube class the most suitable output tubes
are high-powered triodes such as the 2A3. The 807, as a triode with a small resistor in its screen to limit screen dissipation, has also proved very suitable. With two of these tubes in push-pull circuits, it is possible to filter about $\frac{1}{2}$ kw of alternating current containing 5 per cent harmonics, actual circuits and performances being given in section V. Greater power-handling capacity may be obtained by going to transmitting tubes or by using tubes in parallel; $m$ tubes in parallel increasing the capacity $m$ times. In the transmitting tube class it is found that two 833A's in push-pull would be able to filter a power of about $3\frac{1}{2}$ kw.

The alternating-current supply voltage at which filtering is carried out does not affect the power-handling capacitance of the filter, but as $|V_{in}|$ increases with an increase of $|V_{o}|$, a smaller value of $T$ will be necessary (see (5)). Provided that the filter remains stable, the filtering action will be improved by the increase in gain brought about by the reduced value of $T$. If the value of $T$ found is such that it is not possible to obtain sufficient gain from a conventional voltage amplifier (because of the loss in the output tube), and as it is not desirable to increase the number of stages of voltage amplification for stability reasons (see section IV), a circuit of the type described by Jeffrey can be employed. This circuit uses the high input resistance of the cathode follower as the load resistance of the voltage amplifier. A dc coupling is used between the output of the cathode follower and the grid of the power tubes, to avoid the introduction of further phase shifts.

IV. Stability

In the theory given above there is no reason why $G$ should not be indefinitely increased, but in practice $G\beta$ is function of frequency and at some frequency the phase shift may be sufficient to cause the feedback to be positive. The requirements on $G\beta$ as a function of frequency for the system to be stable were discovered by Nyquist and have been discussed in some detail by Bode. In devices of this type it is not desirable to use conditionally stable arrangements, so the requirements reduce to ensuring that whenever the phase shifts become equal to odd multiples of $\pi$, $|G\beta| < 1$. We have confined ourselves to the cases in which phase shift is always less than $\pm \pi$ in the working range and arranged that the gain of the amplifier has dropped below the critical value before appreciable phase shift occurs. This limits us to the case of no more than two stages of resistance-coupled amplification, and, as discussed above, the last tube is essentially a power amplifier with no voltage gain: it often has a gain of about one-half. Under these conditions values of the order of 50 can easily be obtained. This means that unwanted voltages can be reduced in the ratio 1:50 (34 db) in a single stage.

Stability requirements also limit the types of passive network which can be inserted in the feedback path. Ladder type networks, because of their large phase shifts, are not suitable, in spite of their desirable attenuation characteristics. Parallel-T null networks are very suitable as their phase shift does not exceed $\pm \pi/2$, and the greater part of it occurs within a relatively limited region near the null frequency. It is thus possible to design the amplifier to have relatively little phase shift where the parallel-T has its greatest shift and vice versa, and so ensure that the phase shift of the network and amplifier do not greatly reinforce each other.

Bridge-T networks are capable of greater selectivity than the parallel-T, and in applications where frequencies very close to the desired one are to be removed, their use is indicated. However, they were abandoned at 50 ~ because nonlinearity in the iron-cored inductances caused the balance point to vary with load.

V. Some Experimental Results

Fig. 3 shows a circuit using 807 tubes which has been used for filtering a 250-volt, 50 ~ supply with 5 per cent harmonics.

The tube conditions are as follows:

$\mu = 8$
$\mu_0 = 1600 \Omega$
$E_v = 450 \text{ volts}$
$E_c = 40 \text{ volts}$
$I_v = 40 \text{ ma}$
$D_c = 18 \text{ watts}$

Equations (5) and (7) give $T = 18$ and $I_v$ (for 2 tubes) as 1.02 amperes, and a check shows that inequality (6) is fulfilled.

The measured performance of this circuit with $T = 15$ is given in Table I (b). It will be seen that $I_v$ is 1.4 amperes so that the tubes must be operating with their anode current cut off over a small part of the cycle. This would explain the appearance of even harmonics in the output, but fortunately, as noted earlier, the harmonics produced by the device are also divided by the factor $1 + G\beta$. 

---

Table 1 (a) gives the measured performance with $T = 8$, and as would be expected shows better filtering, since the gain of the amplifier is greater.

As the rated maximum dissipation of an 807 is 25 watts, equation (10) would lead us to expect a volt-ampere capacity for the filter of 500. The actual value achieved was 240. The discrepancy may be accounted for by 3 factors:

1. The tubes were run at 18 watts dissipation.
2. $I^2r_a = 2.5$ and is not negligible.
3. It was assumed that two tubes in push-pull would give twice the current-handling capacity, which is not true with the phase splitting device used.

If we take account of the first two factors, (9) gives us a theoretical volt-ampere capacity of 310, which is thus not greatly in excess of the achieved value.

The relative amounts of the individual harmonics can be adjusted by slightly unbalancing the parallel-$T$ network. We have found that the best procedure is to balance the parallel-$T$, as indicated by the vanishing of the fundamental frequency on an oscillograph connected across the grid of the 807, and then slightly unbalance it to increase the fundamental component or reduce any of the harmonics to a minimum. Failure of the filter shows itself in the presence of discontinuities of the wave form on the grid. The results given above represent the sort of typical compromise that can be obtained in practice.

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### VII. Conclusion

The method described above enables filters to be constructed which are capable of handling powers up to several kilowatts without requiring a large number of power-rated components. It appears that these devices are the most satisfactory means yet described for the production of a very pure wave form from an ac source and should have a number of applications, particularly in the field of electrical measurements. A further application that appears promising is the inclusion of a filter of this type between an ac voltage regulator that makes use of saturable reactors and a stabilized dc power supply. This should result in an improvement in the stability of the system, because if a filter is not used, changes in harmonic content, as the ac regulator operates, result in changes in average value of the ac voltage, even though its rms value remains constant. The over-all stability of the system, without the filter, is thus not as great as might be expected at first.

The actual filters described in this paper are of a rather specialized type, but it is felt that the principle of error-actuated filtering is capable of very wide extensions.

![Fig. 4—General arrangement of compensating filter.](image-url)
Signal-to-Noise Ratios of Linear Detectors

R. H. Delano*, Associate IRF

Summary—A method is presented for obtaining the output signal and noise from a linear detector, including the output signal wave form and both the signal and the noise spectra. The method is applicable to any type of amplitude-modulated input signal at any signal-to-noise ratio, provided only that the bandwidth is small compared with the center-band frequency, and requires that certain steps be performed graphically rather than analytically. A comparison is made with square-law detection for a few useful cases. It is demonstrated that the linear detector gives a higher output signal-to-noise ratio than the square-law detector for some types of signal.

I. Introduction

The method presented here is a restatement and extension of principles discussed in the literature to the problem of the signal and noise output of a linear detector when the input signal is a general amplitude-modulated wave. As a rule, only amplitude-modulated waves, including any type of pulse-modulated signals, are detected by a single diode. The noise and signal output of a discriminator using linear detectors can also be derived by this same general method.

The general problem is to obtain the signal and noise spectrum of the output of a linear detector, given the input signal and the input noise power spectrum. From a mathematical standpoint this general problem is not solved until an analytic solution is obtained for the output signal and noise for any input signal (and noise) whatever. The method presented here is not such a solution. First of all, some graphical Fourier analyses have to be performed since the corresponding analytic expressions are much too cumbersome to handle. Second, the ratio of input bandwidth to the center frequency of the band is assumed to be small, and finally the input signal is subjected to more restrictive conditions than merely the one of having its components in the appropriate frequency band. If it is an amplitude-modulated wave, this fact is recognized, and use is made of the necessary relation between the phases of sidebands symmetrical around the carrier frequency. The type of input noise considered is normally distributed random noise. Several analytic solutions, both approximate and exact, have been given in the literature for certain less general problems connected with linear detection of signal and noise.

The basic philosophy involved in the method is simple, and is commonly referred to as the adiabatic assumption. For the AM signal, for example, it is merely recognized that the signal may be factored into two parts, a slowly varying envelope and a carrier-frequency sine wave. The output signal for an unmodulated carrier is merely the dc output. Similarly, the output signal with modulation is this same dc, now varying slowly with time as a function of the amplitude of the input sine wave, and hence, of the envelope. Thus as a prerequisite, the solution for the dc output with noise and sine-wave signal input is required. Similarly, the output noise spectrum with AM signal can be modified from this simple case of a sine-wave signal at the input.

II. The Output Signal with AM Signal Input

Consider first the dc output, which is the mean or average value of the output, when a sine-wave and noise are applied to the input of a nonlinear device. In general, this solution is given by

\[ I = F(V_s, V_n^2) \]  

where

- \( I \) = mean or dc output current (output voltage for and envelope detector)
- \( V_s \) = peak amplitude of the input sine-wave (peak carrier amplitude)
- \( V_n \) = input noise voltage
- \( V_n^2 \) = mean square input noise voltage

For normally distributed random noise, whose probability distribution is not a function of its spectrum, it is easily demonstrated that the dc output is not a function of the spectral distribution of the input noise, but depends only on its mean square amplitude. Given the general detector characteristic

\[ I = F_d(V_n) \]  

\( V_n \) = input voltage to the nonlinear device

\( I \) = output current (or voltage) from the nonlinear device.

The probability density of the output, prior to filtering, is

\[ P(I)dI = P(V)dV \]  

where

- \( P(I) \) = probability density of \( I \), the output current.
- \( P(V) \) = probability density of the input voltage.

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† Hughes Aircraft Company, Culver City, Calif.


and the mean value of the output is then

\[ I = \int_{-\infty}^{\infty} IP_d(I) dI = \int_{-\infty}^{\infty} P_i(V) F_d(V) dV. \]  

(4)

There is clearly no dependence on the input spectrum in this relation; in other words, the mean or dc output is a function only of the total input noise power. Furthermore, filtering of ac components of the output can have no effect on the dc value of the output, so this relation is always applicable.

For normally distributed random noise, the function \( F(V_i, V_2) \) is uniquely determined by the detector characteristic \( F_d(V) \). For the ideal linear detector whose transfer characteristic is

\[ F_d(V) = \alpha V \quad \text{for} \quad V > 0 \]

\[ = 0 \quad \text{for} \quad V < 0. \]

The corresponding dc output has already been derived by Bennett

\[ I = \alpha \sqrt{\frac{W_n}{2\pi}} t F_1\left(-\frac{1}{2}; 1; -\frac{W_1}{W_n}\right). \]  

(5)

Where

\[ W_n = \frac{V_n^2}{2} \]

\[ W_1 = \frac{1}{2} V_2^2 \]

\( W_n \) = input power envelope

\( t F_1(a; c; z) \) = confluent hypergeometric function

\[ = 1 + \frac{a}{1!} + \frac{a(a+1)}{c(c+1)} \frac{z^2}{2!} + \cdots. \]

Furthermore, Bennett has shown that the output of an ideal envelope detector differs only by the constant factor \( \pi \) from the low-frequency output of an ideal linear detector. Since constant factors do not affect signal-to-noise ratios, this solution for the ideal linear detector is applicable to either case.

Since the frequency components of the envelope differ by at least one order of magnitude from the carrier frequency, the time variation of the actual envelope \( V_i(t) \) may be introduced into this expression. The result is the output signal as a function of time, although, to be more exact, the contribution to the output due to noise must first be subtracted out, giving

\[ I_s(t) = \text{output signal as a function of time} \]

\[ y = \frac{V_s}{W_n^{1/2}} \quad \text{signal voltage envelope (peak)} \]

\[ \frac{W_n}{W_n^{1/2}} \quad \text{rms input noise voltage} \]

The frequency components of the signal must be obtained by a graphical Fourier analysis of \( I_s(t) \). From a practical standpoint, such a procedure is much more effective than attempting to derive analytic expressions for the Fourier components of \( I_s \) for any arbitrary \( V_i(t) \). This process may be carried out with the help of Fig. 1 where \( I_s \) is given as a function of \( y \). To use a

\[ I_s(t) = \sum_{n=1}^{\infty} C_n \cos (\omega_n t + \phi_n) \]  

(7)

\( \omega_n = 2\pi(n/T) = 2\pi f_n \)

\( \phi_n \) = random phase angle (uniform probability density from 0 to \( 2\pi \)).

The amplitude of the noise components \( C_n \) is related to the noise power spectrum by the relation.
\[ \frac{1}{2} C_s^2 = w(f_s)\Delta f = w(f_s) \frac{1}{T} \]

\[ w(f) = \text{noise power spectrum-mean square volts per cycle}. \]

The input noise spectrum is determined by the bandpass characteristics of the narrow-band elements of the input circuits, since prior to narrow banding, the spectrum is uniform over all frequencies. For a general nonlinear device, it has been demonstrated that the output noise can be represented as the sum of intermodulation products of all orders. Since only the spectrum and statistical properties of the input noise are known, a direct Fourier integral of the output cannot be formulated, and this representation is the most convenient one. For a square-law detector the only nonzero intermodulation products are the second-order ones, signal-noise and noise-noise intermodulation. These products give difference frequency beats in the audio or video region. Since the phases of these difference frequency beats are also random for the case of noise, the mean square amplitudes or the "power" spectra of the two kinds of intermodulation add linearly in the output. It turns out conveniently that these second-order intermodulation products are the only significant contributors to the output of an ideal linear detector also.

For a sine-wave signal and noise, the difference frequency beat between signal and noise gives an output component whose amplitude is proportional to the product of the sine-wave amplitude and the noise-component amplitude. The proportionality constant is a function of the input signal-to-noise ratio (a constant for constant \( W_s \)). Thus, the shape of the signal-to-noise output spectrum is the same as that of the input noise, only shifted towards dc by the carrier or signal frequency. Of course, this shift leaves the lower half of the spectrum at negative frequencies. In the actual output it is necessary to reflect these portions back onto positive frequencies, and add the power spectra. Suppose, for the moment, that the spectrum is left as it is, with both positive and negative frequency components, and modified to take into account variations in signal amplitude with time.

As the input signal envelope varies, the amplitude of each of the beat-frequency output components varies in a related manner determined by the functional dependence of the proportionality constant on instantaneous signal-to-noise ratio. If the amplitude of a sine-wave, whatever its frequency or phase, varies periodically, the wave itself may be expanded into a carrier at the sine-wave frequency plus symmetrical modulation sidebands, even if some of the sidebands fall at negative frequencies and later have to be reflected back onto the positive half of the scale. Thus, the above spectrum becomes a function of time and the actual steady-state frequency spectrum is obtained by summing a series of power spectra corresponding to the carrier and sideband amplitudes in the above Fourier analysis of a single component. Mathematically this procedure is developed as follows:

The amplitude of the signal-to-noise beat at the difference frequency \( f_s - f_i \) is given by Bennett

\[ A_{ss} = \frac{\alpha C_s}{2} \left[ \frac{W_s}{\pi W_n} d F_1 \left( \frac{1}{2} ; \frac{W_s}{W_n} \right) \right] \]

\[ = \frac{\alpha C_s}{\sqrt{8\pi}} d F_1 (\frac{1}{2} ; 2 ; -\frac{1}{2} y^2). \] (8)

Plotting \( A_{ss} \) over a cycle of the input modulation and performing a graphical Fourier analysis gives the expansion

\[ A_{ss}(t) = \frac{\alpha C_s}{\sqrt{8\pi}} \left[ a_0 + \sum_{n=1}^{a} a_n \cos (n \omega_s t + \theta_n) \right] \] (9)

\[ 2\pi f_m = \omega_m = 2\pi \text{ times the fundamental envelope modulation frequency}. \]

\[ A_{ss} \text{ is plotted versus } y \text{ in Fig. 2.} \]

Fig. 2—Hypergeometric function used to obtain \( A_{ss} \).

The \( a_0 \) coefficient gives the carrier amplitude of each beat component and the modulation sidebands are of amplitude \( a_n/2 \). The \( a_0 \) term gives an output spectrum equal to \( \alpha^2 a_0^2/8\pi \) times the shifted input spectrum and the \( a_1 \) terms give similar spectra shifted in frequency by \( n f_m \) relative to the \( a_0 \) spectrum. Thus, the output noise power spectrum due to signal-to-noise intermodulation is

\[ w_{ss}(f) = \frac{\alpha^2}{8\pi} \left[ a_0^2 w_s(f) + \sum_{n=1}^{a} \frac{a_n^2}{4} (w_s(f - n f_m) + w_s(f + n f_m)) \right] \] (10)

where

\[ w_s(f) = w(f_i + f) \text{ with } f \text{ both positive and negative} \]

\( f_s = \text{input carrier frequency}. \)

If the input noise power spectrum \( w(f) \) is symmetrical around the signal carrier frequency, this expression simplifies to

\[ w_{ss}(f) = \frac{\alpha^2}{8\pi} \left[ \frac{\alpha^2}{2} w_s(f) + \sum_{n=1}^{a} \frac{a_n^2}{2} w_s(f - n f_m) \right] \] (11)

At the conclusion of this summation process, all portions of the noise power spectrum at negative frequen-
cies may be reflected onto the positive side, and added to any portions already there.

The procedure for noise-to-noise intermodulation is identical, except that the output spectrum for an unvarying sine-wave signal is different. This spectrum is proportional to

\[ w(f) = \frac{1}{W_n} \int_{0}^{\infty} w(\lambda) w(\lambda + f) d\lambda \]  \hspace{1cm} (12)

with \( f \) both positive and negative, and it is always symmetrical around zero frequency or dc. The expression for the amplitude of a single beat frequency component at the frequency \( f = f \) is

\[ A_{nn} = \frac{\alpha \nu e}{\sqrt{8\pi W_n}} \left[ b_0 + \sum_{n=1}^{\infty} b_n \cos (n\omega_n t + \theta_n) \right]. \]  \hspace{1cm} (14)

Here, \( A_{nn} \) is plotted versus \( y \) in Fig. 3.

A graphical Fourier analysis of \( A_{nn} \) plotted versus time over a period of the input modulation gives the expansion

\[ A_{nn} = \frac{\alpha \nu e}{\sqrt{8\pi W_n}} \left[ b_0 + \sum_{n=1}^{\infty} b_n \cos (n\omega_n t + \theta_n) \right]. \]  \hspace{1cm} (15)

The corresponding noise spectrum, allowing both positive and negative frequencies, is

\[ w_n(f) = \frac{\alpha^2}{8\pi} \left[ b_0^2 w_i(f) + \sum_{n=1}^{\infty} b_n^2 \frac{w_i(f + n\omega_n)}{2} \right]. \]  \hspace{1cm} (16)

After summation, any portions of this spectrum at negative frequencies may be reflected onto the positive scale and added to any already at positive frequencies. Similarly, the total output noise spectrum is the sum of the spectra due to noise-to-noise and signal-to-noise intermodulation.

\[ w_n(f) = w_n(f) + w_{nn}(f). \]  \hspace{1cm} (17)

IV. THE TOTAL OUTPUT SIGNAL AND NOISE

The preceding calculations, including the graphical Fourier analyses, are only necessary if the spectral distribution of the signal and noise in the output is desired. The total output signal and noise power for a sine-wave signal input are functions of the ratio of total input signal power to total input noise power only, provided none of the output components are removed by filtering. The total mean signal and noise power may be obtained by taking the average of these expressions over a period of the modulation for AM signal inputs. For signal, this mean power output is

\[ P_s = \frac{\alpha^2 W_n}{2\pi} \left[ F(\frac{1}{2}; 1; 1 - \frac{1}{2} y^2) \right]. \]  \hspace{1cm} (18)

The total output noise power due to signal-to-noise intermodulation is

\[ W_{nn} = \frac{\alpha^2 W_n}{8\pi} \left[ \frac{\gamma^2}{r} F(\frac{1}{2}; 1; 1 - \frac{1}{2} y^2) \right]. \]  \hspace{1cm} (19)

and due to noise-to-noise intermodulation is

\[ W_{nn} = \frac{\alpha^2 W_n}{8\pi} \left[ \frac{e^{-w_{1/4}W_n}}{r} F(\frac{1}{2}; 1; 1 - \frac{1}{2} y^2) \right]. \]  \hspace{1cm} (20)

In general, the mean power for the signal is not too meaningful, since information is derived from the ac components of the signal. For the noise, however, the shape of the spectrum can be roughly estimated, and the total noise power is a useful quantity.

V. SQUARE-LAW DETECTION

The same techniques are, of course, applicable to square-law detection except that the analytic solution is already known. Given the square-law characteristic,

\[ I = \alpha_i V_n \]  \hspace{1cm} (21)

\[ \alpha_i = \text{detector coefficient}. \]

The output signal and noise can be expressed directly in terms of the input signal and noise. Suppose the input signal is given by

\[ E_i = V_e \left[ c_0 + \sum_{n=1}^{\infty} c_n \cos (n\omega_n t + \theta_n) \right] \cos \omega t, \]  \hspace{1cm} (22)

\[ \text{See Section 4.5 of footnote reference 3.} \]
and the input noise power spectrum by \( w(f) \) again, then the low-frequency output signal is simply,

\[
V(t) = \alpha V_e \left[ c_0 + \sum_{n=1}^\infty c_n \cos (n \omega_{at} + \theta_n) \right]^2 \cos^2 \omega t
\]

\[
= \frac{1}{2} \alpha V_e^2 \left[ c_0 + \sum_{n=1}^\infty c_n \cos (n \omega_{at} + \theta_n) \right]^2.
\]  \hspace{1cm} (22)

The signal-to-noise intermodulation output spectrum is

\[
w_n(f) = \alpha^2 V_e^2 \left[ c_0^2 w_0(f) + \frac{1}{2} \sum_{n=1}^\infty c_n^2 w_0(f + n f_a) \right].
\]  \hspace{1cm} (24)

Reflection onto the positive frequency scale is again applicable after this summation. The noise-to-noise power spectrum is also

\[
w_n(f) = \alpha^2 \int_{-\infty}^{\infty} w(\lambda) w(\lambda + f) d\lambda.
\]  \hspace{1cm} (25)

where \( f \) is both positive and negative. If \( w(f) \) is symmetrical around the signal carrier frequency, this expression is simply

\[
w_n(f) = \alpha^2 V_e^2 \left[ c_0^2 w_0(f) + \frac{1}{2} \sum_{n=1}^\infty c_n^2 w_0(f + n f_a) \right].
\]  \hspace{1cm} (26)

Reflection onto the positive frequency scale is again applicable after this summation. The noise-to-noise power spectrum is also

\[
w_n(f) = \alpha^2 \int_{-\infty}^{\infty} w(\lambda) w(\lambda + f) d\lambda.
\]  \hspace{1cm} (27)

where \( W_n = \) average input signal power.

These expressions will be used to obtain a simplified comparison of signal-to-noise ratios of gated, pulse-modulated signals for square-law and linear detectors in the next section. Let us, however, state the basic conclusions of the preceding theoretical development at this point.

Conclusions:

1. A general method has been presented for determining the output signal and noise of any nonlinear device, and in particular of an ideal linear detector, given the solution when the input signal is a constant sine wave.

2. Either the output spectra or the total output noise power, or both, may be derived.

3. The restrictions on the input signal and noise are as follows:
   (a) The signal is an AM wave.
   (b) The ratio of input bandwidth to center frequency is small.
   (c) The noise is random.

VI. A SIMPLIFIED COMPARISON OF SQUARE-LAW AND LINEAR DETECTORS FOR SIGNAL-TO-NOISE RATIO

The preceding method for obtaining the total output noise power for a linear or envelope detector and for a square-law detector offers a simple basis for comparison of these two types of detectors. The shape of the output noise spectra for the two cases is different, but usually not sufficiently different to produce any very large error if the shapes are assumed to be the same. In this case the total noise power output is the only possible variable. For small amounts of sinusoidal modulation either of a continuous carrier or of rectangular pulse signals, the ratio of total signal power to total noise power in the output for linear and square-law detectors can be calculated. The ratio of the output signal-to-noise ratios so obtained is a good basis for comparing the relative merits of the two detectors for a variety of applications. This calculation is made as follows:

A. Square-Law Detector

Let us consider either an uninterrupted carrier or rectangular pulses of the carrier frequency. The input signal is then

\[
E_i = V_e \sin \omega_{at} \text{ during the pulses}
\]

\[
= 0 \text{ otherwise.}
\]  \hspace{1cm} (28)

Let the pulses have a duty cycle \( k = f_r T \) where

\[
f_r = \text{pulse repetition frequency}
\]

\[
T = \text{pulselength}
\]

\[
k = 1 \text{ for the uninterrupted carrier,}
\]

and the pulses be surrounded by a gate \( t_g \) in length

\[
k_g = \frac{f_r T}{f_r T} \geq k
\]

\[
k_g = 1 \text{ for the uninterrupted carrier.}
\]

The output signal is given by

\[
V_o = \alpha E_i^2 = \frac{1}{2} \alpha V_e^2 \text{ during the pulses}
\]

\[
= 0 \text{ otherwise.}
\]  \hspace{1cm} (29)
Referring to (26) and (27), the total output noise power is seen to be the sum of the following two quantities

\[ W_{sn} = \alpha^2 V_s^2 W_n k \]

\[ W_{nn} = \alpha^2 W_n^2 k_o. \]

The ratio of total output signal-to-noise power may be defined as

\[ \lambda_o = \frac{\frac{1}{2} \alpha V_s^2 k}{\alpha^2 W_n^2 k_o + \alpha V_s^2 k} = \frac{\lambda_o^2}{k_o + 2\lambda_o} \]

where \( \lambda_o \) is defined as \( \lambda_o = \frac{1}{2} V_s^2/W_n \)

peak input signal power = average input noise power before gating.

If the pulses or the carrier is modulated sinusoidally at a small percentage modulation the output signal is

\[ V_o = \frac{1}{2} \alpha V_s(1 + a \cos \omega_m t)^2 \approx \frac{1}{2} \alpha V_s(1 + 2a \cos \omega_d t); \]

and the appropriate factor by which to multiply the ratio \( \lambda_L/\lambda_o \) is

\[ \frac{1}{\lambda_o^2} \left[ \int \frac{dI_o}{dy} \cos \omega_o t \right]^2 \]

By definition

\[ P^2 = \lambda_o \left[ \int \frac{d}{d\lambda_o} (\lambda F_1(-\frac{1}{2}; 1; -\lambda_0) - 1) \right]^2. \]

For some applications where the total energy in the pulse is of most importance, the ratio \( \lambda_L/\lambda_o \) itself is the most accurate basis of comparison for the two detectors. However, when the modulation frequency component is the desired or useful output, the ratio \( \lambda_L/\lambda_o P^2 \) gives a better measure of the relative merits of square-law and linear detection.

**B. Linear or Envelope Detector**

The analogous expressions for the linear detector are simply

\[ V_o = \sqrt{\frac{W_o}{2\pi} \int \left\{ F_1\left(-\frac{1}{2}; 1; -\frac{V_o^2}{2W_o}\right) - 1 \right\} \text{ during the pulses} \]

\[ = 0 \text{ otherwise.} \]

\[ W_{sn} = \frac{\alpha^2}{8\pi} k V_s^2 F_1\left(\frac{1}{2}; 2; -\frac{V_s^2}{2W_n}\right) \]

\[ W_{nn} = \frac{\alpha^2}{8\pi} \left[ k W_n F_1\left(\frac{1}{2}; 1; -\frac{V_s^2}{2W_n}\right) + (k_o - k) W_n \right] \]

\[ \lambda_L = \left( \frac{k_o - k}{k} \right) + F_1\left(\frac{1}{2}; 1; -\lambda_0\right) + 2\lambda_0 F_1\left(\frac{1}{2}; 2; -\lambda_0\right) \]

However, when a small sinusoidal modulation is applied, the signal-to-noise ratio for the output modulation component is not \( 2\lambda_La^2 \). In general, the output signal is

\[ \frac{a y \left( \frac{dI_o}{dy} \right) \cos \omega_o t,} \]

and the appropriate factor by which to multiply the ratio \( \lambda_L/\lambda_o \) is

\[ \frac{1}{\lambda_o^2} \left[ \int \frac{dI_o}{dy} \cos \omega_o t \right]^2 \]

\[ = \frac{1}{\lambda_o} \left[ \int \frac{d}{d\lambda_o} (\lambda F_1(-\frac{1}{2}; 1; -\lambda_0) - 1) \right]^2 \]

\[ = \frac{1}{\lambda_o} \left[ \int \frac{d}{d\lambda_o} \log (\lambda F_1(-\frac{1}{2}; 1; -\lambda_0) - 1) \right]^2 \]

\[ = \frac{1}{\lambda_o} \left[ \int \frac{d}{d\lambda_o} \log \left(\frac{d}{d\lambda_o} (\lambda F_1(-\frac{1}{2}; 1; -\lambda_0) - 1) \right) \right]^2. \]
a higher output signal-to-noise ratio than the square-law detector for equal pulse and gate widths, or for the uninterrupted carrier. This result is in direct contrast to many well-known proofs that the square-law detector is optimum, regarding signal-to-noise ratio. The reason for this discrepancy is easily explained, however. First of all the series expansions for the hypergeometric functions show clearly that at very low signal-to-noise ratios the output of the linear detector is the same as the output of a square-law detector in signal-to-noise ratio, for at very low signal-to-noise ratios

\[ I_s \to \alpha \sqrt{\frac{W_n}{2\pi}} [1 + \frac{1}{2} \frac{W_s}{W_n} + \cdots - 1] \]

\[ = \frac{\alpha W_s}{\sqrt{8\pi W_n}} \quad (43) \]

\[ W_{n,n} \to \frac{\alpha^2 W_n}{8\pi} \quad (44) \]

\[ \frac{I_s}{W_{n,1/2}} \to \frac{W_s}{W_n} \quad (45) \]

This result is identical to that obtained for a square-law detector. Naturally, signal-to-noise intermodulation is ignored at very low signal-to-noise ratios since noise-to-noise intermodulation is much larger. It is precisely because signal-to-noise intermodulation is ignored in these general proofs alluded to above that the results shown here are in contrast to them.

The reason for better signal-to-noise ratios from the linear detector is that the signal-to-noise intermodulation is of greater magnitude in the square-law detector than in the linear detector near unity signal-to-noise ratio, where it is beginning to be of importance. At very high signal-to-noise ratios also the two detectors give the same output modulation frequency signal-to-noise ratio, for at very high signal-to-noise ratios the linear detector gives,

\[ I_s \to \frac{\alpha \sqrt{2W_s}}{\pi} = \frac{\alpha V_s}{\pi} \quad (46) \]

\[ W_{n,n} \to \frac{\alpha^2 W_n}{\pi^2} \quad (47) \]

\[ \frac{I_s}{W_{n,1/2}} \to \frac{W_s}{2W_n} \quad (48) \]

whereas the square-law detector gives \( \sqrt{W_s/2W_n} \) as (22) and (26) show. However, when small percentage modulations about the mean value of input are considered, the square-law detector doubles the percentage modulation in the output, whereas the linear detector is linear at high signal-to-noise ratios in this respect. Thus, the factor of two above is cancelled out, and the two detectors give identical output signal-to-noise ratios for the output modulation.

Finally, let us note the serious decrease in performance of the linear detector in the case of modulated pulses with very large gates regarding the output modulation frequency signal-to-noise ratio. The explanation of this result is likewise simple. As Fig. 3 shows, the signal suppresses the noise when it is large relative to the noise in the linear detector. During a wide gate, the signal is on only a small fraction of the time and most of the noise is not suppressed—the noise-to-noise intermodulation noise, that is. For these cases the linear detector is definitely inferior to the square-law detector. Experimental work in the Electronics Department of the Hughes Aircraft Company has shown excellent agreement with the theory presented here, and has clearly demonstrated the inaccuracies resulting from applying square-law theory to many calculations on circuits using linear detectors.

APPENDIX I

RELATIONS BETWEEN THE CONFLUENT HYPERGEOMETRIC FUNCTIONS AND BESSEL FUNCTIONS

The expansions of the three confluent hypergeometric functions used in this paper in terms of imaginary Bessel functions are given below:

\[ iF_i(-\frac{1}{2}; 1; -x) = e^{-x/2} \left[ I_0 \left( \frac{x}{2} \right) + xI_0 \left( \frac{x}{2} \right) + x^2I_0 \left( \frac{x}{2} \right) \right] \quad (49) \]

\[ iF_i(\frac{1}{2}; 2; -x) = e^{-x/2} I_1 \left( \frac{x}{2} \right) \quad (50) \]

\[ iF_i(\frac{1}{2}; 1; -x) = e^{-x/2} I_0 \left( \frac{x}{2} \right) \quad (51) \]
A Video-Frequency Noise Spectrum Analyzer*

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Summary—A wave analyzer capable of measuring noise spectra in the video-frequency range is described. Part I of the paper discusses requirements in noise-analyzer design. Part II is a detailed description of a practical instrument.

A frequency range of 50 kc to 10 Mc is covered, without bandswitching, at virtually constant sensitivity. The width of the analyzing pass band, 33 kc, remains constant over the measuring range. The lowest spectral level that can be measured is 7 microvolts per root kilocycle.

INTRODUCTION

CERTAIN PROPERTIES of noise which have a direct bearing on the design of a noise-spectrum analyzer are found to differ considerably from those of periodic signals.† The first section of the present paper is devoted to a discussion of some of these properties and the requirements which the analyzer must fulfill. The relative advantages, in noise applications, of the tuned-circuit and the heterodyne methods of frequency analysis are compared; the heterodyne proves to be preferable. In Part II, a practical instrument is described.

PART I

MEASUREMENT OF NOISE SPECTRA

1. Basic Considerations

A wave analyzer should meet the following general conditions:

1. Adequate frequency coverage with maximum simplicity of tuning and stability.

2. Adequate sensitivity, approximately the same at all frequencies to be measured, with good gain stability.

3. Definition and identification of the measured frequency band.

4. Wide dynamic operating range.

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5. Suppression, or at least accurate evaluation, of errors.

The frequency distribution of a given signal is experimentally determined by passing the signal through a circuit whose response is confined to a narrow band of frequencies, and then examining the result. The band must be movable over the frequency range to be measured. A periodic signal yields a set of sinusoidal components, harmonically related in frequency, each having definite amplitude and phase.

For a nonperiodic function, such as random noise, no single sinusoidal component can be isolated. The output of the filter retains the erratic behavior of noise, no matter how narrow the pass band is made. While, for the periodic case, the mean-square amplitude of an isolated component is independent of the bandwidth of the filter, for noise the mean-square value of the filter output approaches, as the bandwidth is decreased, direct proportionality to the bandwidth.

II. Analytical Representation of Noise Spectra

These experimental facts imply that it is possible to express the time-average frequency distribution of a signal by a continuous function \( y(f) \), which can be so chosen that the mean-square value associated with that part of the spectrum lying between any two frequencies \( a \) and \( b \) is given by

\[
\langle E^2(a, b) \rangle_{av} = \int_a^b y^2(f) df. \tag{1}
\]

The time over which the average is taken is required to be large compared to \( a^{-1} \) and \( (b-a)^{-1} \). The definition of \( y^2(f) \) is

\[
y^2(f) = \lim_{\Delta f \to 0} \frac{\langle E^2(f, \Delta f) \rangle_{av}}{\Delta f} \tag{2}
\]

where \( \langle E^2(f, \Delta f) \rangle_{av} \) is the mean-square value of those components lying in a band of width \( \Delta f \) and of center frequency \( f \). Again the time over which the average is taken must be large compared to \( f^{-1} \) and \( (\Delta f)^{-1} \). The function \( y^2(f) \) is the mean-square spectrum (or power spectrum) of the noise signal \( e \); it is a continuous function of frequency. If the noise signal consists of a voltage-time function, then \( y^2(f) \) has the units volts-squared-per-unit-frequency. In studying the frequency distribution of various noise sources, the rms spectrum \( y(f) \) (referred to as the spectral level at frequency \( f \)) is of interest. Its measurement at a single, sharply defined frequency is an unpleasant idealization, since the measuring bandwidth is necessarily finite.

The spectrum defined contains no quantity comparable to phase in the periodic case; our representation
is analogous to the power spectrum, consisting of the squares of the amplitudes.

III. Integrated Response of Circuits to Noise

We shall consider below a parameter called the "integrated response" of a circuit to noise. We proceed as follows: The mean-square value of the output of a circuit of amplitude response \( m(f) \) to which noise having spectral level \( y(f) \) is applied is, from (1),

\[
[E^2]_{av} = \int_{0}^{\infty} m^2(f)y^2(f)df. \tag{3}
\]

Suppose that \( m \) has an appreciable value only within a certain range of frequencies. If the input spectrum is of constant level \( y_o \) over this range,

\[
[E^2]_{av} = y_o^2 \int_{0}^{\infty} m^2(f)df. \tag{4}
\]

The integrated response \( M \) of the circuit is defined as

\[
M = \sqrt{\int_{0}^{\infty} m^2(f)df}. \tag{5}
\]

This quantity relates the rms value of the output noise to the input spectral level, provided the latter is constant over the pass band of the circuit:

\[
E = My_o. \tag{6}
\]

\( M \) may be calculated from (5). The integral may be computed when the analytic form of \( m(f) \) is known, or it may be evaluated graphically from sine-wave determinations of \( m(f) \). The value of \( M \), together with a measurement of \( E \), permits calculation of \( y_o \). This is the basis for the design of a noise-spectrum analyzer.

IV. Errors Due to Finite Bandwidth of Filter

If noise is applied to a narrow-band-pass filter, then the rms output of the filter, together with the integrated response of the filter, gives approximately the spectral level at the center frequency of the filter

\[
y_o = \frac{E}{M} + \varepsilon(y, m). \tag{7}
\]

The error \( \varepsilon \) is a function of the filter characteristic and that part of the spectral distribution which lies within the pass band of the filter. Two theorems concerning this error, which have some practical importance, are as follows:

1. For a filter amplitude response that is symmetric about a frequency \( f_o \), only the even derivatives of \( y(f) \) computed at \( f_o \) contribute to the error.

2. The error may be attributed to an error in the frequency at which the spectral level is measured; this error in frequency is always smaller than the filter pass band, and, for a symmetric characteristic, is smaller than half the pass band:

\[
y(f) = \frac{E}{M}, \quad |f_1 - f_o| < \text{pass band}. \tag{8}
\]

These theorems depend on the continuity of the noise spectrum; they fail, for example, if a periodic component lies within the pass band.

The narrower the pass band, the more accurate the measurement of the spectrum. The width of the pass band has a lower limit, set by one or the other of two considerations: sensitivity, and meter fluctuations. The effective noise sensitivity of a filter is proportional to the square root of its bandwidth. More restrictive are fluctuations in the filter output. We may represent the noise output of a narrow filter as a sine wave, the amplitude of which varies in a random manner with time:

\[
e = R(t) \sin 2\pi ft. \tag{9}
\]

The frequency \( f_o \) is the center frequency of the filter, and the envelope \( R(t) \) consists of noise having frequency components between zero and one-half the bandwidth of the filter. These low-frequency components cause objectionable fluctuations in the final measuring circuit. The narrower the filter, the more prominent are the lower frequencies in the spectrum of \( R(t) \), and the greater the corresponding time required to obtain a measurement of prescribed accuracy.

The problem of measuring noise spectra is seen to reduce essentially to the design of a band-pass filter of variable center frequency, with band pass narrow enough to yield accurate measurements, but wide enough to satisfy practical requirements of sensitivity and reasonable time constants.

V. Basic Circuits

Three fundamental circuits are available to meet the problem: the resonant inductance-capacitance-resistance circuit,\(^6\) the selective feedback-amplifier circuit employing a bridged-\(T\) or twin \(T\),\(^9\)\(^11\) and the heterodyne circuit.\(^9\)\(^11\)\(^12\)\(^16\)

A high-selectivity inductance-capacitance resonant circuit satisfies conditions 3, 4, and 5. The measured frequency band is uniquely determined, the dynamic range of the input is restricted only to amplitudes for which the circuit components remain linear, and the response of the circuit to noise can be accurately calcu-


ated from sine-wave calibration. For either series or parallel circuits, however, conditions 1 and 2 are extremely difficult to satisfy simultaneously.

The bridged-T circuit, being a null-transmission arrangement, must be employed as a feedback element in a high-gain amplifier. For the frequency range under consideration, this method is difficult to apply.

The heterodyne circuit shown in Fig. 1 can easily be made to satisfy conditions 1 and 2 over very wide frequency ranges. By suitable care in design, adequate performance with regard to conditions 3, 4, and 5 may be secured. The heterodyne system has the advantage that the width and character of the measured band are not functions of the center frequency. The ultimate measuring bandwidth is determined at audio frequency, where characteristics are readily controlled. Variations of the integrated noise gain with frequency depend only on the frequency response of the first modulator and the input filter.

VI. Considerations Pertinent to the Heterodyne Circuit

(a) Frequency Identification. The identification of the measured band is evident from consideration of Fig. 1. The intermediate-frequency band is placed between the range to be measured and the local oscillator fundamental, and certain of the input signal components, as shown, lie within the intermediate-frequency band. To ensure that only frequencies below the intermediate frequency are measured, a low-pass filter designed to cut off slightly below the intermediate frequency is inserted ahead of the first modulator. The measured components are then uniquely determined, and all other possible primary modulation components are suppressed. Under certain conditions, secondary products due to frequency components outside the band under scrutiny may cause errors in measurement. This method, using simple difference-frequency components, permits measurement of the entire range from near zero frequency up to the intermediate-frequency band by tuning the local oscillator over a frequency ratio of only 2 to 1.

(b) Effect of Internal Interference on Sensitivity. The maximum sensitivity that may be obtained is intimately connected with the interference that originates within the instrument. This interference may be periodic, random, or both.

The effectiveness of the interference in obscuring the signal depends considerably on the output-circuit design. The choice of the final meter circuit lies essentially between the linear rectifier and the quadratic rectifier. When the signal is smaller than the interference, it is more completely obscured in the response of the linear rectifier than in the quadratic rectifier; also, the relation between the signal and the indication is more complicated. In the square-law circuit, the average output is just the sum of the mean-square values of the signal and interference. By use of a suitable bridge circuit, it is possible to cancel out the average effect of the interference, and leave a response due entirely to the signal. This cannot be done conveniently with a linear rectifier when the signal is smaller than the interference. Partially, in consequence of these considerations, the quadratic bridge circuit was adopted in the practical design.

There is a limit to the canceling-out process because of the fact that all amplifier stages must handle the entire interference-plus-signal amplitude, prior to cancellation, with linearity. As the signal becomes smaller this amplitude becomes larger, to produce sufficient output from the signal to actuate the meter circuit. Eventually, some stage will be overdriven because of the amplitude of the interference alone. If the signal and interference are both noise, then the least measurable signal level is independent of the analyzer bandwidth. When the minimum measurable output voltage is small compared with the overdrive voltage, then the least measurable level is proportional to $b/a$.

A second limitation is associated with fluctuations of the output indicator due to interference of the random variety and to the random nature of the noise signal under observation. Such fluctuation may be reduced by use of a large time constant in the meter circuit. A compromise is necessary, however, as the time constant is limited by the practical consideration of the time required to obtain a reading, and ultimately by the rate at which the over-all sensitivity of the analyzer drifts. In the instrument as constructed, signals 6 db below the interference level could be accurately measured.

(c) Errors Due to Nonlinearity. The entire spectrum of the input signal which falls within the frequency range to be investigated is applied to the input of the first modulator. Thus the total signal at the first modulator may be many times larger than that portion of it which is effective in giving a measurement of the spectral level at any specific center frequency. In some cases, the resulting signal may be large enough to produce appreciable secondary modulation products in addition to the desired primary ones. In extreme cases, the signal may be large enough to cause overload of the modulator, with consequent reduction of the over-all gain of the analyzer. Simple broad filters of low attenuation


placed ahead of the first modulator will minimize this by restricting the signal to the frequency band of interest at the moment. An important measure to reduce the significance of these effects is the introduction of attenuation ahead of the first modulator. It is preferable to insert all attenuation at this point and operate subsequent amplifiers at full gain.

The error in measurement introduced by the presence of secondary modulation products is significant when dealing with portions of a spectrum which are considerably lower than the rest, such as narrow, deep holes, and the cutoff region of a sharp-cutoff filter. It is frequently found that the hole, as measured, appears narrower and less deep for noise than for sine waves. This effect is explained by the fact that, along with the primary modulation products, there is always a set of secondary products, consisting of noise components beating with each other and sometimes also subsequently with the local oscillator. These secondary products arise from frequencies outside the band nominally under measurement; they lie within the intermediate-frequency pass band and are indistinguishable from the primary ones. When the frequency region being measured has about the same amplitude as the rest of the spectrum, the secondary products are always drowned out by the primary ones; however, at a hole in the spectrum, the primary products may be so reduced as to be small compared with the secondary products. The reading then will be too large, and the hole will appear to be filled in; the deeper and narrower the hole, the smaller the primary components, and the larger the region which contributes secondary components. This same effect at the edge of a band-pass filter makes the noise appear to spill over outside the filter sine-wave characteristic. This type of error, caused by signal components outside the nominal measurement band, can be completely eliminated only by suppressing the extraneous frequencies with a band-pass filter preceding the first modulator. It can be minimized by careful choice of circuit and operating conditions. Furthermore, it can readily be detected by introducing additional attenuation ahead of the first modulator. If second- and higher-order products are of significant magnitude, the output of the analyzer will fall off more rapidly than the added attenuation would justify.

A somewhat different type of error is introduced when overload conditions occur. The over-all analyzer gain having been reduced by the overloading, peaks in the spectrum will be flattened out, and will appear to be of smaller amplitude than they really are. Actually, all parts of the spectrum would appear to be of lower than true amplitude, but, except at the peaks, there is an excellent chance that the extraneous secondary products discussed above will contribute so much energy to the band under measurement that the effect of gain reduction due to overload will not be very marked. Overload occurs only for larger amplitudes than those which produce objectionable secondary products. Hence, the net result of overload is to flatten out a spectrum containing peaks and holes. Overload can easily be detected, since it results in a nonlinear relationship between the input attenuator setting and the output reading. In this case, the output will decrease less rapidly than added attenuation at the input would justify.

If the signal mean-square voltage $E_s^2$ is greater than the maximum allowable mean-square voltage $E_m^2$ which may be applied to the modulator, then only the fraction $E_m/E_s$ of the signal amplitude $E_s$ can be applied to the first modulator. The expression for the minimum measurable signal amplitude is higher than before by the factor $E_s/E_m$. It should be noted that $E_m$ is dependent on the noise spectrum of the signal.

(d) Dynamic Range. The dynamic range of an instrument refers to the range of input amplitudes over which it may be used. The dynamic range may be extended indefinitely in the direction of large amplitudes by increasing the available attenuation ahead of the first modulator; in the other direction, little increase of sensitivity is gained by the use of untuned preamplifiers. The noise in the first modulator, which is the principal source of interference originating within the instrument, is only slightly above that of an amplifier stage, provided the modulator has been carefully designed. As we have seen, the range of spectral levels which can be measured in any particular case depends greatly on the nature of the signal. If large variations in level over the spectrum are present, removal of sufficient attenuation to permit measurement of the low-level portions may result in overdrive of the first modulator. A high peak-to-rms ratio leads to overdrive at lower spectral levels than a low ratio. Whenever the level of the secondary products that lie in the intermediate-frequency pass band becomes comparable to that due to the primary components, serious errors will occur in the measurement. The significant quantity remains the minimum accurately measurable level, which can be computed from the effective peak-to-rms ratio of the signal, the maximum allowable input to the first modulator, and the maximum analyzer gain.

VII. Summary

The foregoing comparison between the heterodyne and tuned-circuit methods of noise-spectrum analysis may be summarized as follows: the heterodyne method is greatly superior for obtaining wide frequency coverage at constant sensitivity and with simplicity of tuning. It is inferior in dynamic operating range and in the suppression of errors, but design procedures giving satisfactory performance with respect to these qualities can be worked out. The tuned circuit has the disadvantage that the entire amplifying and measuring system must be designed to operate at all frequencies in the range to be measured. In the heterodyne system, all circuits after the first modulator operate over fixed, narrow frequency bands. For the present purpose, the heterodyne circuit is decidedly preferable.
the source, at least up to 10 Mc. A noise signal 10 Mc wide of uniform level 0.1 volt per root kc has an rms value of 10 volts, and may have an effective peak value of 20 or 25 volts in the case of Gaussian noise. In Part I it was shown that, given optimum design at subsequent circuits, the larger the nonoverdriving voltage which can be applied to the modulator, the lower the minimum measurable spectral level. The input probe should accordingly be designed not to overload at a lower voltage than the first modulator. In the instrument developed, the 6Y6G cathode follower handles 6 volts peak-to-peak at the input, driving a 100-ohm load with a voltage amplification of 0.3. The input capacitance is about 11 $\mu$F, and the input resistance is greater than 0.1 megohm. The cathode follower is built into a probe unit which may be brought into close physical proximity of the source of signal being measured.

II. High-Frequency Attenuator

The high-frequency attenuator is an integral part of the measuring system of the analyzer. Readings are taken by setting this attenuator so that the output meter indicates some standard value. The analyzer gain may be so adjusted that the loss introduced by the attenuator gives the input spectral level directly in decibels above a suitable arbitrary level. Location of the attenuator ahead of the first modulator provides the greatest possible dynamic operating range, but involves the difficulty of covering a wide band of frequencies. The attenuator is designed to provide a range of 70-db in 1-db steps at an impedance level of 100 ohms; there is negligible frequency discrimination up to 10 Mc.

The several sections consist of pi networks, individually shielded, each being controlled by a toggle switch. By connecting composition resistors directly to the lugs of the switches and soldering the returns securely to a heavy copper base plate, inductance and common-path effects were readily overcome.

III. Balanced Modulator

A balanced modulator (Fig. 4) is used as a mixer in order to permit measurement at such low frequencies that the local oscillator falls within the pass band of the intermediate-frequency amplifier. The balanced modu-

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**Fig. 2**—Front-panel view of the video-frequency noise-spectrum analyzer.

**Fig. 3**—Block diagram of the video-frequency noise-spectrum analyzer.

**Fig. 4**—Balanced-modulator schematic diagram.
lator prevents the carrier of the local oscillator from appearing in the modulator output and overloading the intermediate-frequency amplifier.

Before being applied to the balanced modulator, the signal must be passed through a low-pass filter to prevent components at frequencies in the intermediate-frequency band or at image frequencies from contributing to the output of the modulator. The integrated response of the filter in the intermediate-frequency band of frequencies is about 100 db below the pass band. The filter response is flat within 1 db up to 9.3 Mc.

A signal balanced to ground for application to the balanced modulator is obtained by passing the output of the low-pass filter through a phase splitter (V₁ of Fig. 4). The difference in output impedance of the plate and cathode circuits (the latter, effectively a cathode follower, has a much lower impedance) results in difficult and critical balance of the modulator input circuits if the phase splitter is coupled directly to the modulator. The difficulties appear to be due to space-charge coupling between the No. 1 and No. 3 grids of the 6SA7 converter tubes. The addition of an extra cathode follower (V₂ of Fig. 4), between the plate circuit of the phase splitter and the appropriate modulator grid, eliminates the need for neutralization of the space-charge coupling, and gives satisfactory operation.

In the design of the modulator circuit, 6SA7 pentagrid converter tubes (V₁ and V₄ in Fig. 4) are used to facilitate isolation of the signal circuits from the local oscillator. Means are provided for balancing both phase and amplitude. Amplitude balance is accomplished by means of a capacitance voltage divider which controls the relative proportion of the local-oscillator carrier fed to each converter tube. The phasing is done by means of a balancing capacitor across the output tank circuit. Both coarse and fine controls are provided for both amplitude and phase balancing.

Although exact settings of the balancing controls for optimum suppression of the carrier depend slightly on the frequency of the local oscillator, usually the balance obtained at some frequency in the neighborhood of 50 kc holds adequately over the entire range. Balancing is not required at all for readings at frequencies above 500 kc.

Provided the local-oscillator output remains reasonably constant over the tuning range, the modulator conversion gain is not dependent on frequency. The complete balanced modulator circuit is found to have a constant conversion gain from 10 cps to 10 Mc. Thus, with input circuits flat over the range to be measured, it is possible to realize essentially constant sensitivity.

IV. Variable-Frequency Local Oscillator

The chief requirements in the design of the local oscillator (Fig. 5) are frequency stability and constancy

of output over the tuning range. Stability is particularly important at the low-frequency end, since a small drift in oscillator frequency produces a large fractional change in the frequency of the band being measured.

Constancy of output over the tuning range is needed because, even with optimal choice of oscillator drive for the mixer tubes, the conversion transconductance of the latter varies with oscillator voltage. It was found that a typical grounded-plate Hartley oscillator operating between 10 and 20 Mc delivered approximately twice the voltage output at the high-frequency as at the low-frequency limit. A frequency-sensitive degenerative feedback network (C₄, C₅, and R₁ of Fig. 5) is added to make the output substantially constant over the band. By adjustment of C₄, the outputs can be made equal at any selected pair of high and low frequencies. By this means, a deviation of less than 1 db can be procured over the entire tuning range. The resistor R₁ acts both as part of the feedback circuit and also as the bias resistor for the oscillator.

It is necessary to shield the oscillator unit carefully and by-pass all power leads, including the heaters, thoroughly, in order to prevent undesired coupling with the rest of the instrument.

V. Intermediate-Frequency Circuits

The intermediate-frequency amplifier employs two transformer-coupled stages giving a gain of 60 db at the center frequency of 9,875 kc.

The second modulator employs a 6SA7 pentagrid converter tube, with capacitance neutralization of the space-charge coupling between the oscillator and signal grids to prevent crystal oscillator voltage from appearing on the signal grid. A voltmeter may be switched to measure the bias on either grid. The bias on the oscillator grid is a measure of the drive secured from the crystal oscillator. The direct voltage developed by the signal grid is a useful preliminary indication of unbalance in the first modulator.

The crystal oscillator operates at 9,875 kc, the center of the intermediate-frequency-amplifier pass band. Thus frequencies both above and below this are converted to audio frequency in the mixer.

VI. Audio Circuit

The audio circuit comprises a set of low-pass filters, an attenuator, an amplifier, and a square-law meter circuit.

The width of the sample of the input-signal spectrum which is measured is determined by the low-pass filter between the second modulator and the audio-frequency amplifier. The sample is just twice the bandwidth of the filter. Provision of a choice of several filters having different cutoff frequencies, specifically, 5 and 15 kc, permits the sample bandwidth to be varied.

A conventional amplifier employing two 6SJ7 pentodes in cascade to feed a 6J5 triode output stage provides the necessary audio-frequency amplification.

An attenuator providing up to 30 db loss in three 10-db steps is used as an auxiliary to the main video-frequency attenuator. A continuous gain control permits adjustment of the over-all sensitivity of the instrument to a predetermined standard calibration value.

The output indicator circuit consists of a thermistor bridge. This has several advantages. It exhibits an accurately quadratic characteristic, high sensitivity, and tolerance of severe overload, and it maintains constant sensitivity over extended periods of use. Further, the bridge circuit can be rebalanced without affecting the sensitivity, so that the average effects of interference arising within the analyzer are suppressed. The thermistor arms of the bridge circuit are mounted within a temperature-controlled box in order to eliminate the effects of changes in the ambient temperature. The use of two thermistors permits cancellation of the effect of the small temperature changes which occur in the box due to lag in the thermostat response. Only one thermistor receives the signal to be measured.

A useful range of 10 db may be covered by the indicating scale. An input power of about 2 milliwatts will provide full-scale deflection of the meter. This power is readily supplied by the 6J5 final audio-amplifier tube with an excellent reserve capability for handling peaks. It is evident that any interference voltage present, whether originating internally or externally, may be combined with the direct current in the measuring thermistor to obtain the original bridge balance. The addition of signal, even if considerably weaker than the interference, then produces a reading. The effect of interference evidently cannot be suppressed if, at an earlier point in the analyzer, it and the desired signal intermodulate.

VII. Conclusion

The experience gained with two models of the video-frequency spectrum analyzer has shown that the instrument has a maximum sensitivity of 7 microvolts per root kilocycle with high impedance input. The sensitivity is constant to ±0.5 db over the range 50 kc to 9.3 Mc. The internal noise level, referred to the modulator input, is 5 microvolts per root kilocycle, or 15 microvolts per root kilocycle referred to the cathode-follower input.

The attenuators permit measurements over a wide range of signal-input levels, while the probe input permits these to be made with little disturbance of the signal source under investigation. The range of levels which the instrument will measure has a minimum set by the sensitivity, and a maximum determined by the voltage at which intermodulation begins. The maximum tolerable input is determined in terms of the effective peak voltage applied to the input circuit. Hence, the wave form and frequency distribution will affect the spectral level at which intermodulation occurs. It may occur that components in a remote part of the spectrum will overdrive the input before those under consideration can be measured, even though the latter be above the threshold level. If a noise signal is connected of which the effective peak value is four times its rms value and the spectrum is flat out to and negligible above 10 Mc, the greatest measurable spectral level is 7.5 millivolts per root kilocycle.

As long as the principal modulation products of the balanced modulator are the result of the signal noise components beating with the local oscillator, and not with each other, holes in the spectrum will be accurately measured; if a hole is narrow and deep, however, secondary-modulation products, consisting of difference frequencies between noise components beating with the local oscillator or of sum frequencies, may be sufficient to make the hole appear considerably filled in. No corresponding difficulty appears with peaks, for which second-order effects are always negligible. The presence of peaks may, however, interfere with measurements elsewhere in the spectrum.

The analyzer is not designed to measure signals with wave forms having large peak-to-rms ratios, such as radar pulses.

The exceptional tuning range without bandswitching and with constant sensitivity simplifies the use of the analyzer markedly. The method of determining bandwidth gives an almost ideal square-topped characteristic whose width is accurately known. If desired, selection between several such widths is possible. The output measuring circuit is simple, reliable, and rugged.
Terminal Impedance and Generalized Two-Wire-Line Theory*

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Summary—Conventional transmission-line theory takes no account of variations in the parameters of the line and of coupling near a terminating impedance, or near any other change in its uniform properties. A theory is derived which substitutes for these effects a simple terminal-zone network of lumped series and shunt elements which may be evaluated for each type of termination or discontinuity. The apparent terminal impedance $Z_a$ (which is the impedance actually measured on a lossless line at a distance $X/2$ from the termination) consists of this network in combination with the theoretical, isolated impedance, $Z_o$, of the load. $Z_a$ for a fixed termination with a given $Z$, may vary greatly with the nature of the connection to the line, the relative orientation of line and load, and the type of line and its dimensions.

INTRODUCTION

The derivation of the transmission-line equations involves fundamental assumptions that are often overlooked. Among them is the requirement that all elements of length $A\xi$ of a two-wire line be identical. Since this is not true near the ends of the line at $z = 0$ and $z = s$ where it joins the generator and load, the conventional differential equations and the currents, voltages, and impedances derived from them are not accurate in terminal zones at each end. The length $d$ of each zone depends on the termination; it is of the order of magnitude $10b$, where $b$ is the spacing of the two-wire line, provided the terminations are completely outside the region bounded by the planes $z = 0$ and $z = s$. In each terminal zone the parameters of the line vary, and the line may be coupled to adjacent parts of the termination. A physically significant impedance may be defined only outside the terminal zones; i.e., at such distances from the load that the scalar potential difference between opposite points on the two conductors is equal to the line integral of the electric field from one point to the other.

The apparent terminal impedance of a particular impedor may be studied using Fig. 1 without specializing the formulation to this particular circuit. The load in this figure is to the right of the plane containing the line-load junction.

An essential postulate implied in the conventional equations is that the line must be balanced. Electrically, this means

$$q_{2L}(w') = -q_{1L}(w'); I_{2L}(w') = -I_{1L}(w'), \quad (1)$$

where $q_{2L}(w')$, $[q_{2L}(w')]$ is the charge per unit length on conductor 1 [2] of the line at a distance $w'$ from the load-line junction. $I_{1L}(w')$, $[I_{2L}(w')]$ is the total current in conductor 1 [2] at the same cross section. Distributions in the line that satisfy (1) involve restrictions on the charges and currents in the termination and, hence, on the electromagnetic field associated with all charges and currents in the entire circuit. Similar restrictions apply to the scalar and vector potentials as follows: All parts of the line, load, generator, and associated circuits must be so constructed that the scalar potential $\Phi$ and the axial component $A_z$ of the vector potential, at corresponding points on the surfaces of the conductors of the line, satisfy the following conditions:

$$\Phi_z(w) = -\Phi_1(w); A_{2z}(w) = -A_{1z}(w). \quad (2)$$

The potential differences are then given by:

$$V'(w) = \Phi_1(w) - \Phi_2(w) = 2\Phi_1(w); \quad (3)$$

$$W'_z(w) = A_{1z}(w) - A_{2z}(w) = 2A_{1z}(w).$$

The total potentials and potential differences are the sums of components due to charges and currents on the line (subscript $L$) and on the terminations (subscript $T$). For example,

$$V'(w) = V'_T(w) + V'_T(w); \quad (4)$$

Therefore, (2) and (3) are true if a subscript $L$ or $T$ is added to each potential.

The conditions (2) with added subscript $T$ imply that the load (or generator) consists of identical halves 1 and 2 of which #1 is attached or coupled to conductor 1 of the line, #2 to conductor 2 of the line. The following con-
Rations must obtain at equal distances \( u' \) from the junctions with the line along the axes of the conductors of each half of the termination:

\[
q_{2r}(u') = - q_{1r}(u') \quad \text{and} \quad I_{2r}(u') = - I_{1r}(u')
\]

\[
I_{2r}(u') = I_{1r}(u') \quad \text{and} \quad I_{2r}(u') = I_{1r}(u')
\]

(5)

If the halves are geometrical images in the plane \( y = 0 \), put with signs of charges and directions of currents opposite to those of mirror images, all conditions (1)–(5) are satisfied. However, there are configurations in which the halves are not geometrical images in the plane \( y = 0 \), that also satisfy conditions (1)–(5).

**Generalized Potential Differences**

Referring to Fig. 1, the potential differences between points \( P_{1L} \) and \( P_{2L} \) on the surfaces of the conductors of the line are given by (4) with

\[
W_{sL} = \frac{1}{2\pi} \int_0^\pi \int_0^\pi I_{sL}(w') P_{L}(w, w') dw'
\]

\[
W_{rL}(w) = \frac{1}{2\pi} \int_0^\pi \int_0^\pi I_{rL}(w') P_{T}(w, u') du'
\]

\[
V_{L}(w) = \frac{1}{2\pi} \int_0^\pi \int_0^\pi q_{L}(w') P_{L}(w, w') dw'
\]

\[
V_{r}(w) = \frac{1}{2\pi} \int_0^\pi \int_0^\pi q_{r}(w') P_{T}(w, u') du'
\]

where

\[
P_{L}(w, w') = \left[ \frac{e^{-\beta R_a}}{R_a} - \frac{e^{-\beta R_b}}{R_b} \right]
\]

\[
P_{T}(w, u') = \left[ \frac{e^{-\beta R_{1T}}}{R_{1T}} - \frac{e^{-\beta R_{2T}}}{R_{2T}} \right]
\]

\[
\beta = \omega \sqrt{\frac{\xi}{\nu}} \quad \xi = \epsilon_\epsilon - j\sigma/\omega
\]

and where \( \nu = \nu_{w'} \), \( \nu_0 = 10^7/4\pi \) meters/second; \( \epsilon_\epsilon = \epsilon_{r\epsilon} \), \( \epsilon_0 = 8.85 \times 10^{-12} \) farads/meter; \( \nu = 1/\mu_r \) is the relative permeability, \( \epsilon_{r\epsilon} \) is the relative effective dielectric constant; and \( \sigma_r \) is the effective conductivity in mhos/meter of the medium in which the entire circuit is immersed.

The distances \( R_a = \sqrt{(w-w')^2 + a^2} \), \( R_b = \sqrt{(w-w')^2 + b^2} \), \( R_{1T} \) and \( R_{2T} \) are in Fig. 1; \( z \) is the length of the line, \( s_T \) is half the distance around the contour of the termination.

In order to evaluate (6) and (7), the charges and currents at \( w' \) in the line and at \( u' \) in the termination are expanded in series. Currents and charges are continuous at the line-load junction so that

\[
q_{L}(w' \to 0) = q_{T}(w' \to 0)
\]

\[
I_{sL}(w' \to 0) = I_{sT}(w' \to 0)
\]

(10)

Using the equation of continuity, \( dI_{sL}/ds + j\omega q_{L} = 0 \), the following expansions may be written:

\[
q_{L}(w') = q_{L}(w) + (w' - w) \frac{1}{i\omega} \frac{\partial I_{sL}(w)}{\partial w^2}
\]

\[
I_{sL}(w') = I_{sL}(w) + (w' - w) j\omega q_{L}(w)
\]

\[
q_{r}(w') = q_{r}(w) - (u' + w) \frac{1}{j\omega} \frac{\partial I_{sL}(w)}{\partial w^2}
\]

\[
I_{sT}(w') = I_{sT}(w) - (u' + w) j\omega q_{L}(w)
\]

(11)

Note that

\[
I_{sT}(w') = I_{sT}(w') \cos \psi(w')
\]

where \( I_{sT}(w') \) is the total axial current at \( u' \) in the termination, and \( \psi(w') \) is the angle between the direction of the current at \( u' \) and the \( z \) axis.

Substitution of (11) and (12) in (6) and (7), and subsequent substitution of the resulting integrals in (4), gives:

\[
W_{s}(w) = \frac{1}{2\pi} \left[ I_{sL}(w) [k_0(w) + k_{sr}(w)] + j\omega q_{L}(w) [k_1(w) + k_{1T}(w)] /\beta \right]
\]

\[
V(w) = \frac{1}{2\pi} \left[ q_{L}(w) [k_0(w) + k_{sr}(w)] + \frac{1}{j\omega} \frac{\partial I_{sL}(w)}{\partial w} [k_1(w) + k_{1T}(w)] /\beta \right],
\]

(14)

where

\[
k_0(w) = \int_0^s P_{L}(w, w') dw' = \int_0^s \left( \frac{1}{R_a} - \frac{1}{R_b} \right) dw'
\]

\[
= \sinh^{-1} \frac{w}{a} - \sinh^{-1} \frac{w}{b} + \sinh^{-1} \frac{z}{a}
\]

\[
= \sinh^{-1} \frac{2}{b}
\]

(15a)

\[
k_{sr}(w) = \int_0^s P_{T}(w, u') \cos \psi(u') du'
\]

\[
\pm \int_0^s \left( \frac{1}{R_{1T}} - \frac{1}{R_{2T}} \right) \cos \psi(u') du'
\]

(15b)

\[
k_{sr}(w) = \int_0^s P_{T}(w, u') du'
\]

\[
= \int_0^s \left( \frac{1}{R_{1T}} - \frac{1}{R_{2T}} \right) du'
\]

(15c)

\[
k_1(w) = \beta \int_0^s (w' - w) P_{L}(w, w') dw'
\]

\[
= \beta \int_0^s (w' - w) \left( \frac{1}{R_a} - \frac{1}{R_b} \right) dw'
\]

\[
= \beta \left[ \sqrt{(w^2 + b^2 - \sqrt{w^2 + a^2} - \sqrt{2b^2 + a^2} + \sqrt{2b^2 + a^2})} \right]
\]

(15d)

\[
k_{1T}(w) = - \beta \int_0^s (u' + w) P_{T}(w, u') \cos \psi(u') du'
\]

(15e)

\[
k_{1T}(w) = - \beta \int_0^s (u' + w) P_{T}(w, u') \cos \psi(u') du'
\]

characterize the transmission-line-end effect for a line in air where \( \beta = 0 \) (but do not include coupling to the terminal) are shown in Fig. 2. The corresponding values of \( I^s(w) \), \( C_0(w) \), and \( R_0(w) = [I^s(w)/C_0(w)]^{1/2} \) are in Fig. 3.

The approximate integrals in (15) differ negligibly from the exact integrals if \( \beta b \) satisfies the following inequality:

\[ |\beta b| \ll 1. \]  

The terms neglected on the right in (15) contribute to radiation from the line; (16) makes this negligible. Let

\[ I^r(w) = I^s(w) + I^t(w) = [k_0(w) + k_{ar}(w)]/2\pi v \]  

\[ k_{ar}(w) = \beta \int_0^w (u + w) P^r(w, u) du' \]  

\[ = -\frac{1}{R_{1T}} \int_0^w (u + w) \left( \frac{1}{R_{1T}} - \frac{1}{R_{2T}} \right) \cos \psi(u') du'. \]  

Note that

\[ \beta^2 = \omega^2/\nu = -j\omega \beta \]  

With (17) substituted in (13) and (14), the following expressions are valid at all points along the line:

\[ W_s(w) = I^r(w) \left[ I_{sl}(w) + j\omega L_s(w) p(w)/\beta \right] \]  

\[ V(w) = \frac{j\omega}{y(w)} \left[ q_L(w) + \frac{1}{j\omega} \frac{\partial^2 I_{sl}(w)}{\partial w^2} \right] p'(w). \]  

For some purposes, the ratio functions \( a_1(w) \) and \( \phi_1(w) \) are useful.

\[ a_1(w) = I^r(w)/I^s(w) \quad W_s(w)/W_{sl}(w); \]  

\[ \phi_1(w) = c(w)/c_0(w) \quad \phi_{L}(w)/V_{L}(w). \]  

When \( a_1(w) = 1 \), there is no inductive coupling; when \( \phi_1(w) \), there is no capacitive coupling between line and termination.

At points outside the terminal zones, where conventional line theory is adequate, these relations reduce to

\[ W_s(w) = I^s(w) I_{sl}(w); \quad V(w) = j\omega q_L(w)/y_0. \]  

\[ I^s = k_0(w \to \infty) / 2\pi = (\ln b/a)/2\pi v \]  

\[ y_0 = \gamma_0 + j\omega c_0 = j\omega \beta \]  

\[ = j\omega \beta / [\ln (b/a)]. \]  

The functions \( k_0(w) \), \( k_1(w)/\beta \beta b \), and \( p_0(w)/\beta \beta b \) which characterize the transmission-line-end effect for a line in air where \( \beta = 0 \) (but do not include coupling to the terminal) are shown in Fig. 2. The corresponding values of \( I^s(w) \), \( C_0(w) \), and \( R_0(w) = [I^s(w)/C_0(w)]^{1/2} \) are in Fig. 3.

Fig. 2—The functions \( k_0(w) \), \( k_1(w)/\beta \beta b \), at the load end of a long line in air which is not coupled to the load.

Fig. 3—The external inductance and capacitance per unit length and characteristic resistance at the load end of a line in air with no coupling to the load, \( k_0(w) = 2 \times 10^{-7} \text{ henry/m} \); \( c_0(w) = 2 \pi \times 8.85 \times 10^{-12}/k_0(w) \text{ farads/m} \); \( R_0(w) = \sqrt{I_0^2(w)/C_0^2(w)} \).

**Generalized Differential Equations for Voltage and Current**

The generalized differential equations for current and voltage may be derived from the equations defining the scalar and vector potentials, namely, \( E = -\text{grad} \Phi \) and \( A + j(\beta^2/\nu) \Phi = 0 \); and the relation, \( E_{t1} = z_1^t I_{t1} \), where \( z_1^t \) is the internal impedance per unit length of conductor 1 and \( E_{t1} \) is the tangential electric field at its surface. Noting that \( z_1^t = Z_1^t = z_1^t/2 \), where \( z_1^t \) is the linear impedance per unit length of conductor 1 and \( Z_1^t \) is the characteristic impedance of the line, one obtains

\[ E_{t1} = \frac{I_{t1}}{z_1^t}. \]
the internal impedance per loop unit length of the line, and using (1), (3), and \( w = s - z \), the following equations are obtained:

\[
\frac{\partial V(w)}{\partial w} = z I_{sl}(w) + j\omega W_{s}(w) \quad (23)
\]

\[
\frac{\partial W_{s}(w)}{\partial w} = \frac{j\beta^2}{\omega} V_{L}(w). \quad (24)
\]

By eliminating \( W_{s}(w) \) from (19), using (23), and \( q_{L}(w) \) using (20),

\[
I_{sl}(w) = \frac{j\omega \xi^{*}(w)p(w)p' (w) \partial I_{sl}(w)}{\beta^2} \frac{\partial}{\partial w^2} \quad (25)
\]

where

\[
z(w) = z^i + j\omega \xi^{*}(w). \quad (26)
\]

Since \( p(w) \) and \( p'(w) \) are small,

\[
p(w)p'^{*} \ll 1; \quad (27)
\]

and since, in order of magnitude,

\[
\frac{\partial^2 I_{sl}(w)}{\partial w^2} = -\beta^2 I_{sl}(w); \quad (28)
\]

it follows subject to (27), that (25) with (18) reduces to,

\[
I_{sl}(w) = \frac{1}{z(w)} \left[ \frac{\partial V(w)}{\partial w} - \frac{k_{0}(w) + k_{o}r(w)}{k_{0}(w) + k_{o}r'(w)} \beta p(w) V(w) \right]. \quad (29)
\]

When there is no inductive coupling between load and line, \( k_{o}r(w) = 0 \) and the factor before \( \beta \) reduces to \( \phi(w) \). Outside the terminal zone (29) assumes the conventional form, namely,

\[
I_{sl}(w) = \frac{1}{z_0} \frac{\partial I^{'*}(w)}{\partial w} \quad z_0 = z^i + j\omega \xi^{*}. \quad (30)
\]

Differentiation of (23) with respect to \( w \), using (4) gives

\[
\frac{\partial^2 V(w)}{\partial w^2} + \beta^2 V_{L}(w) = \frac{\partial}{\partial w} \left[ z I_{sl}(w) + j\omega W_{sr}(w) \right]. \quad (31)
\]

This equation may be simplified by replacing the correction term \( W_{sr}(w) \) by its principal part obtained from (13). Thus, with (26) and (17a),

\[
z I_{sl}(w) + j\omega W_{sr}(w) = I_{sl}(w) [z^i + j\omega \xi^{*}(w)] = I_{sl}(w) [z(w) - j\omega \xi^{*}(w)]. \quad (32)
\]

With the leading term in (29), the derivative of (32) is given by

\[
\frac{\partial}{\partial w} \left[ z I_{sl}(w) + j\omega W_{sr}(w) \right] = \frac{\partial}{\partial w} \left[ z(w) - j\omega \xi^{*}(w) \frac{\partial V(w)}{\partial w} \right] \quad (33)
\]

Since the leading part of \( z(w) \) is \( j\omega \xi^{*}(w) \) and this varies slowly with \( w \) even in the terminal zone, the principal part in (33) is that given on the right. For terminations in which there is no \( z \) component of current near the line load junction, \( \xi^{*}(w) \) is zero, and (33) is exact. Substitution of (33) in (31), rearrangement of terms noting that \( V_L(w)/V(w) = k_0(w)/[k_0(w) + k_0r'(w)] \), and use of (17a), gives the following final equation for \( V(w) \):

\[
\frac{\partial^2 V(w)}{\partial w^2} - \gamma^2 V(w) = 0, \quad (34)
\]

where

\[
\gamma^2 (w) \equiv z(w) \gamma(w) = \left[ z^i + j\omega \xi^{*}(w) \right] \left[ g(w) + j\omega c(w) \right]. \quad (35)
\]

Since a small error in the small quantity \( z^i \) in the terminal zone is of no consequence,

\[
\gamma^2 (w) \equiv z^i \alpha_1(w) \phi_1(w), \quad (36a)
\]

where

\[
\gamma^2 (w) = \gamma^2 (w - z) = z_0 \gamma_0 = \left[ z^i + j\omega \xi^{*}(w) \right] \left[ g_0 + j\omega c_0 \right]. \quad (36b)
\]

The parameter \( \gamma \), which is independent of \( w \), is the conventional complex propagation constant.

The scalar potential difference at any cross section of the line is obtained by solving (34); the current is then given by (29).

**APPENDIX TERMINAL IMPEDANCE**

Transmission-line measurements usually depend on distribution curves or resonance curves obtained on parts of the line quite far from the terminations. The data so obtained are then interpreted using conventional formulas that are valid only outside the terminal zones. This procedure is correct only if sufficiently long sections of line are included as a part of the termination, so that \( z = s \) is not the actual end of the smooth line. If conventional formulas are assumed to apply to the actual end of the line, the impedance apparently terminating it includes the effect of errors made in using incorrect line parameters and formulas in the terminal zone. This apparent terminal impedance \( Z_{at} \) at \( z = s \) is not the ratio of the actual scalar potential difference to current. Since \( Z_{at} \) involves the properties of the transmission line, the same geometrical structure may have very different apparent terminal impedances when connected as load to different transmission lines. Merely varying the spacing of a line, for example, may change the apparent terminal impedance of a given load by a large amount.

For reasons similar to those which make it impossible to have the uniform properties of a long transmission line continue to its junction with an arbitrary impedor, it is also impossible to define for an arbitrary circuit element an impedance that is independent of the circuit to which it is connected. The degree of coupling of such an element to the adjacent parts of the circuit varies with the configuration of conductors; it may be large or small, but never zero. As explained in the literature,\(^1\) it is for-

\(^1\) See pages 419 and 461 of footnote reference 1.
mally possible to separate the coupling between two parts of a complete circuit into two self-impedances and a mutual impedance. Except when the distribution of current is greatly affected by the mutual term, the self-term of the load differs negligibly from the self-impedance when isolated with an equal potential difference maintained at its terminals by a fictitious "concentrated" source. This is true of the coupling between a transmission line and its load. Accordingly, the transmission-line may be analyzed as if it had a physically dimensionless load, and the load as if it were isolated and driven by a fictitious, dimensionless source that maintains the required potential difference $V(s) = \Phi_1(s) - \Phi_2(s)$ at its terminals (Fig. 4), provided account is taken of the closed form or by numerical methods for a particular termination.

If $Z_T$ is assumed connected in series with the load and $Y_T$ in parallel with it, $\gamma^2$ may be substituted for $\gamma(z)$ in (34), $z_0$ for $p(w)$, and zero for $p(w)$; also the factor before $\beta$ in (29) may be replaced by unity. It is now possible to obtain solutions in conventional form, namely,

$$I'(w) = I'(s)[\cosh \gamma w + Y_w Z_c \sinh \gamma w] \quad (40a)$$
$$I'(w) = I'(s) \frac{Z_c}{Z_c} [\sinh \gamma w + Y_w Z_c \cosh \gamma w], \quad (40b)$$

where $Z_c$ is the conventional characteristic impedance of the line and where $Y_w$ consists of the impedance

$$Z_s = V(s)/I(s) \quad (41)$$
in series with $Z_T$ and in parallel with $Y_T$, $L_T$ and $C_T$ are positive or negative depending on the termination. Thus, an approximately equivalent circuit of an actual two-wire line of length $s$ with a terminal zone (extending from $z = s - d$ to $z = s$ where the line joins the load) is a fictitious line of length $s$ without terminal zone, but with the network of lumped elements shown in Fig. 5 serving as connection between the line and the "isolated" impedance $Z_s$ of the load. Note that the impedances apparently terminating both actual and fictitious lines are $Z_{ro}$. Neglecting line losses, this is the impedance looking toward the load at a distance $\lambda/2$ from it. $Z_{ro}$ is a direct measurable quantity, $Z_s$ is not; $Z_{ro}$ varies with the type of line and the nature of the connection, $Z_s$ is by definition independent of the line.

**Conclusion**

A theory has been developed that is an essential supplement to conventional transmission-line analysis. It has been applied to determine the apparent impedance of an open end, of a bridged end, and of antennas connected (1) as end loads\(^4\) in the plane and perpendicular to the line. The integrals (38) and (39) may be evaluated either in closed form or by numerical methods for a particular termination.

If $Z_T$ is assumed connected in series with the load and $Y_T$ in parallel with it, $\gamma^2$ may be substituted for $\gamma(z)$ in (34), $z_0$ for $p(w)$, and zero for $p(w)$; also the factor before $\beta$ in (29) may be replaced by unity. It is now possible to obtain solutions in conventional form, namely,

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where $Z_c$ is the conventional characteristic impedance of the line and where $Y_w$ consists of the impedance

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On the Energy-Spectrum of An Almost Periodic Succession of Pulses*

G. G. MACFARLANE†, ASSOCIATE, IRE

Summary—An analysis is presented of the effect of (i) irregularities in the amplitude of pulses, and (ii) jitter in the repetition rate on the energy-spectrum of a succession of pulses.

In both cases the spectrum has two components: (a) a line spectrum, and (b) a continuous spectrum. In case (i), the envelope of both spectra is proportional to the envelope of the energy-spectrum of a single pulse and the lines in the line spectrum are spaced by the repetition frequency (see Figs. 1 and 2). In case (ii), the envelopes of the spectra are not the same as the envelope of a single pulse, and the lines are spaced by the mean repetition frequency (see Figs. 3 and 4).

I. INTRODUCTION

In the frequency analysis of a succession of pulses it is generally assumed that the spacing and amplitude of pulses are perfectly constant. This allows the spectrum to be calculated by straightforward Fourier analysis. However, in practice owing to causes which may be systematic or random, all pulses are not equally spaced and/or do not have exactly the same amplitude. This paper is devoted to the description of a method for calculating the energy-frequency spectrum of such an almost periodic succession of pulses in which the irregularity occurs in a random way.

The method is illustrated in two cases of special interest. In the first, the pulses all have the same shape and are regularly spaced, but they vary in amplitude in a random manner about a mean value. The output from a superregenerative receiver operated in the linear mode is like this. In the second, the pulses are identical in shape, but their spacing varies in a random manner about a mean spacing. Barkhausen noise in a magnetic amplifier is thought to be like this. The first case considered is PAM (pulse amplitude modulation) and the second is PPM (pulse position modulation), each using fluctuation noise as the modulation wave.

II. THE ENERGY SPECTRUM OF A REGULAR SUCCESSION OF PULSES ALL OF THE SAME SHAPE, BUT OF DIFFERENT AMPLITUDES

We shall first derive an expression for the spectral energy-density of 2N+1 pulses in terms of the amplitudes $a_n$ of the pulses.

Let the spectrum of a single pulse of unit amplitude occurring at zero time be $G(\omega)$. The spectrum of the same pulse occurring at time $t$ is

\[ G(\omega) \exp(i\omega t). \]

The spectrum of 2N+1 pulses spaced $T$ apart and with amplitudes $a_n$, as shown in Fig. 1, is the sum of the spectra of the individual pulses, and is therefore

\[ S(\omega) = G(\omega) \sum_{-N}^{N} a_n \exp(i\omega T). \]

The spectral energy density, which is the mean value of the square of the modulus of $S(\omega)$ per unit time, is

\[ R_N(\omega) = \frac{1}{(2N+1)T} \left| \frac{1}{T} \sum_{-N}^{N} a_n \exp(i\omega T) \right|^2. \]

\[ = \frac{1}{(2N+1)T} \left[ \sum_{-N}^{N} a_n \cos(n\omega T) \right]^2. \]

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The approximate integrals in (15) differ negligibly from the exact integrals if $|\beta b| \ll 1$.

The terms neglected on the right in (15) contribute to radiation from the line; (16) makes this negligible. Let

$$l'(w) = l_0'(w) + l_T'(w) = [k_0(w) + k_{0T}(w)]/2\pi$$

$$\omega y^{-1}(w) = j\omega y_{0}(w) + y_{0T}(w)$$

$$= [k_0(w) + k_{0T}(w)]/2\pi \xi$$

$$y(w) = g(w) + j\omega c(w)$$

$$p(w) = [k_1(w) + k_{1T}(w)]/[k_0(w) + k_{0T}(w)]$$

$$= [k_1(w) + k_{1T}(w)]/[k_0(w) + k_{0T}(w)]$$

$$p_0(w) = k_1(w)/k_0(w).$$

Note that

$$\beta^2 = \omega^2/\nu = -j\omega y_{0}(w) y(w) [k_0(w) + k_{0T}(w)]/k_0(w).$$

With (17) substituted in (13) and (14), the following expressions are valid at all points along the line:

$$W_s(w) = l'(w) / l_0'(w) = W_s(w) / W_{sL}(w):$$

$$\phi_1(w) = c(w)/c_0(w) = V_L(w)/V(w).$$

When $a_1(w) = 1$, there is no inductive coupling; when $\phi_1(w)$, there is no capacitive coupling between line and termination.

At points outside the terminal zones, where conventional line theory is adequate, these relations reduce to

$$W_s(w) = l_0'(w); \quad V(w) = j\omega q_L(w)/y_0$$

$$l_0' = k_0(w \to \infty)/2\pi = [\ln b/a]/2\pi$$

$$y_0 = g_0 + j\omega c_0 = j\omega 2\pi/k_0(w \to \infty)$$

$$= j\omega \pi/[\ln b/a].$$

The functions $k_0(w)$, $k_1(w)/\beta_0$, and $p_0(w)/\beta_0$, which characterize the transmission-line-end effect for a line in air where $\beta = \beta_0$ (but do not include coupling to the termination) are shown in Fig. 2. The corresponding values of $l_0'(w)$, $c_0(w)$, and $R_c(w) = [l_0'(w)/c_0(w)]^{1/2}$ are in Fig. 3.

Fig. 2—The functions $k_0(w)$, $k_1(w)/\beta_0$, at the load end of a long line in air which is not coupled to the load.

Fig. 3—The external inductance and capacitance per unit length and characteristic resistance at the load-end of a line in air with no coupling to the load, $l_0'(w) = 2 \times 10^{-7}$ henry/m; $c_0(w) = 2 \times 8.85 \times 10^{-12}$ farads/m; $R_c(w) = \sqrt{l_0'(w)/c_0(w)}$.

**Generalized Differential Equations for Voltage and Current**

The generalized differential equations for current and voltage may be derived from the equations defining the scalar and vector potentials, namely, $E = -\nabla \Phi - j\omega A$; div $A + j(\beta^2/\omega)\Phi = 0$; and the relation, $E_{AI} = z_1 I'_{AI}$, where $z_1$ is the internal impedance per unit length of conductor 1 and $E_{AI}$ is the tangential electric field at its surface. Noting that $z_1' = z_1 = z^2/2$, where $z$

\[ See \ page \ 348 \ of \ footnote \ reference \ 1.\]
s the internal impedance per loop unit length of the line, and using (1), (3), and \( w = s - z \), the following equations are obtained:

\[
\frac{\partial I_{sL}(w)}{\partial w} = z I_{sL}(w) + j \omega W_s(w) \tag{23}
\]

\[
\frac{\partial W_{sL}(w)}{\partial w} = \frac{j \beta^2}{\omega} V_L(w). \tag{24}
\]

By eliminating \( W_s(w) \) from (19), using (23), and \( q_L(w) \) using (20),

\[
z(w) I_{sL}(w) - \frac{j \omega l_s(w)p(w)p'(w)}{\beta} \frac{\partial I_{sL}(w)}{\partial w} = \frac{\partial V(w)}{\partial w} - \frac{j \omega l_s(w)y'(w)p(w)}{\beta} V(w) = 0, \tag{25}
\]

where

\[ z(w) = z^i + j \omega l_s(w). \tag{26} \]

Since \( p(w) \) and \( p'(w) \) are small,

\[ p(w)p'^2(w) \ll 1; \tag{27} \]

and since, in order of magnitude,

\[
\beta^2 I_{sL}(w)/\partial w^2 = -\beta^2 I_{sL}(w); \tag{28}
\]

it follows subject to (27), that (25) with (18) reduces to,

\[
l_{sL}(w) = \frac{1}{z(w)} \left[ \frac{\partial V(w)}{\partial w} + \frac{k_s(w) + k_{cr}(w)}{k_s(w) + k_{cr}(w)} \beta p(w) V(w) \right]. \tag{29}
\]

When there is no inductive coupling between load and line, \( k_{cr}(w) = 0 \) and the factor before \( \beta \) reduces to \( \phi_l(w) \). Outside the terminal zone (29) assumes the conventional form, namely,

\[
I_{sL}(w) = \frac{1}{z_0} \frac{\partial V(w)}{\partial w}; \quad z_0 = z^i + j \omega l_s. \tag{30}
\]

Differentiation of (23) with respect to \( w \), using (4) gives

\[
\frac{\partial^2 V(w)}{\partial w^2} + \beta^2 V_L(w) = \frac{\partial}{\partial w} \left[ z I_{sL}(w) + j \omega W_{sT}(w) \right]. \tag{31}
\]

This equation may be simplified by replacing the correction term \( W_{sT}(w) \) by its principal part obtained from (13). Thus, with (26) and (17a),

\[
z I_{sL}(w) + j \omega W_{sT}(w) = I_{sL}(w) z^i + j \omega l_s \phi_l(w) = I_{sL}(w) \left[ z(w) - j \omega l_s \phi_l(w) \right]. \tag{32}
\]

With the leading term in (29), the derivative of (32) is given by

\[
\frac{\partial}{\partial w} \left[ z I_{sL}(w) + j \omega W_{sT}(w) \right] = \frac{\partial}{\partial w} \left[ \frac{z(w) - j \omega l_s \phi_l(w)}{z(w)} \frac{\partial V(w)}{\partial w} \right] = \frac{z(w) - j \omega l_s \phi_l(w)}{z(w)} \frac{\partial^2 V(w)}{\partial w^2}. \tag{33}
\]

Since the leading part of \( z(w) \) is \( j \omega l_s \phi_l(w) \) and this varies slowly with \( w \) even in the terminal zone, the principal part in (33) is that given on the right. For terminations in which there is no \( z \) component of current near the line load junction, \( l_{sT}(w) \) is zero, and (33) is exact. Substitution of (33) in (31), rearrangement of terms noting that \( V_{sL}(w)/V_{sT}(w) = k_s(w)/[k_s(w) + k_{cr}(w)] \), and use of (17a), gives the following final equation for \( V(w) \):

\[
\frac{\partial^2 V(w)}{\partial w^2} - \gamma^2(w) V(w) = 0, \tag{34}
\]

where

\[
\gamma^2(w) = z(w) y(w) = |z^i + j \omega l_s \phi_l| |g(w) + j \omega c (w)|. \tag{35}
\]

Since a small error in the small quantity \( z^i \) in the terminal zone is of no consequence,

\[
\gamma^2(w) \approx \gamma^2 l_1(w) \phi_1(w). \tag{36a}
\]

where

\[
\gamma^2 \equiv \gamma^2(w \rightarrow x) = z_0 y_0 = (z^i + j \omega l_s \phi_l) (g_0 + j \omega c_0). \tag{36b}
\]

The parameter \( \gamma \), which is independent of \( w \), is the conventional complex propagation constant.

The scalar potential difference at any cross section of the line is obtained by solving (34); the current is then given by (29).

**Approximate Solutions of the Equations—Apparent Terminal Impedance**

Transmission-line measurements usually depend on distribution curves or resonance curves obtained on parts of the line quite far from the terminations. The data so obtained are then interpreted using conventional formulas that are valid only outside the terminal zones. This procedure is correct only if sufficiently long sections of line are included as a part of the termination, so that \( z = s \) is not the actual end of the smooth line. If conventional formulas are assumed to apply to the actual end of the line, the impedance *apparently* terminating it includes the effect of errors made in using incorrect line parameters and formulas in the terminal zone. This *apparent* terminal impedance \( Z_{as} \) at \( z = s \) is not the ratio of the actual scalar potential difference to current. Since \( Z_{as} \) involves the properties of the transmission line, the same geometrical structure may have very different apparent terminal impedances when connected as load to different transmission lines. Merely varying the spacing of a line, for example, may change the apparent terminal impedance of a given load by a large amount.

For reasons similar to those which make it impossible to have the uniform properties of a long transmission line continue to its junction with an arbitrary impedor, it is also impossible to define for an arbitrary circuit element an impedance that is independent of the circuit to which it is connected. The degree of coupling of such an element to the adjacent parts of the circuit varies with the configuration of conductors; it may be large or small, but never zero. As explained in the literature,1 it is for-

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1 See pages 419 and 461 of footnote reference 1.
nally possible to separate the coupling between two parts of a complete circuit into two self-impedances and a mutual impedance. Except when the distribution of current is greatly affected by the mutual term, the self-term of the load differs negligibly from the self-impedances when isolated with an equal potential difference maintained at its terminals by a fictitious "concentrated" source. This is true of the coupling between a transmission line and its load. Accordingly, the transmission-line may be analyzed as if it had a physically dimensionless load, and the load as if it were isolated and driven by a fictitious, dimensionless source that maintains the required potential difference \( V(s) = \Phi(s) - \Phi_s(s) \) at its terminals (Fig. 4), provided account is taken of the coupling between them. This may be done approximately by means of a suitable equivalent network that represents the coupling as if lumped at the junction instead of distributed over short distances near it. By concentrating end effects and coupling effects in the network of lumped elements, the terminal zone may be replaced by a fictitious section of conventional line that has as its termination the apparent terminal impedance \( Z_w \). This is composed of the fictitious impedance \( Z_s \) of the load (treated as if isolated and given by \( Z_s = V(s)/I(s) \) as shown in Fig. 4) in conjunction with the appropriate terminal-zone network that includes a series element \( Z_T \) and a shunt element \( Y_T \) defined as follows:

\[
Z_T = \int_0^d [z(w) - z_0]dw = jw \int_0^d [l'(w) - l_{e'}]dw
\]

\[
= jw \int_0^d [l'(w)a_1(w) - l_{e'}]dw = jwI_T \quad (38)
\]

\[
Y_T = \int_0^d [y(w) - y_0]dw = jw \int_0^d [c(w) - c_0]dw
\]

\[
= jw \int_0^d [c_0(w)\phi_1(w) - c_0]dw = jwC_T. \quad (39)
\]

The integrals (38) and (39) may be evaluated either in closed form or by numerical methods for a particular termination.

If \( Z_T \) is assumed connected in series with the load and \( Y_T \) in parallel with it, \( \gamma \) may be substituted for \( \gamma^2(w) \) in (34), \( z_0 \) for \( z(w) \), and zero for \( \rho(w) \); also the factor before \( \beta \) in (29) may be replaced by unity. It is now possible to obtain solutions in conventional form, namely,

\[
V(w) = V(s)[\cosh \gamma w + Y_sZ_c \sinh \gamma w] \quad (40a)
\]

\[
I(w) = \frac{V(s)}{Z_c} [\sinh \gamma w + Y_sZ_c \cosh \gamma w], \quad (40b)
\]

where \( Z_c \) is the conventional characteristic impedance of the line and where \( Y_s \) consists of the impedance

\[
Z = V(s)/I(s) \quad (41)
\]

in series with \( Z_T \) and in parallel with \( Y_T \). \( L_T \) and \( C_T \) are positive or negative depending on the termination. Thus, an approximately equivalent circuit of an actual two-wire line of length \( s \) with a terminal zone (extending from \( z = s - d \) to \( z = s \) where the line joins the load) is a fictitious line of length \( s \) without terminal zone, but with the network of lumped elements shown in Fig. 5 serving as connection between the line and the "isolated" impedance \( Z_s \) of the load. Note that the impedances apparently terminating both actual and fictitious lines are \( Z_{se} \). Neglecting line losses, this is the impedance looking toward the load at a distance \( \lambda/2 \) from it. \( Z_{se} \) is a direct measurable quantity, \( Z_s \) is not; \( Z_{se} \) varies with the type of line and the nature of the connection, \( Z_s \) is by definition independent of the line.

**Conclusion**

A theory has been developed that is an essential supplement to conventional transmission-line analysis. It has been applied to determine the apparent impedance of an open end, of a bridged end, and of antennas connected (1) as end loads in the plane and perpendicular.

On the Energy-Spectrum of An Almost Periodic Succession of Pulses

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Summary—An analysis is presented of the effect of (i) irregularities in the amplitude of pulses, and (ii) jitter in the repetition rate on the energy-spectrum of a succession of pulses.

II. THE ENERGY SPECTRUM OF A REGULAR SUCCESSION OF PULSES ALL OF THE SAME SHAPE, BUT OF DIFFERENT AMPLITUDES

We shall firstly derive an expression for the spectral energy-density of $2N+1$ pulses in terms of the amplitudes $a_n$ of the pulses.

Let the spectrum of a single pulse of unit amplitude occurring at zero time be $G(\omega)$. The spectrum of the same pulse occurring at time $t$ is

$$G(\omega) \exp(i\omega t).$$

The spectrum of $2N+1$ pulses spaced $T$ apart and with amplitudes $a_n$, as shown in Fig. 1, is the sum of the spectra of the individual pulses, and is therefore

$$S(\omega) = G(\omega) \sum_{-N}^{N} a_n \exp(i\omega T).$$

The spectral energy density, which is the mean value of the square of the modulus of $S(\omega)$ per unit time, is

$$R_N(\omega) = \frac{|S(\omega)|^2}{(2N+1)T}$$

$$= \left[ \frac{|G(\omega)|^2}{T} \left\{ \frac{1}{2N+1} \left[ \sum_{-N}^{N} a_n \exp(i\omega T)^2 \right] \right\} \right]$$
Let us now assume that the amplitudes \( a_n \) are randomly distributed, and that the probability of the amplitude \( a_n \), having a value between \( x \) and \( x+dx \) is \( q_n(x)dx \). The function \( q_n(x) \) is called the probability function. In order to evaluate the spectral energy-density, equation (4), we require to know the mean-square values of the sums

\[
\sum_{-N}^{N} a_n \cos (n \omega T), \quad \sum_{-N}^{N} a_n \sin (n \omega T). \tag{5}
\]

They are readily obtained when we know the probability functions \( P_X \) of these sums. We can find the functions \( P_X \) in the following way.

Suppose the probability function of the random variable \( x \) is \( q(x) \), then the probability function \( p_n(y) \) of an arbitrary function \( y = f_n(x) \) is obtained from \( q(x) \) by substituting

\[
x = F_n(y),
\]

where \( F_n(y) \) is the inverse of \( f_n(x) \).

Thus

\[
y = f_n(x). \tag{6}
\]

This gives

\[
p_n(y) = q(y) \left| \frac{dF_n(y)}{dy} \right|.
\]

Furthermore, there is a theorem in the theory of probability\(^1\) which states that the mean and the mean-square deviation of the sum of any number of probability distributions is the sum of the means and the mean-square deviations respectively of the component distributions. By the sum of the probability distributions, \( p_n(x) \), is meant the probability distribution of the sum of a number of functions \( f_n(x) \) of the random variable \( x \). Thus the sum of two probability distributions \( p_1(x) \) and \( p_2(x) \) is not \( p_1(x) + p_2(x) \), but

\[
\int_{-\infty}^{\infty} p_1(x-y)p_2(y)dy. \tag{7}
\]

This theorem is directly applicable to our case, for we seek to find the mean-square value \( x_N^2 \) of the sum \( \sum_{-N}^{N} f_n(x) \) of \( 2N+1 \) functions of the random variable \( x \), and this is equal to the sum of the square of the mean, \( (\bar{x})^2 \), and the mean-square deviation, \( \rho_N^2 \), of \( \sum_{-N}^{N} f_n(x) \). Or

\[
\bar{x}_N^2 = \rho_N^2 + (\bar{x}_N)^2; \tag{10}
\]

while the theorem above states that

\[
\bar{x}_N = \sum_{-N}^{N} X_n \tag{11}
\]

and

\[
\rho_N^2 = \sum_{-N}^{N} \sigma^2_{n}. \tag{12}
\]

where \( X_n \) is the mean and \( \sigma^2_{n} \) the mean-square deviation of the distribution \( p_n(x) \) of the function \( f_n(x) \). That is,

\[
X_n = \int_{-\infty}^{\infty} x p_n(x)dx = \int_{-\infty}^{\infty} f_n(x)q(x)dx \tag{13}
\]

and

\[
\sigma^2_{n} = \int_{-\infty}^{\infty} [f_n(x) - X_n]^2 q(x)dx. \tag{14}
\]

Herein use is made of (8) to express \( X_n \) and \( \sigma^2_{n} \) in terms of the probability function \( q(x) \) of the random variable \( x \).

Having established these general formulas, let us return to the discussion of (3) for the spectral energy-density. Consider the series

\[
\sum_{-N}^{N} f_n(x) = \sum_{-N}^{N} a_n \sin (n \omega T). \tag{15}
\]

Here

\[
f_n(x) = x \sin (n \omega T).
\]

Therefore, from (13) and (14),

\[
X_n = \sin (n \omega T) \int_{-\infty}^{\infty} xq(x)dx = \sin (n \omega T) \bar{x}, \text{ say.} \tag{16}
\]

and

\[
\sigma^2_{n} = \sin^2 (n \omega T) \int_{-\infty}^{\infty} (x - \bar{x})^2 q(x)dx = \sin^2 (n \omega T) (1 - \bar{x}^2), \text{ say.} \tag{17}
\]

Substitute \( X_n \) and \( \sigma^2_{n} \) from (16) and (17) into (10) to (12). This gives the mean-square value of the sum (15)

\[
\left\{ \left[ \sum_{-N}^{N} a_n \sin (n \omega T) \right]^2 \right\}_{\lambda_N} = \left( 1 - \bar{x}^2 \right) \sum_{-N}^{N} \sin^2 (n \omega T) + (\bar{x})^2 \left[ \sum_{-N}^{N} \sin (n \omega T) \right]^2. \tag{18}
\]

Similarly,

\[
\left\{ \left[ \sum_{-N}^{N} a_n \cos (n \omega T) \right]^2 \right\}_{\lambda_N}.
\]

---

\begin{align}
&= (1 - \overline{A})^2 \sum_{n}^{N} \cos^2 (n\omega T) \\
&\quad + (\overline{A})^2 \left[ \sum_{n}^{N} \cos (n\omega T) \right]^2.
\end{align}

Therefore,
\[ \int \frac{1}{2N + 1} \left\{ \sum_{n}^{N} a_n \exp (in\omega T) \right\}^2 \, d\omega = (1 - \overline{A})^2 + \frac{(\overline{A})^2}{2N + 1} \left[ \sum_{n}^{N} \sin (n\omega T) \right]^2 \\
&\quad + \frac{(\overline{A})^2}{2N + 1} \left[ \sum_{n}^{N} \cos (n\omega T) \right]^2. \tag{20}
\]

Now let the number of pulses increase indefinitely, so that we can let \( N \to \infty \).

For all values of \( \omega T \) and
\[ \lim_{N \to \infty} \frac{1}{2N + 1} \left[ \sum_{n}^{N} \cos (n\omega T) \right]^2 = 0 \]
for the probability function \( q(x) \) is therefore
\[ R(\omega) = \lim_{N \to \infty} R_x(\omega) \]
\[ = \frac{(1 - \overline{A})^2}{2N + 1} \left[ \sum_{n}^{N} \cos (n\omega T) \right]^2 \left[ \frac{2\pi}{T} \delta \left( \omega - \frac{2m\pi}{T} \right) \right] \left[ \frac{|G(\omega)|^2}{T} \right]. \tag{21} \]

where \( \delta(x) \) is the unit impulse function of Dirac which is infinite at \( x = 0 \) but zero everywhere else, although the integral of \( \delta(x) \) over any range including the point \( x = 0 \) is unity. The spectral energy-density of an infinite succession of pulses of equal spacing, \( T \), and similar shape but with amplitudes which are randomly distributed according to the probability function \( q(x) \) is therefore
\[ R(\omega) = \lim_{N \to \infty} R_x(\omega) \]
\[ = \frac{(1 - \overline{A})^2}{2N + 1} \left[ \sum_{n}^{N} \cos (n\omega T) \right]^2 \left[ \frac{2\pi}{T} \delta \left( \omega - \frac{2m\pi}{T} \right) \right] \left[ \frac{|G(\omega)|^2}{T} \right]. \tag{21} \]

where
\[ \overline{A} = \int_{-\infty}^{+\infty} xq(x) \, dx \tag{22} \]
and
\[ (\overline{A} - \overline{\overline{A}})^2 = \int_{-\infty}^{+\infty} (x - \overline{A})^2 q(x) \, dx. \tag{23} \]

The energy spectrum consists of two distinct component spectra (i) a continuous distribution of density \((A - \overline{A})^2 |G(\omega)|^2 / T^2\) with intensity proportional to the spectral density of a single pulse (Fig. 2a), and (ii) a line spectrum with separation of lines in frequency by \( T \), the recurrence frequency of the pulses. The power in the line of frequency \( m/T \) is
\[ \frac{2\pi}{T} \left( \frac{A^2}{T} \right) - \left( \frac{2m\pi}{T} \right) \left( |G\left( \frac{2m\pi}{T} \right)|^2 \right). \tag{24} \]

If the power in the lines is plotted against frequency, the power spectrum of Fig. 2(b) is obtained. The envelope of the lines is proportional to the spectral density of a single pulse.
of the random variable \( x \). For the first of these two, we have

\[
\phi_n(x) = \sin(\omega nT + x).
\]

From (13) and (14) the mean value and the square of the standard deviation of the probability distribution of \( \sin(\omega(nT + x)) \) are, respectively,

\[
\begin{align*}
X_n &= \frac{1}{\sigma\sqrt{2\pi}} \int_{-\infty}^{\infty} \sin(\omega(nT + x)) \exp\left(-\frac{x^2}{2\sigma^2}\right) dx \\
&= \exp\left(-\frac{1}{2}\sigma^4\omega^2 \sin(\omega T)\right) \\
\sigma_n^2 &= \frac{1}{\sigma\sqrt{2\pi}} \int_{-\infty}^{\infty} \sin^2(\omega(nT + x)) \exp\left(-\frac{x^2}{2\sigma^2}\right) dx - X_n^2 \\
&= \frac{1}{2} \left[ 1 - \exp\left(-2\sigma^4\omega^2\right) \right] \\
&\quad - \left[ 1 - \exp\left(-\sigma^4\omega^2\right) \right] \exp\left(-\sigma^4\omega^2\right) \sin^2(\omega T).
\end{align*}
\]

Therefore, from (10) and (11) the mean-square value of \( \frac{1}{N} \sum_{-N}^{N} \phi_n(x) \) is

\[
\begin{align*}
&\left\{ \left( \sum_{-N}^{N} \sin(\omega(nT + a_n)) \right)^2 \right\}_{\text{Av.}} \\
&= \frac{2N+1}{2} \left[ 1 - \exp\left(-2\sigma^4\omega^2\right) \right] \\
&\quad + \exp\left(-\sigma^4\omega^2\right) \left[ \sum_{-N}^{N} \sin(\omega T) \right]^2 \\
&\quad - \exp\left(-\sigma^4\omega^2\right) \left[ 1 - \exp\left(-\sigma^4\omega^2\right) \right] \sum_{-N}^{N} \sin^2(\omega T).
\end{align*}
\]

Similarly

\[
\begin{align*}
&\left\{ \left( \sum_{-N}^{N} \cos(\omega(nT + a_n)) \right)^2 \right\}_{\text{Av.}} \\
&= \frac{2N+1}{2} \left[ 1 - \exp\left(-2\sigma^4\omega^2\right) \right] \\
&\quad + \exp\left(-\sigma^4\omega^2\right) \left[ \sum_{-N}^{N} \cos(\omega T) \right]^2 \\
&\quad - \exp\left(-\sigma^4\omega^2\right) \left[ 1 - \exp\left(-\sigma^4\omega^2\right) \right] \sum_{-N}^{N} \cos^2(\omega T).
\end{align*}
\]

Therefore

\[
\begin{align*}
&\left\{ \frac{1}{2N+1} \left[ \sum_{-N}^{N} \exp \left\{ i\omega(nT + a_n) \right\} \right]^2 \right\}_{\text{Av.}} \\
&= 1 - \exp\left(-\sigma^4\omega^2\right) + \frac{\exp\left(-\sigma^4\omega^2\right)}{2N+1} \left[ \sum_{-N}^{N} \sin(\omega T) \right]^2 \\
&\quad + \left[ \sum_{-N}^{N} \cos(\omega T) \right]^2.
\end{align*}
\]

The energy spectrum is now obtained as before by letting \( N \to \infty \). The result is
Here again the energy spectrum consists of two distinct component spectra: a continuous spectrum and a line spectrum. If the pulses recurred with perfectly regular intervals \( \sigma \) would be zero, and the spectrum would be solely a line spectrum. The average power in the lines would then be

\[
R(\omega) = \left\{1 - \exp \left(-\sigma^2 \omega^2\right) + \exp \left(-\sigma^2 \omega^2\right) \frac{2\pi}{T} \delta \left(\omega - \frac{2m\pi}{T}\right) \right\} |G(\omega)|^2. \tag{33}
\]

When, however, the repetition rate is not steady, the average power in the lines is reduced to

\[
\frac{1}{T} \sum_{m=-\infty}^{\infty} G \left(\frac{2m\pi}{T}\right) |G(\omega)|^2 \exp \left(-\frac{4m^2 \sigma^2}{T^2}\right) \frac{2\pi}{T} \approx \frac{1}{T} \int_{-\infty}^{+\infty} \exp \left(-\sigma^2 \omega^2\right) |G(\omega)|^2 d\omega. \tag{34}
\]

and energy is distributed continuously over the frequency band of a single pulse. The power in this continuous spectrum is noise power. The ratio of signal-to-noise powers is approximately:

\[
\frac{\int_{-\infty}^{+\infty} \exp \left(-\sigma^2 \omega^2\right) |G(\omega)|^2 d\omega}{\int_{-\infty}^{+\infty} |1 - \exp \left(-\sigma^2 \omega^2\right)| |G(\omega)|^2 d\omega}
\]

There is a difference in the forms of the energy spectra given by (21) and (33), which is interesting. In (21), component and the power spectrum of the lines is the same as the energy-density spectrum of a single pulse. In (33) however, this is not the case, for the factors multiplying \(|G(\omega)|^2\) depend on \(\omega\). Thus, the form of the envelope is altered for both signal line-spectrum and noise continuous-spectrum. The form of the energy-density spectrum of the noise is approximately \(\omega^2\) times that of a single pulse. This is illustrated in Figs. 4(a) and 4(b).

IV. THE SPECTRUM OF REGULAR PULSES OF UNEQUAL LENGTH

The method of finding the energy-frequency spectrum of a temporal variation used in the two problems discussed in parts II and III, is applicable to many more cases. The above examples have been chosen for their practical importance. A further example would be the output from a superregenerative receiver operated in logarithmic mode. Successive pulses are then unequal in duration due to the noise in the input circuit.

If the spacing of pulses is \(T\) and the spectrum of a pulse occurring at zero time is \(G(\omega, x)\) where \(x\) is the random variable that defines the pulse width, then the spectral energy-density is

\[
R(\omega) = \frac{1}{T} \left\{(\bar{A} - A)^2 + \frac{2\pi}{T} \delta \left(\omega - \frac{2m\pi}{T}\right)(\bar{A})^2\right\}.
\]

In this equation

\[
\bar{A} = \int_{-\infty}^{+\infty} G(\omega, x)q(x)dx,
\]

\[
(\bar{A} - A)^2 = \int_{-\infty}^{+\infty} G^2(\omega, x)q(x)dx - (\bar{A})^2,
\]

and \(q(x)\) is the probability function.

If the pulses are all rectangular of amplitude \(1/\tau\) and of mean width \(\tau\), and if the distribution function is Gaussian about zero as mean with standard deviation \(\sigma\)

\[
\bar{A} = \frac{1}{\omega} \int_{-\omega/2}^{\omega/2} \exp \left(i\tau y - 2\sigma^2 y^2\right)dy
\]

and

\[
(\bar{A} - \bar{A})^2 = \frac{1}{\omega^2} \int_{-\omega/2}^{\omega/2} dy \int_{-\omega/2}^{\omega/2} \exp \left(i\tau x - 2\sigma^2 x^2\right)dx - (\bar{A})^2.
\]

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Problems Related to Measuring the Field Strength of High-Frequency Electromagnetic Fields*

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Summary—Discussion and comparison of the analysis of two cases of an electron of uniform velocity acted on by a uniform magnetic field parallel to the electron direction, and a high-frequency electromagnetic field with the electric component transverse to the beam, shows that, for both a plane parallel-plate field and a rectangular cavity field, the field amplitude may be expressed quite simply in terms of easily measurable parameters. The simple relationship in question requires that certain velocity conditions on the electron velocity be observed, and that the cyclotron frequency \(|e|H/mc| = \text{frequency of the alternating electromagnetic field.}

Introduction

In a previous paper, the problem of an electron beam directed parallel to a uniform magnetic field and subjected to a transverse high-frequency electromagnetic field was discussed for a case where the alternating electromagnetic field was produced between two plane parallel plates, while the electron beam was directed along an axis midway between the plates.

The purpose of this paper is to present a more complete treatment of the parallel-plate case, and the extension to the case of a rectangular cavity. In the discussion presented here there appear certain velocity conditions on the electron beam which are necessary if one wants to have the solutions in the simplest and most convenient form. The conditions mentioned did not appear in the earlier discussion because the analysis there was not carried out in sufficient detail.

Plane Parallel-Plate Case

The plane parallel-plate situation is shown in Fig. 1. The electron beam of uniform velocity \(v_s\) is introduced at 0 in Fig. 1, with a uniform magnetic field \(H_z\) as indicated. The electric component of the rapidly alternating electromagnetic field is \(E_z\), and at any time \(t\), \(E_z\) is considered to be constant in magnitude over the region between the plates. The fringe field effect was discussed in the previous paper. The quantities \(H_z\), \(v_s\), and \(E_z\) determine the motion of an electron.

The equations of motion and the differential equations to be solved are:

\[
\begin{align*}
mx &= -\left[\frac{e}{c} \right] E_z - \frac{e}{c} yH_z, \\
m\frac{dy}{dt} &= \frac{e}{c} xH_z, \\
mz &= 0
\end{align*}
\]

where \(t_0\) is the time of entrance of the electron into the field region. Using \(E_z = E_{0z} \sin \omega t\), one has:

\[
\begin{align*}
x + \omega^2 t = K_2 \sin \omega t \\
y + \omega^2 t = \frac{\omega}{\omega} (\cos \omega t - \cos \omega t_0)
\end{align*}
\]

where

\[
\omega = \frac{|e| H_z}{m c} \quad \text{and} \quad K_2 = \frac{|e| E_{0z}}{m}
\]

the complete solution for \(x\) is:

\[
x = A x_1(t) + B x_2(t) + K_2 \int_{t-t_0}^{t} x_1(s) x_2'(s) \sin \omega s ds
\]

where

\[
x_1(t) = e^{i\omega t}, \quad x_2(t) = e^{-i\omega t}.
\]

It turns out that only the particular solution, given by the integral, contributes to our solution. The particular solution in \(x\) is:

\[
x = - \frac{K_2}{\omega_0} \left[ \frac{\sin (\omega_0 - \omega) t - \sin (\omega_0 - \omega) t_0}{2(\omega - \omega_0)} - \frac{\sin (\omega_0 + \omega) t - \sin (\omega_0 + \omega) t_0}{2(\omega + \omega_0)} \right] \cos \omega t
\]

\[
+ K_2 \left[ \frac{\cos (\omega_0 - \omega) t - \cos (\omega_0 - \omega) t_0}{2(\omega - \omega_0)} - \frac{\cos (\omega_0 + \omega) t - \cos (\omega_0 + \omega) t_0}{2(\omega + \omega_0)} \right] \sin \omega t
\]

When \(\omega = \omega_0\) the expression becomes:

\[
x = \frac{K_2}{2\omega_0} (t - t_0) \cos \omega t
\]

\[
+ \frac{K_2}{2\omega_0} \sin \omega t - \sin \omega t_0 \cos \omega (t - t_0)
\]

the corresponding solution in \(y\) is:

\[
y = \frac{K_2}{2\omega_0} (t - t_0) \sin \omega t
\]

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Let \((t-t_0)\) be the transit time \(r\) for the electron in the region between the plates. The relations for \(x\) and \(y\) become:

\[
x = -\frac{K_2}{2\omega_e} r \cos \omega_e(t_0 + r)
+ \frac{K_2}{2\omega_e^2} \left[ \sin \omega_e(t_0 + r) - \sin \omega_e t_0 \cos \omega_e r \right]
\]

\[
y = -\frac{K_2}{2\omega_e} r \sin \omega_e(t_0 + r)
+ \frac{K_2}{2\omega_e^2} \left[ 2 \cos \omega_e t_0 (1 - \cos \omega_e r) + \sin \omega_e t_0 \sin \omega_e r \right].
\]

If we regard \(\omega_e\) and \(r\) as fixed quantities, \(x\) and \(y\) are functions of the entrance time \(t_0\). In other words, the distance \(r\) of an electron from the \(z\) axis in a plane \(z=\text{constant}\) at the exit end of the plates is a function only of \(t_0\). It is apparent on looking at the component expressions for \(x\) and \(y\) that this expression for \(r\) is in general complicated. If, however, we impose the condition:

\[\omega_e r = n\pi, \quad n \text{ even}\]

the expressions for \(x\) and \(y\) become simply:

\[x = -\frac{K_2}{2\omega_e} r \cos \omega_e(t_0 + r)\]

\[y = -\frac{K_2}{2\omega_e} r \sin \omega_e(t_0 + r);\]

hence

\[
r = \frac{K_2}{2\omega_e} r
\]

since

\[
\tau = \frac{z_0}{v_e} \quad \text{and} \quad K_2 = -\frac{|e|}{m} E_{0z}
\]

\[
|E_{0z}| = \left( \frac{2m r v_e}{e z_0} \right) \left( \frac{2 r v_e H_s}{z_0 c} \right) \text{esu/cm.}
\]

\(z_0\) is the length of the plates.

It has been pointed out in (1) that neglecting the fringe field can introduce considerable error if the above expression is used to relate \(E_{0z}\) to \(r\), depending on the geometry and the frequency.

### Cavity Case

The use of a rectangular cavity (Fig. 2) in place of the parallel plates avoids the need of considering the fringe field question. The equations of motion for an electron are as before except that the field \(E_z\) now has a space dependence in the variables \(y\) and \(z\).

\[
m\ddot{x} = -\left| e \right| \left( E_{0z} + \frac{1}{c} \gamma H_s \right)
\]

\[
m\ddot{y} = -\left| e \right| \frac{\gamma}{c} H_s
\]

at \(t = t_0\)

\[
E_z = E_{0z} \cos \frac{\pi}{z_0} y \sin \frac{\pi}{z_0} z \sin \omega t
\]

(for the lowest mode in a rectangular cavity).

Now the quantity \(\cos \pi/2y \geq 1\) in the region of the electron path and is a slowly varying function as long as the beam deflection \(r\) remains small in comparison with \(y_s\).

We write \(z = v_s(t-t_0)\) where \(t_0\) is the time of entrance of the electron into the cavity at \(x = 0, y = 0, z = 0\). Now with \(E_z = E_{0z}\) \(\sin \pi/z_0 v_s(t-t_0)\) \(\sin \omega t\) the differential equations in \(x\) and \(y\) are:

\[
\dot{x} + \omega_e^2 x = K_2 \sin b(t-t_0) \sin \omega t
\]

\[
\dot{y} + \omega_e^2 y = \omega_e K_2 \sin b(t-t_0) \sin \omega t
\]

The particular solutions are given by:

\[
x = -\frac{K_2}{\omega_e} \int_{t=t_0}^{t=t_0+1} \sin \omega_s(s-t) \sin b(s-t_0) \sin \omega ds ds
\]

\[
y = \frac{K_2}{\omega_e} \int_{t=t_0}^{t=t_0+1} (1 - \cos \omega_s(s-t)) \sin b(s-t_0) \sin \omega ds ds
\]

and from the first of the two integrals one obtains:

\[
\begin{align*}
x &= \frac{K_2}{4\omega_e} \cos \omega_e t \cos b t_0 \cos (-\omega_e + \omega + b) t - \cos (-\omega_e + \omega + b) t_0
\cos (\omega_e + \omega + b) t - \cos (\omega_e + \omega + b) t_0
\end{align*}
\]

\[
+ \frac{K_2}{4\omega_e} \sin \omega_e t \cos b t_0 \sin (-\omega_e + \omega + b) t - \sin (-\omega_e + \omega + b) t_0
\sin (\omega_e + \omega + b) t - \sin (\omega_e + \omega + b) t_0
\end{align*}
\]
There are four values of \( \omega \), \((\omega = \omega_c \pm b\) and \(\omega = - \omega_c \pm b\)) for which various terms in \( x \) become indeterminate. In the parallel-plate case the corresponding situation arose only for \( \omega = \pm \omega_c \). Consider now one of the four possible cases for which \( \omega = \omega_c - b \); when the indeterminate terms are subjected to l'Hospital's rule one obtains:

\[
x = \frac{K_z}{4\omega_c} \left\{ - (t-t_0) \sin (\omega_c t-bt_0) + \frac{1}{(\omega_c-b)} \sin (\omega_c-b) t \sin b(t-t_0) - \sin (\omega_c-b) t \sin \omega_t(t-t_0) \right\}.
\]

The corresponding expression for \( y \) when \( \omega = \omega_c - b \) is similar to that for \( x \), but contains additional terms:

\[
y = \frac{K_z}{4\omega_c} \left\{ (t-t_0) \cos (\omega_c t-bt_0) - \frac{1}{(\omega_c-b)} \sin (\omega_c-b) t \cos b(t-t_0) - \cos \omega_t(t-t_0) \sin (\omega_c-b) t_0 \right\}.
\]

Hence

\[ b = \frac{\pi}{z_0}, \]

therefore, when \( \tau \) is the transit time \((t-t_0)\) of the electron through the cavity in the \( z \) direction, \( z = z_0 \) in the above expression, and

\[ b\tau = \frac{\pi}{z_0} (t-t_0) = \frac{\pi}{z_0} z; \]

In addition, we may notice that use of the condition

\[ (\omega_c - b) \tau = k\pi \quad k \text{ odd} \]

will greatly simplify the expressions for \( x \) and \( y \). Moreover, \( b\tau = \pi \), making the above condition:

\[ \omega_c \tau = (k+1)\pi \quad k \text{ odd} \]

or

\[ \omega_c \tau = n\pi \quad n \text{ even}, \]
Detection at Radio Frequencies by Superconductivity

J. V. LEBACQZ†, MEMBER, IRE, C. W. CLARK‡, M. C. WILLIAMS‡, AND D. H. ANDREWS†

Summary—Detection of radio-frequency currents by superconducting columbium nitride has been studied as a function of rf current, bias current, and temperature. The existence of nonlinear resistance in the zone between super and normal conduction has been observed, with specially high values of $dR/dI$ for currents less than one ma; this is believed to be due to the effect of the magnetic field of the current superconductivity (Silasbee hypothesis). The observed values of $dR/dI$ account in a rough quantitative way for the observed rectified potentials and their variation with rf current, bias current, and temperature at 1 Mc. Increasing rectification at higher frequencies is observed, but not yet explainable.

INTRODUCTION

In one of the earliest series of experiments ever made at very low temperatures, the Dutch physicist, H. Kamerlingh Onnes, discovered that many metals, when cooled to within a few degrees of absolute zero, suddenly lose all traces of electrical resistance and become perfect or “super” conductors. For any particular metal, the temperature interval in which this state of superconductivity is entered as the metal is cooled, is called the transition range. It is usually a zone a few hundredths of a degree wide, the midpoint of which is called the transition temperature.

This discovery was soon followed by another, that although resistance had disappeared completely below the transition point, it could be restored by passing sufficient current through the metal, the amount of this restoring current increasing as the temperature was lowered further below the transition range. The resistance was also found to be restored by the presence of a sufficiently intense magnetic field around the metal.

In 1917, Silsbee pointed out that the strength of the restoring current reported in a number of cases was just sufficient to produce at the surface of the metal a magnetic field equivalent to that which had been found just sufficient to restore resistance when externally applied. It was thus apparent that one had at these temperatures a resistance $R$ dependent on current $I$, providing a possible means of rectification of alternating currents and detection of radio signals. But the values of $dR/dI$ found were so small that this possibility was not investigated further for many years.

† Johns Hopkins University, Baltimore, Md.
‡ U. S. Army Ballistic Research Laboratory, Aberdeen, Md.
§ H. Kamerlingh Onnes, Leiden Comm., No. 122b, 1911.
Recently in studying the use of a superconducting alloy, columbium nitride (CnN) as a bolometer for detecting very weak infrared signals, it was found that, at certain values of the bias current, there was an abnormal amount of extraneous noise. This turned out to be due to pick up from a local commercial broadcasting station. It appears that with very small bias currents in this alloy at special regions of temperature, \( dR/dI \) attains values many times larger than in metallic elements and that considerable rectification therefore takes place, due to nonlinear resistance. Preliminary observations on this phenomenon have been reported recently, in addition to discussions of the use of CnN as an infrared bolometer, so that we shall refer to the earlier papers to cover descriptions of the details of the preparation of the alloy and its mounting as a bolometer, and deal here with the results which point toward the Sills hypothesis as the explanation of the detection observed.

![Cryostat or refrigerating can to maintain bolometer at 15.5 K](image)

Fig. 1—Cryostat or refrigerating can to maintain bolometer at 15.5 K. Bolometer is shown enlarged X2 in lower left corner.

### Apparatus

The bolometer is normally made from a 1-cm length of columbium nitride ribbon, 1 mm wide and 0.025 mm thick, cemented on the flattened face of a small copper rod by a thin layer of bakelite lacquer which insulates the ribbon from the rod electrically, but leaves it in good thermal contact. The normal resistance \( R_N \) at 15.8 K is about 1 ohm. Fig. 1 shows such a bolometer in the lower left-hand corner, and in the main drawing shows the refrigerating can or cryostat in which the bolometer is mounted. The cryostat holds a liter each of liquid hydrogen and liquid nitrogen, and will maintain the bolometer in the sensitive region of temperature for about twenty hours before refilling. A rock salt window is provided for passing in infrared signals; radio signals are sent down the electrical leads from the upper end of the can or through two special short leads (not shown) which can be placed next to the rock salt window.

![Electrical circuit](image)

Fig. 2—Electrical circuit.

The electrical circuit is shown in Fig. 2. Provision is also made for connecting a Brown recording potentiometer (BRP-1) across the bolometer to measure the IR drop due to the DC bias current \( (I_d) \). Another Brown recording potentiometer (BRP-2) equipped with a rectifier, can be connected in parallel with the oscillograph to record audio signals received. A type K, Leeds and Northrup potentiometer is sometimes connected di-

![Values of resistance (R/\( R_N \)) as a function of temperature (T) and bias current (\( I_d \)) for bolometer #29 R 24](image)

Fig. 3—Values of resistance \( (R/\( R_N \)) \) as a function of temperature \( (T) \) and bias current \( (I_d) \) for bolometer #29 R 24.
rectly across the bolometer to measure rectified potential from unmodulated rf input signals. The audio amplifier gives a band pass of about 200–2,000 cps, the signal from the bolometer passing first into a low-impedance transformer and then through two stages of audio amplification. The rectified signal will be denoted by $\Delta E_R$, and given in microvolts (rms) denoted by $\mu V$.

**RESULTS**

Fig. 3 shows the variation of resistance with temperature for five different values of direct current passing through the sample. By cross plotting these and interpolating, there were obtained the values of $dR/dI$ given in the second column of Table I corresponding to the values of dc in the first column labelled $I_d$.

![Figure 3](image3.png)

Time (Temp.)
Each Division = 40 Sec.

![Figure 4](image4.png)

Time (Temp.)
Each Division = 40 Sec.

Fig. 4—Superimposed Brown recording potentiometer tracings showing rectified voltage ($\Delta E_R$) obtained as temperature is varied from region of normal resistance through transition to region of superconductivity with various values of bias current. Lower curves show change of dc voltage with temperature for same region.

Fig. 4 shows a typical variation of audio signal with temperature and with bias current. This figure was prepared by cutting out and superimposing the BRP-2 tracings. The BRP-1 tracings showing the corresponding change of dc potential (and thus of resistance) are shown below each set, for 0.1 ma dc on the left and for 0.2 ma on the right. The rf signal used had a frequency of 1 Mc, 80 per cent modulated at 400 cps, and a current of about 1 ma was flowing through the bolometer. It is seen that the audio signal reaches a peak value at a temperature in the middle of the transition zone where $dR/dT$ is a maximum. It reaches a maximum with bias current at about 0.4 ma.

**TABLE I**

<table>
<thead>
<tr>
<th>$I_d$ (ma)</th>
<th>$dR/dI$ (ohm amp)</th>
<th>$\Delta E_R$ (calc.) (microvolts)</th>
<th>$\Delta E_R$ (obs.) (microvolts)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(a) 0.0</td>
<td>0</td>
<td>0</td>
<td>6.7</td>
</tr>
<tr>
<td>0.1</td>
<td>485</td>
<td>17</td>
<td>18</td>
</tr>
<tr>
<td>0.2</td>
<td>415</td>
<td>24</td>
<td>28</td>
</tr>
<tr>
<td>(b) 0.1</td>
<td>485</td>
<td>14</td>
<td>18</td>
</tr>
<tr>
<td>0.2</td>
<td>415</td>
<td>24</td>
<td>28</td>
</tr>
<tr>
<td>0.3</td>
<td>375</td>
<td>32</td>
<td>53</td>
</tr>
<tr>
<td>0.4</td>
<td>335</td>
<td>38</td>
<td>48</td>
</tr>
<tr>
<td>(c) 0.5</td>
<td>295</td>
<td>42</td>
<td>48</td>
</tr>
<tr>
<td>0.6</td>
<td>250</td>
<td>36</td>
<td>35</td>
</tr>
<tr>
<td>0.7</td>
<td>220</td>
<td>32</td>
<td>30</td>
</tr>
<tr>
<td>0.8</td>
<td>180</td>
<td>25.6</td>
<td>23.8</td>
</tr>
<tr>
<td>0.9</td>
<td>160</td>
<td>22.8</td>
<td>18.8</td>
</tr>
<tr>
<td>1.0</td>
<td>120</td>
<td>17.1</td>
<td>16.2</td>
</tr>
<tr>
<td>1.5</td>
<td>80</td>
<td>11.4</td>
<td>9.7</td>
</tr>
<tr>
<td>2.0</td>
<td>70</td>
<td>10.0</td>
<td>6.3</td>
</tr>
<tr>
<td>3.0</td>
<td>60</td>
<td>8.6</td>
<td>4.0</td>
</tr>
<tr>
<td>4.0</td>
<td>50</td>
<td>7.2</td>
<td>2.9</td>
</tr>
<tr>
<td>5.0</td>
<td>40</td>
<td>5.7</td>
<td>2.3</td>
</tr>
</tbody>
</table>

Values for $I_d=0.535$ ma

| (l) 0     | 0                | 0                                | 16                              |
| 0.1       | 485              | 27.5                             | 31                              |
| 0.2       | 415              | 47                               | 48                              |
| 0.3       | 375              | 65                               | 65                              |
| 0.4       | 335              | 79                               | 80                              |
| 0.5       | 295              | 83                               | 103                             |
| 0.6       | 250              | 90                               | 118                             |
| 0.7       | 220              | 96                               | 125                             |
| 0.8       | 150              | 106                              | 125                             |
| 0.9       | 100              | 108                              | >125                            |
| (c) 1.0   | 90               | 109                              | >125                            |
| 1.5       | 80               | 69                               | 92                              |
| 2.0       | 70               | 40                               | 41                              |
| 3.0       | 60               | 34                               | 35                              |
| 4.0       | 50               | 29                               | 26                              |
| 5.0       | 40               | 23                               | 22                              |

Values for $I_d=1.07$ ma

![Figure 5](image5.png)

Fig. 5—Maximum rectified voltage for bolometer #29R24 as observed and calculated (see Table I), plotted against bias current $I_d$ for two values of $I_{RF}$.
In the right-hand column of Table I there are shown the maximum (with temperature variation) audio voltages (rms) observed when passing in an rf signal at 1 Mc, 80 per cent modulated at 500 cps, using bolometer 29R24. The first part of the table gives the results obtained when an rf current of 0.535 ma was passing through the bolometer; and the second part of the table, when the rf current was 1.07 ma. Fig. 5 shows these results plotted as \( \Delta E_B \) against \( I_d \), the bias current. The calculated results in the third column will be discussed in the following section on Theory.

![Fig. 6—Rectified voltage (\( \Delta E_B \)) as a function of temperature (\( T \)) for constant bias current of 41 \( \mu \)a and rf current of 0.4 \( \mu \)a. Calculated values are shown as solid line, observed values as circles. Bolometer #29 R 24.](image)

Fig. 6 shows the values of rectified voltage obtained when an rf signal at 1 Mc, unmodulated, was applied, the rf current being 0.4 ma and the bias current, 41 \( \mu \)a, being held constant as temperature was varied. The calculated curve is discussed below in the section on Theory.

In order to show variation with frequency, percentage modulation, and direction of bias current, three other sets of curves from a previous paper\(^8\) are included as Figs. 7 and 8. These are the maximum audio signal voltages observed as temperature is varied in the superconducting transition. These maxima, in turn, are a function of the above variables.

**Theory**

Since we are dealing with a strip of metal mounted as a bolometer for infra red detection, it is natural to ask whether rectification is not due in part to heating. Fusion\(^7\) has made a careful study of the response of these bolometers to chopped infrared radiation and has determined their thermal time constants with considerable accuracy, finding values between 0.005 and 0.0005 second in general. The heat capacity of the CbN has been measured by Armstrong,\(^9\) so that the thermal conductivity away from the sample can be calculated in the usual manner. As previously shown,\(^6\) this is sufficiently large so that the heating from the relatively small currents used should produce less than 1 microvolt of rectified potential. Moreover, the bias current,\(^7,8\) which gives maximum response with infrared signals, i.e., with modulated external heating, is about 20 ma, whereas the bias current which gives maximum response with modulated rf is less than 1 ma. In addition to this, there are the large rectified voltages observed with unmodulated...
observed rectified voltages and their dependence on field of the current, can explain the major part of the observed rectified voltages and their dependence on temperature, rf current, and bias current as variables.

Consider first the condition of zero-bias current and the passage of unmodulated rf through the CbN. As the current rises from zero to its peak value \( I_{rf} \) in the first quarter of the rf cycle, the resistance will rise rapidly, giving a larger value of instantaneous potential across the sample at the peak value of the rf current than would be obtained for constant \( R \). However in the third quarter of the cycle with the current passing in the opposite direction, the same excess of potential is created in the opposite direction, since the resistance rises as a function of the magnetic field, and does not depend on the direction of the current. Thus no rectified potential is to be expected for zero bias current.

If, now, a bias current \( I_d \) is introduced, somewhat less than the peak rf current, the excess \( E \) will be greater when the instantaneous rf current \( I_{rf} \) flows with the bias current than when it flows against it, and rectification should result. Thus, if \( dR/dI \) were constant, the rectified potential \( \Delta E_R \) should rise to a maximum at the point where \( I_{rf} \) equals \( I_d \); and as \( I_d \) is increased beyond \( I_{rf} \), \( \Delta E_R \) should remain constant at its maximum value.

This can be summarized in the following equations, which are similar to the usual expressions for nonlinear resistance:

\[
E = R(I_d + I_{rf}) \tag{1}
\]

\[
R = R_0 + (dR/dI)(I_d + I_{rf}) \tag{2}
\]

\[
I_{rf} = I_{rf} \sin 2\pi f t \tag{3}
\]

where, in addition to the symbols already defined, we have \( R_0 \), resistance for zero current; \( E \), instantaneous potential; \( f \), frequency; and \( t \), time.

Substituting and integrating over a complete cycle, we get, for the rectified potential:

\[
\Delta E_R = \frac{1}{2}(dR/dI)I_{rf}^2 \tag{4}
\]

Actually \( dR/dI \) is not a constant but falls with increasing \( I \), as may be seen from Table I. The result is that \( \Delta E_R \) does not reach a constant maximum value with increasing \( I_d \) but goes through a maximum and then falls rapidly, as may be seen from Fig. 5, this fall being due to the rapid decrease in \( dR/dI \) at values where \( I_d > I_{rf} \).

The problem of calculating \( \Delta E_R \) thus falls into two parts:

(i) The region where \( I_d > I_{rf} \) and the variation of \( dR/dI \) is appreciable;

(ii) The region where \( I_d < I_{rf} \) and the rectification is incomplete, approaching zero as \( I_d \) tends to zero.

For the second of these two parts, we can use the approximate expression:

\[
\Delta E_R = \frac{I_d}{I_{rf}} \frac{1}{2} \left( \frac{dR}{dI} \right) I_{rf}^2 \tag{5}
\]

or we can integrate graphically from a plot of \( E \) against \( I \), which is constructed from the observed values of \( dR/dI \).

In Table I, the second and third values of \( \Delta E_R \) calculated in the series marked (a), have been obtained by graphical integration, the first value for zero bias current being identically zero.

The series of values marked (b) have been calculated from (5), using the mean value of \( dR/dI \) for the range covered.

The series of values marked (c) have been calculated from (4), in the range where \( I_d > I_{rf} \).

Because the rf was modulated, the rms values of the observed voltages have been used as an approximation to compensate for this additional complication in the variation of the current.

It is to be observed, first, that the difference between the values of series (a), calculated by graphical integration, differ only slightly from those of series (b), calculated on the basis of a mean value for \( dR/dI \). It thus appears that the shape of the curve for \( dR/dI \) is such that, for a rough comparison such as we are making, this approximation is permissible.

Considering the difficulties of the measurements, problems of temperature control, losses in the wires leading into and out of the low-temperature chambers, and the as yet unexplained frequency dependence, we cannot expect too close a correspondence between calculated and observed values actually found. The agreement may, under these circumstances, be regarded as tending to confirm the explanation of the rectified voltages as due to the Silsbee phenomenon.

The calculated curve in Fig. 6 is also obtained with the help of (5) and explains the variation of \( \Delta E_R \) with temperature as due to the variation of \( dR/dI \), which is a maximum in midtransition, and becomes zero at temperatures both above and below the transition.

The presence of rectified voltage in the absence of bias current is not explained by this theory, and must be attributed to some additional phenomenon which is dependent on the direction of the current, as may be seen from the curves in Fig. 7 (a), (b), (c).

Since the publication of the first report on superconducting rectification, several investigators have checked on the rectification to be expected when relatively large currents are used, in general, sufficient to carry the resistance completely through the transition.


12 Privately printed report under O. N. R. Contract, Physics Dept., Rutgers University, to be published shortly.
Generalized Theory of the Band-Pass Low-Pass Analogy*

P. R. AIGRAIN†, STUDENT, IRE, B. R. TEARE, JR.‡, SENIOR MEMBER, IRE, AND E. M. WILLIAMS‡, SENIOR MEMBER, IRE

Summary—A generalized theory of band-pass low-pass equivalents is developed using Laplace transform analysis, and it is shown that video equivalents exist not only for symmetrical systems, but also for dissymmetrical systems with low-level modulation. Several illustrations are given.

THE BAND-PASS low-pass analogy1 provides a means of simplifying the analysis of amplitude-modulated carrier systems which otherwise would be extremely tedious, especially when transient phenomena are involved.

This useful alternative analysis has not been extensively applied, however, possibly because its general theory, breadth of application, and limitations have not been widely understood. The purpose of this paper is to give a rigorous derivation of the band-pass low-pass analogy, and to discuss and illustrate its use in the classes of transient and steady-state problems in which it is applicable.

The concept of the equivalent video amplifier is illustrated by the two systems of Fig. 1. Fig. 1(a) shows a generalized amplitude-modulated carrier system comprising an oscillator as carrier source, a modulated stage in which the carrier envelope is made to vary linearly with the instantaneous amplitude of the signal a(t), a band-pass network which includes all the high and intermediate frequency sections of the transmitter, receiver, and transmission link, and a linear detector, the output b(t) of which is similar in form to the variable part of the carrier signal envelope at its input. In an ideal system b(t) would be identical (except for a possible constant multiplying factor) with a(t). However, in most realizable systems there are distinct differences between b(t) and a(t). The equivalent video system, shown in Fig. 1(b), is a hypothetical low-pass amplifier which, for the same input signal a(t), has a signal output of the same form b(t) as the band-pass system. The elements of the equivalent system are usually, though not necessarily, physically realizable.

Fig. 1—(a) Above, generalized amplitude-modulated carrier system. (b) Below, the video equivalent.

It has been pointed out by Cherry2 that the existence of an equivalent video system is a sufficient condition

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‡ Carnegie Institute of Technology, Pittsburgh, Pa.
for a linear relation between $b(t)$ and $a(t)$ and that the nonexistence of a video equivalent in any system is an indication of a nonlinear relationship between detector signal output and modulated stage signal input. For this reason the theory of the band-pass low-pass analogy takes on an interesting general significance quite apart from its utility in the solution of specific problems.

The theory of the video equivalent is derived by use of the Laplace transform. An alternative approach using Fourier transforms could be adapted from Cherry. We begin by considering the carrier system of Fig. 1(a) in detail. An oscillator supplies a sinusoidal voltage of unit amplitude and frequency $\omega_0$ to a modulated stage in which the modulating input signal $a(t)$ is applied. The modulated stage is assumed to have such idealized properties that its output is a voltage $1 + a(t)\sin \omega_0 t$, which is the input to a band-pass amplifier. The output of the band-pass amplifier is a voltage of form $[A_0 + b(t)]\sin (\omega_0 t + \phi(t))$, in which $A_0$ is a constant, and this voltage is supplied to a detector whose properties are assumed to be such that the output voltage is the variable part of the envelope of the input: that is, $b(t)$. The arbitrary part of the system, the band-pass amplifier, is described by its transfer function $G_B(j\omega)$, the ratio of output to input under steady-state ac operation. The transfer function may also be written as $A(\omega)e^{j\theta(\omega)}$ in which $A$ and $\theta$ are the amplitude and phase functions.

The problem is to determine under what conditions there is a low-pass amplifier which is equivalent to the carrier system, an amplifier which will give the same output $b(t)$ for the same input $a(t)$.

The form of the response $h(t)$ of the band-pass amplifier when the input is $f(t)\sin \omega_0 t$ may be expressed in terms of Laplace transforms as

$$h(t) = \frac{1}{2j} \left[ e^{j\omega_0 t}L^{-1}[F(s)G_B(s + j\omega_0)] - 3e^{-j\omega_0 t}L^{-1}[F(s)G_B(s - j\omega_0)] \right]$$

(1)

in which $F(s)$ is the transform of $f(t)$, and $L^{-1}$ indicates the operation of taking the inverse transform. If $s$ is considered to be real, this can be assumed without loss of generality, the expression (1) may be simplified$^6$ to

$$h(t) = 3e^{j\omega_0 t}L^{-1}[F(s)G_B(s + j\omega_0)]$$

(2)

This may be expressed more usefully by introducing new quantities $p(t)$, $q(t)$, $m(t)$, and $\phi(t)$ as follows:

$$L^{-1}[F(s)G_B(s + j\omega_0)] = p(t) + jq(t) = m(t)e^{j\phi(t)}$$

(3)

where $p(t)$ and $q(t)$ are the real and imaginary parts of the inverse transforms, and

$$m(t) = \sqrt{[p(t)]^2 + [q(t)]^2} = L^{-1}[F(s)G_B(s + j\omega_0)]$$

(4)

The symbols $R$ and $3$ will be used to designate operations of taking the real and imaginary parts, respectively.

$$\phi = \tan^{-1} \frac{q(t)}{p(t)} = \text{angle of } L^{-1}[F(s)G_B(s + j\omega_0)]$$

(5)

Then

$$h(t) = 3m(t)e^{j(\omega_0 t + \phi(t))} = m(t)\sin(\omega_0 t + \phi(t))$$

(6)

In the case under consideration

$$f(t) = 1 + a(t);$$

hence its transform is

$$F(s) = \frac{1}{s} + A(s)$$

(9)

and we are concerned only with $m(t)$, the envelope of the modulated carrier wave $h(t)$, since the variable part of $m(t)$ is the output of the system. From (4) and (9)

$$m(t) = L^{-1}\left\{ \frac{1}{s}G_B(s + j\omega_0) + A(s)G_B(s + j\omega_0) \right\}$$

(10)

It is assumed that the carrier frequency has been applied for a very long time, and it will next be shown that the first term of (10) is the steady-state response to the carrier alone. The final value theorem$^7$ gives for the first term of (10)

$$\lim_{t \to \infty} L^{-1} \frac{1}{s}G_B(s + j\omega_0) = G_B(j\omega_0).$$

(11)

In terms of amplitude and phase functions this is $A(\omega_0)e^{j\theta(\omega_0)}$. For brevity $A(\omega_0)$ will be written $A_\omega$ and $\theta(\omega_0)$ as $\theta_\omega$. Then $m(t)$ may be written as the absolute value of the sum of the two complex numbers,

$$m(t) = |A_\omega e^{j\theta_\omega} + L^{-1}A(s)G_B(s + j\omega_0)|$$

(13)

in which $s$ is real ($s$ is integrating variable and integration is over a contour in complex $s$ plane). The first term gives the response to the carrier alone and is constant, and the second term is, in general, a function of time. Reference to (2) to (5) shows that the whole expression (13) is the amplitude of a rotating vector $\omega_0$ which represents the output of the band-pass amplifier. At any time $t$ this expression may be represented by the vector diagram in the complex plane shown in Fig. 2. The out-

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put of the detector consists of the varying or ac component \( m(t) \), that is, the second term (13).

Consider now the equivalent video amplifier shown in Fig. 1(b) which has a transfer function \( G_{L}(s) \). The equivalence means that for the same input \( a(t) \), its output \( b(t) \) is the same as the variable part of the output \( m(t) \) of the amplitude-modulated carrier system. Now

\[
b(t) = L^{-1}G_{L}(s)A(s),
\]

and the problem of determining the video equivalent reduces to that of finding whether and when it is possible to find a function \( G_{L}(s) \) which is the transfer function of some circuit, not necessarily a physically realizable one, such that

\[
b(t) = m(t) - C.
\]

\( C \) being constant or the dc component of \( m(t) \) attributable to the first term of (13). A necessary condition for the existence of a single video equivalent circuit is that when \( a(t) \) is multiplied by a constant factor \( K \), \( b(t) \) is also multiplied by the same factor \( K \), and \( m(t) \) is changed to

\[
A e^{j \theta_{0}} + L^{-1}KA(s)G_{b}(s + j \omega_{0}) = KB(t) + C.
\]

Thus there is equivalence only if the variable part of \( m(t) \) is proportional to

\[
L^{-1}A(s)G_{b}(s + j \omega_{0})
\]

From Fig. 2 it may be seen that this is true in general for only two distinct conditions: Case I, in which \( A e^{j \theta_{0}} \) and \( L^{-1}A(s)G_{b}(s + j \omega_{0}) \) have the same phase angle \( \theta_{0} \) being real \( \theta_{0} \), and Case II, in which

\[
A e^{j \theta_{0}} \gg L^{-1}A(s)G_{b}(s + j \omega_{0})
\]

and changes in the resultant vector are proportional to the magnitude of the second, very small component. These cases will be considered separately.

Case I: \( A e^{j \theta_{0}} \) and \( A, G_{b}(s + j \omega_{0}) \) have the same phase angle \( \theta_{0} \), hence \( e^{j \theta_{0}}A_{1} \), \( A_{2}G_{b}(s + j \omega_{0}) \) is real. A comparison of (13) and (14) shows that the video equivalent exists and

\[
G_{1}(s) = e^{-j \phi_{1}}G_{b}(s + j \omega_{0}).
\]

The fact that both members of (16) are real means that \( A(\omega) \), the amplitude function for \( G_{b}(\omega) \), is symmetrical \(^4\) and the phase function \( \phi(\omega) - \phi_{1} \) is antisymmetrical.

\(^4\) That this symmetry is implied may be shown by considering the right-hand member of (10) to be a function of \( s \) and expanding it in a power series in \( s \):

\[
e^{j \theta_{0}}G_{b}(s + j \omega_{0}) = e^{j \theta_{0}} + j \omega_{0} e^{j \theta_{1}} + \omega_{0}^{2} e^{j \theta_{2}} + \cdots
\]

Since this function is real when \( \phi(s) \) is real, the coefficients are real. By substituting first \( s = -j \omega \) and then \( s = j \omega \) it may be shown that

\[
e^{j \theta_{0}}G_{b}(\omega + j \omega_{0}) \text{ is conjugate to } e^{j \theta_{0}}G_{b}(\omega - j \omega_{0}).
\]

Now the transfer function \( G(\omega) \) may be written in polar form \( A(\omega) \).

Thus

\[
A(\omega + j \Delta \omega) e^{j \Delta \omega}(s + j \omega - s) \text{ is conjugate to } A(\omega - j \Delta \omega) e^{j \Delta \omega}(s + j \omega + s).
\]

Accordingly, if \( \Delta \omega \) is any departure of frequency from \( \omega \)

\[
A(\omega + j \Delta \omega)(s + j \omega - s) \Delta \omega = A(\omega - j \Delta \omega) e^{j \Delta \omega}(s + j \omega + s) \Delta \omega
\]

where \( A(\omega) \) is a statement of the symmetry conditions.

about \( \omega = \omega_{0} \) and, conversely, any system which has this symmetry will make (16) real and falls into this case. The case in which there is symmetry of amplitude and phase characteristics about the carrier frequency is the case previously discussed in the literature. \(^1\) It may be observed that no physically realizable system can be exactly symmetrical about a frequency \( \omega_{0} \), since network theory requires that \( A(\omega) \) shall always be symmetrical about \( \omega = 0 \), and any function having two axes of symmetry must extend to infinity. Symmetry can be obtained approximately, however, if the amplifier pass-band is only a small fraction of \( \omega_{0} \). The effect of this approximation on the accuracy of results obtained from the video equivalent for any real network can be shown to be related to the "switching angle," or phase of the carrier voltage within its signal envelope. The actual transient response of an approximately symmetrical carrier system is independent of this switching angle at all instants of time except those immediately following the initiation of the transient, and it is during this initial period that the video-equivalent response is not very precisely indicative of the prototype carrier system response.

Case II:

\[
| A e^{j \theta_{0}} \gg L^{-1}A(s)G_{b}(s + j \omega_{0}) |
\]

This is the case in which the depth of modulation in the input signal to the detector is small. The expression (13) for \( m(t) \) can be simplified because of low modulation depth as follows:

\[
m(t) = A e^{j \theta_{0}} + \Re e^{j \theta_{0}}L^{-1}A(s)G_{b}(s + j \omega_{0})
\]

Because of inequality (17) the imaginary part contributes a negligible amount to the magnitude, and since \( e^{j \theta_{0}} \) is a constant independent of \( s \), it may be moved inside the operator \( L^{-1} \) giving

\[
m(t) \Rightarrow A_{b} + L^{-1}A(s) \Re e^{j \theta_{0}}G_{b}(s + j \omega_{0})
\]

By comparison of (18), (14), and (15) it is evident that a circuit with a transfer function

\[
G_{2}(s) = \Re e^{j \theta_{0}}G_{b}(s + j \omega_{0})
\]

is the video equivalent.

The case has not previously been fully developed in the literature although it may be deduced from Poch and Epstein. \(^4\) A small depth of modulation at the detector can be the result either of small modulation depth at the modulated stage or of reinserted or "exalted" carrier operation, such as with injection of a large locally

\(^1\) In taking the real part of the right-hand expression \( s \) is considered to be real. After having taken the real part \( s + j \omega \) is substituted in the new expression to obtain the equivalent video transfer function.

generated carrier, or by the use of a sharply tuned carrier emphasis filter.

It is important to note that the video equivalent of a circuit in this case is not the same as that of the same circuit or form of circuit under symmetrical conditions.

When the conditions of either cases I or II are not met, there is no video equivalent and, furthermore, the carrier system is nonlinear in the sense described in this paper. This case is treated in detail by Cherry, who derives a more elaborate equivalent low-pass system, the output of which is the square root of the sum of the squares of the individual outputs of two parallel linear low-pass networks. In conclusion, several illustrations are given in video equivalents for some common carrier systems.

Case I: Consider a carrier amplifier with the characteristics of a series R, L, C circuit, Fig. 3(a), to

\[ G_B(\omega) = R + j\left(\omega L - \frac{1}{\omega C}\right) \]

or, in terms of \( \omega_0 = 1/\sqrt{LC} \), with the approximation \( \omega/\omega_0 + \omega_0/\omega \approx 2 \),

\[ G_B(\omega) = R + \frac{2j(\omega - \omega_0)}{\omega_0^2C} G_B(s+j\omega_0) = R + 2sL. \]

and \( \theta_b \) is zero so that \( G_1(s) = R + 2sL \). The video equivalent, as is generally known for the case, consists of a resistance \( R \) in series with an inductance equal to twice the inductance in the original band-pass system.

As a second illustration consider the two-stage amplifier of Fig. 4(a), which is made approximately symmetrical for a carrier frequency \( \omega_0 \) by setting

\[ Q_1 = Q_2 \quad \frac{1}{Q} \ll 1 \]

\[ \omega_1 = \omega_0 + D \quad D \ll \omega_0 \]

\[ \omega_2 = \omega_0 - D \quad D \ll \omega. \]

The steady-state gain is of the form

\[ \frac{E_{out}}{E_{in}} = g^2 Z_2. \]

The impedance transfer function is

\[ G_P(\omega) = Z_1(\omega)Z_2(\omega), \]

or in symmetrical form

\[ G_B(\omega) = \frac{1}{C_2C_3[\omega_0^2Q^2 + 4D^2 - 4(\omega - \omega_0)^2 + 4Q\omega_0(\omega - \omega_0)]} \]

and

\[ G_B(s+j\omega_0) = G_L(s) = \frac{1}{C_0C_3[\omega_0^2Q^2 + 4D^2 + 4s^2 + 4Q\omega_0s]} \]

This video equivalent is realized in Fig. 4(b) if

\[ R = \frac{1}{C_0C_3[\omega_0^2Q^2 + 4D^2]} \]

\[ C = 4Q\omega_0C_0C_3 \]

\[ L = \frac{1}{Q\omega_0C_2C_3[\omega_0^2Q^2 + 4D^2]} \]

Case II: Consider again the series R, L, C circuit of Fig. 3(a) with \( \omega_0 = 1/\sqrt{LC} \), but in the case in which \( Q \) is too low to make the approximation \( \omega/\omega_0 + \omega_0/\omega \approx 2 \). Then it can be shown that

\[ G_L(s) = R + \frac{Ls}{1 + LCs^2}, \]

and the circuit of Fig. 3(b) is a video equivalent for low modulation levels.
Contributors to the Proceedings of the I.R.E.

Donald H. Andrews was born in SouthINGTON, Conn., on June 11, 1898. He spent one year at Phillips Academy, Andover, and entered Yale in the Fall of 1916, receiving the B.A. degree in 1920. In 1923 he was awarded the Ph.D. degree, and remained at Yale for one year as research assistant to Dr. John Johnston. Following this he went to the University of California on a National Research Fellowship to continue research in low temperature calorimetry during 1924–1925. He was granted an International Research Fellowship and went to Leiden, Holland, in August, 1925, where he continued research at the Physics Laboratory of the University in absolute zero temperature studies. In August, 1926, he was appointed Research Fellow at the Bartol Research Foundation of the Franklin Institute in Philadelphia, Pa., where he remained until September, 1927. At that time he came to The Johns Hopkins University to teach thermodynamics and direct research in calorimetry.

Dr. Andrews was promoted to associate professor in 1929, and to professor of chemistry in 1930, the chair he had held since that time. Since 1943 he has devoted full research time to the "Cryogeny Laboratory," a research unit which he organized for studies at very low temperatures. Dr. Andrews served as Director of the Chemistry laboratory from 1936 to 1942. He initiated the organization of an atomic research unit in 1936, which was carried on by members of the staff, later becoming associated with the Atomic Bomb Project. He has served as chairman of the Baltimore branch of the Association of Scientists for Atomic Education, and as a member of the Board for Foreign Policy Association, of the United Nations Association, and of the United World Federalists in Maryland. In 1936 he was asked to serve as a member of the First Scientific Commission of l'Institut International de Froid, and in 1948 he served as a representative of the United States on the organizational commission which met in Amsterdam.

For a photograph and biography of Pierre R. Aigrain, see page 900 of the August, 1949, issue of the PROCEEDINGS OF THE I.R.E.

W. Lindsay Black (M'36—SM'43) was born on February 8, 1900, at Asbury Park, N. J. In 1918 he joined the engineering department of the Western Electric Co., which became the Bell Telephone Laboratories in 1925, where he is now a member of the technical staff. He has been engaged in the development and installation of radio and carrier apparatus and systems, and, more recently, of apparatus and systems particularly, including such equipment for radio broadcasting and sound reproduction.

Mr. Black is a member of the Acoustical Society of America, the American Institute of Electrical Engineers, and the Audio Engineering Society.

Chester W. Clark was born in San Francisco, Calif., on July 18, 1906. He received the B.S. and M.S. degrees in chemistry from the University of California in 1927 and 1929, respectively. In 1935 he was awarded the Ph.D. degree in physics from the University of Leyden, Netherlands. During 1929–1933 he was associated with the Standard Oil Company of California, as a research chemist. He taught in the College of Chemistry at the University of California from 1935 to 1937, and at the San Francisco Junior College from 1937 to 1941. During the period 1941–1946 Dr. Clark was a member of the Ordnance Department of the U. S. Army, returning to teaching in 1946 at the Johns Hopkins University. In 1947 he became a low-temperature consultant and physicist at the Naval Research Laboratory, leaving there to accept the position of scientific assistant to the director, Ballistic Research Laboratories, at the Aberdeen Proving Grounds, Md., where he is now engaged. He holds the rank of lieutenant colonel in the U. S. Army.

Philip S. Jastram (M'45) was born in Providence, R. I., on February 28, 1920. He received the S.B. degree from Harvard University in 1943, and the Ph.D. degree from the University of Michigan in 1948. From 1943 to 1945 he was a member of the staff at Radio Research Laboratories in Cambridge. During the last two years, Dr. Jastram has been carrying on research in nuclear physics at the University of Michigan, where he is an instructor in physics. He is a member of the American Physical Society, Sigma Xi, and Phi Beta Kappa.
Contributors to the Proceedings of the I.R.E.

Ronald King (A'30-SM'43) was born on September 19, 1905, at Williamstown, Mass. He received the B.A. degree in 1927 and then the M.S. degree in 1929 from the University of Rochester, and the Ph.D. degree from the University of Wisconsin in 1932. He was an American-German exchange student at Munich from 1928 to 1929; a White Fellow in physics at Cornell University from 1929 to 1930; and a Fellow in electrical engineering at the University of Wisconsin from 1930 to 1932. He continued at Wisconsin as a research assistant from 1932 to 1934. From 1934 to 1936 he was an instructor in physics at Lafayette College, serving as an assistant professor in 1935.

During 1937 and 1938, Dr. King was a Guggenheim Fellow at Berlin. In 1938 he became instructor in physics and communication engineering at Harvard University, advancing to assistant professor in 1939 and to associate professor in 1942. He was appointed Gordon McKay professor of applied physics at Harvard University in 1946. Dr. King is a Fellow in the American Physical Society, the American Association for the Advancement of Science, and the American Academy of Arts and Sciences.

Gordon P. McCouch (S'41-A'45) was born on May 5, 1920 in London, England. He received the A.B. degree in 1941, and the M.A. degree in 1948, both from Harvard University. He has been a licensed radio amateur since 1936. From 1942 through 1945, he was a research associate at the Radio Research Laboratory; of this period, the latter two years were spent in England and continental Europe with the Telecommunications Research Establishment and the American British Laboratory. At this time he was engaged in field applications of radar countermeasures and VHF radiotelephone equipment; for this work he received the Army-Navy Certificate of Appreciation.

Since 1945 Mr. McCouch has been a member of the engineering staff of Aircraft Radio Corp., where he has been principally concerned with signal generator development. While on leave of absence from ARC during 1948, Mr. McCouch held a teaching fellowship at Harvard University.

G. G. MacFarlane (S'35-A'39) was born in 1916 at Airdrie, Scotland. He received the B.Sc. degree with first class honors in electrical engineering in 1937 from Glasgow University. From 1937 to 1939 he pursued postgraduate research at the Technische Hochschule, in Dresden, and was awarded the Dr. Ing. degree.

Since 1939 Dr. Macfarlane has been engaged in mathematical research at the Telecommunications Research Establishment, Ministry of Supply, in Malvern, England, where he is in charge of the theoretical physics division. He has published a number of papers on waveguides, tropospheric wave propagation, noise, and mathematics.

Dr. Macfarlane is an associate member of the Institute of Electrical Engineers, and a Fellow of the Physical Society, London.

Hermon H. Scott (M'35-SM'43) was born in Somerville, Mass., on March 28, 1909. He holds the degrees of B.Sc. and M.Sc. from the Massachusetts Institute of Technology, and was associated with the General Radio Co. for 15 years, first as sales and development engineer, and later as executive engineer in charge of audio frequency, acoustic, and broadcast developments. He is the inventor of RC tuned circuits and oscillators, the dynamic noise suppressor, and other circuits. Leading companies throughout the world are licensed to use these inventions, including Electric and Musical Industries, Ltd., the British manufacturers of His Master's Voice and Columbia records and phonographs.

At the present time, Mr. Scott is president and director of engineering of Hermon Hosmer Scott, Inc., Cambridge, Mass., which company manufactures dynamic noise suppressors, amplifiers, broadcast station and laboratory equipment, and sound-measuring apparatus. The company is also active in the development of electronic musical instruments.

Mr. Scott is also a fellow of the Acoustical Society of America, a member of the American Institute of Electrical Engineers, the Audio Engineering Society, and the American Institute of Physics, and associate member of the Society of Motion Picture Engineers. He is also Chairman of the Boston Section of the Institute of Radio Engineers.

Gordon Newstead was born in Cairo, Egypt, on July 1, 1917. In 1939 he received the degree of Bachelor of Engineering with Honours from the University of Melbourne, and also obtained first class Honors in Applied Mathematics. After graduation Mr. Newstead was employed as a senior demonstrator in electrical engineering by the University of Melbourne, and then as a communications engineer with the department of Civil Aviation. In 1945 he was awarded the degree of Master of Electrical Engineering for a thesis on aviation radio covering the work on which he was engaged while with the department.

Since 1944, Mr. Newstead has been lecturer in electrical engineering at the University of Tasmania, concerned with electronics and communications. His research work and publications have been on circuit theory and the ionosphere. Mr. Newstead is an associate member of the Institution of Engineers of Australia.

H. H. Scott
Correspondence

Positive-Grid Characteristics of a Triode*

In a recent paper by Wood, the grid in a positive-grid triode was considered as a virtual cathode, and the author sought to explain the deviation of his theoretical curves from the experimental ones by means of secondary emission, as stated in his observation (C). I cannot agree with him in his explanation.

At first, as Wood indicated in his paper, when the plate voltage is greater than the grid voltage, secondary-emission current, if any, can only flow from the grid towards the plate, and thus adds to the plate current. Consequently, any secondary emission present in this condition will simply alter his result in the opposite sense to his observation (C).

Regarding his deviations, I should like to state that they actually came from his assumption that the grid could be considered as a "virtual cathode." Thus, when he directly applied his equation (1), he actually fixed the field pattern of the total potential distribution to one which corresponds to the maximum or space-charge-limited current, assuming the grid plane to be replaced by an emitter. This is a limit which usually cannot be realized in an actual triode. Should we insist on assuming the grid plane as an emitter, the emission would be mostly temperature-saturated, instead of space-charge-limited.

In order to clarify this point, it is well to refer back to Jaffe's paper, where Jaffe indicated that the equation (1) used by Wood corresponds to the maximum current which can enter the grid and plate of a triode, when the grid voltage is greater than the grid voltage, secondary-emission current, if any, can only flow from the grid towards the plate, and thus adds to the plate current. Consequently, any secondary emission present in this condition will simply alter his result in the opposite sense to his observation (C).

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Institute News and Radio Notes

TECHNICAL COMMITTEE NOTES

The Standards Committee at a meeting held July 14 reviewed and approved a scope of activities for each of the technical committees, and will submit the plans to the Board of Directors for final approval. Preliminary arrangements were made for the formation of a technical committee on measurements. It was announced that a proposed standard for reference designations (for the identification of electrical, electronic, and mechanical parts and their associated graphic symbols) has been prepared by the Symbols Committee and presented to the Standards Committee and to the RMA for approval and for possible co-sponsorship. The Master Index of IRE, and tentative definitions prepared by the Definitions Co-ordinating Subcommittee under the chairmanship of Axel G. Jensen, will be published in August. This Subcommittee has set up a task group to co-ordinate the definitions of pulse, another group to deal with definitions of transducer and another with electronics. The joint IRE/AIEE group on Noise Definitions has been given full committee status. It is hoped this group will be able to co-ordinate definitions of all types of noise as related to various fields. A member from each of the technical committees has been appointed to represent his committee on the Annual Review Committee to prepare a Radio Progress Report for 1949 in the field covered by his committee. The Report will be published in the PROCEEDINGS OF THE I.R.E. early in 1950. Ralph Batchler, of Caldwell-Clements Inc., is serving as Chairman of the Annual Review Committee. The Audio Technical Committee, meeting August 2, approved a set of definitions prepared by the Subcommittee for Definitions of Audio Systems and Components under the chairmanship of O. L. Angevine, Jr. The definitions are being submitted to the Standards Committee for approval. Standardization work is underway on methods of measurement for audio systems and components. A subcommittee under the chairmanship of D. E. Maxwell. The Committee has established two subcommittees, one dealing with Noise in Audio Systems, and the other concerned with the Psychology of Aural Perception. The Committee on Electron Tubes and Solid State Devices held a meeting July 15. Approved and submitted to the Standards Committee for standardization was the following material: Grid Emission Currents, Methods of Testing, and Radio Frequency Operating Tests for Power-Output High-Voltage Tubes. The Sound Recording and Reproducing Committee, in session on July 28, formed a new Subcommittee on Optical Recording under the chairmanship of Everett Miller of RCA Victor. The Mechanical Recording Subcommittee, H. E. Roys, Chairman, has been engaged in testing for cutter calibration of disk recording. Because the art is so new, the Subcommittee has not been able to standardize on one method. It is felt that the material prepared by the group will be of value to engineers, however, and copies will be distributed through the IRE Professional Group on Broadcast Engineers. S. J. Begun's Subcommittee on Magnetic Recording has prepared a preliminary draft on Noise Methods of Measurement and Definitions. A proposed standard, Methods of Measurement of Television Signal Levels, prepared by a subcommittee of the Video Techniques Committee, is being circulated widely in the IRE and in industry to obtain comments on the work prior to its acceptance by the Committee. L. W. Morrison is Chairman. The Joint Technical Advisory Committee, convening at all-day sessions July 19 and August 1, was concerned primarily with the forthcoming FCC hearing and the results of the carrier offset tests conducted by JTAC at the FCC Laboratories at Princeton July 6 and 7. The report of these tests will be included in the JTAC's recommendations to the FCC for the September hearing. The JTAC's full report to the FCC will be published and issued as Vol. IV, Proceedings of the JTAC. A petition for the formation of a Professional Group on Broadcast and Television Receivers was approved by the Executive Committee August 9. The sponsor is Virgil M. Graham, Sylvanica Electric Products Inc., Flushing, L. I. The Administrative Committee of the Antennas and Propagation Group met July 15 at Washington, D. C. Constitution and By-Laws were adopted and a membership committee appointed. Plans are being made to hold this group's first national meeting in conjunction with the URSI meeting at Washington October 31 and November 2, 1949.

IRE/AIEE STATE SECOND ANNUAL CONFERENCE FOR OCT. 31, NOV. 1, 2

Electronic instrumentation in nucloinics and medicine will be the theme of the second annual IRE/AIEE joint conference scheduled for October 31, November 1 and 2 at the Hotel Commodore, New York, N. Y. "Evaluation of Radiation Hazards" will be the topic of a special round-table discussion planned for the evening session on November 1.

A four-man panel of scientific experts will lead the discussion and then will answer questions from the audience. Dr. Karl T. Compton, chairman of the Research and Development Board of the National Military Establishment, will address the gathering.

Following the program plan of last year's conference, the coming meeting will devote the first day's proceedings to electronics in medicine, the second day's to nucloinics in medicine, and the third day's to the physical aspects of nucloinics instrumentation.

Simultaneously with the conference the Nucloinics Manufacturers' Exhibit will be held on November 1 and 2 at the Hotel Commodore. This exhibit of nucloinics instrumentation is expected to be one of the largest ever held. Featured will be scalars, Geiger tubes, survey meters, scintillation counters, and industrial instrumentation utilizing radiation.

URSI—IRE GROUPS WILL CONFERENCE AT FALL SESSION IN WASHINGTON

In lieu of the regular fall session of the URSI and the IRE, a joint meeting of the USA National Committee of the URSI and the IRE's newly organized Professional Group on Antennas and Propagation has been scheduled for October 31, November 1 and 2, at Washington, D. C. The session will convene at the National Academy of Sciences, 2101 Constitution Ave., N.W., and in the auditorium of the State Department Building, 21 St. and Virginia Ave., N.W.

Sponsoring the conference are the National URSI Commissions as follows: Commission 2—Tropospheric Radio Propagation, Chairman: Dr. Charles R. Burrows, Director of School of Electrical Engineering, Cornell University; Commission 3—Ionospheric Radio Propagation, Chairman: Dr. Newbern Smith, Chief of Central Radio Propagation Laboratory, National Bureau of Standards; Commission 5—Extraterrestrial Radio Noise, Chairman: Dr. D. H. Menzel, Harvard College Observatory; Commission 6—Radio Waves and Circuits, including General Theory and Antennas, Chairman: Dr. L. C. Van Atta, Naval Research Laboratory.

The IRE Professional Group on Antennas and Propagation will be co-sponsors of the sessions of Commissions 2, 3, and 6.

FLORIDA BROADCASTING COMPANY STARTING TELEVISION PROGRAMS

Radio station WMBR-TV, owned by the Florida Broadcasting Company, commenced broadcasting test patterns last month, and planned to open officially early in October with local television program with a four and one-half hour daily and Sunday schedule. Station officials reported that it might be "at least a year" before network programs can be picked up for local television fans.

WMBR-TV purchased complete transmitter and studio equipment from General Electric, including a 3-day antenna which is mounted on a 500-foot tower, and other equipment such as wave-form rack, film camera channel, rack of transmitter monitoring equipment, and two 16-mm movie projectors.

ELECTRONICS CONFERENCE AIDS RESEARCH, NEW DEVELOPMENTS

Electronic research, development, and application were the concern of the National Electronics Conference held September 26, 27, and 28 at the Edgewater Beach Hotel, Chicago, III., with technical sessions devoted to the general topics of Electronic Instrumentation, Solid-State Studies, Computers, Television, Antennas, Measurements, Magnetic Amplifiers, Research Management, Theory of Communications, Vacuum Tubes, Electromagnetics, Superconics, and Audio-Frequency.

1159 PROCEEDINGS OF THE I.R.E.
Address Sessions of the Radio Fall Meeting conducted under auspices of the IRE and the RMA Engineering Department on October 31, November 1 and 2, at the Hotel Syracuse, Syracuse, N. Y.

R. H. Williamson, chairman of the technical session starting at ten o'clock in the morning of the conference's opening. The session will be devoted to discussion of the following topics: "Measurement of Transient Response of Television Receivers" by J. Van Duyne, of Allen B. DuMont Laboratories Inc.; "Television Transient Response Measurement" by Jerry Minter, of Measurements Corporation; and "Understanding Requirements for Television Receivers" by K. S. Geiges, of Underwriters' Laboratories, Inc.

The afternoon session with J. R. Steen, chairman, will hear the following addresses: "Quality Control from the Producer and Consumer Viewpoints" by A. B. Mundel, of Sonotone Corporation; "Quality Control Gets a Job in Television Manufacturing" by L. Lutger, of Allen DuMont Laboratories Inc.; and "The Navy's Interest in Quality Control" by William R. Palst, Jr., Navy Department, Bureau of Ordnance.

Dr. Ralph E. Lapp of the Research and Development Board, Washington, D. C., will speak on the topic, "The Atomic Bomb and National Security," at the evening session, jointly sponsored by the Syracuse Section of the IRE and the Technology Club of Syracuse.

Chairman of the November 1 morning session is R. A. Hackbusch who has arranged the following program: "An Inter-carrier Sound System for Television Receivers Using the 6BN6" by Walter Stroh, of Zenith Radio Corporation; "Simplification of Television Receivers" by W. B. Whalley, of Sylvania Electric Products Inc.; and "Came the Television Revolution" by Dorman B. Israel, of Emerson Radio and Phonograph Corporation; "Universal Application-Cathode Ray Sweep Transformer with Ceramic Iron Core" by C. E. Torsch, of General Electric Company. The afternoon session, with J. E. Brown, as chairman, will hear addresses on topics "An Evaluation of Television Viewing Filters," by L. E. Martin and R. M. Bowie; and "Characteristics of High-Efficiency Deflection and High Voltage Supply Systems for Kinescopes" by O. H. Schade, of RCA. At the dinner meeting, Kenneth W. Jarvinen, will discuss "The Engineering Aspects of Sin.

At the November 2 morning session, the following topics will be "Pickup Tracking" by H. E. Roys, of RCA Victor Division; "New Audio Amplifier Circuit" by Frank H. MacIntosh, consulting engineer; "The Safety-Vox, A Novel Continuous Tape Magnetic Recorder Reproducer for Industrial Purposes" by R. A. York, of General Electric Company; and "A New Type of Dual Core Loudspeaker" by Harry F. Olson and John Preston of RCA Laboratories and J. H. Cunningham, RCA Victor Division.


RIDER PUBLISHES INDEX FOR ELECTRONICS INFORMATION

John F. Rider, president of the Electronics Research Publishing Co., Inc., has announced that the Electronics Engineering Master Index for 1947 and 1948, and the index for January-June, 1949, will be published next month.

According to Mr. Rider the institution of a new idea in the contents of these publications will be of significance to the electronic engineering industry. Beginning with 1947-1948 index, the contents will cross reference published technical papers and related U. S. patents allowed during the period covered by the volume. The work will provide a single source listing all technical articles published in the United States and abroad, the titles, patents, serial number and number of claims, of all electronic and related-subject patents issued in the United States during 1947 and 1948.

ATOMIC UNIT GRANTS FUNDS TO MINNESOTA UNIVERSITY

Construction of a 50 million-volt proton linear accelerator, which will be the world's most powerful "atom smasher" of its type, will be started at the University of Minnesota this year under a $728,000 grant from the Atomic Energy Commission, according to an announcement by President J. L. Merrill of the University.

DuMont Laboratory Studies New Television Allocation

Dr. Thomas T. Goldsmith, Jr. (A'38-SM'46-F'49), research director of Allen B. DuMont Laboratories Inc., has announced the development of a program to utilize vhf and uhf as integrated components of a national television system that will assure viewers a choice of television services provided by a multinet system and protect owners of television receivers from the need to buy additional equipment, including converters and extra antennas. The DuMont plan, which was presented to the FCC at a hearing in September, also provides 10 additional channels for service to smaller localities and seven uhf channels for the use of educators. It constitutes an alternative to the proposals put forward July 11 by the FCC.

According to Dr. Goldsmith, "DuMont enthusiastically approves the policy of providing more channels for additional television services, but there are certain features of the FCC proposal which should be modified to provide better service to the public and a more competitive national television operation.

For example, while the FCC program provides four vhf channels in only 20 of the first 50 cities of the United States, under the DuMont proposal, approximately 35 of the same 50 cities will be provided with four or more vhf channels.

Dr. Goldsmith also states that the DuMont studies, which have been carried on for nearly two years, indicate that FCC proposed allocations, mixing uhf and vhf frequencies in the same market, would impose an unnecessary financial burden on both station owner and receiving set owner.

Calendar of COMING EVENTS

National Radio Exhibition, Olympia, London, England, September 28 to October 8

SMIE 66th Semiannual Convention, Hollywood, Calif., October 10-14

AIEE Midwest General Meeting, Cincinnati, Ohio, October 17-21

Technical Conference on Antennas, Kansas City Section, IRE, Kansas City, October 28-29

Radio Fall Meeting, Syracuse, N. Y., October 31, November 1-2

1949 Nucleonics Symposium, New York City, October 31, November 1-2

URSI/IRE Joint Meeting, October 31, November 1-2, Washington, D. C.

Southwestern IRE Conference, Dallas, Tex., December 9-10

1950 IRE National Convention, New York, N. Y., March 6-9
Industrial Engineering Notes

National Bureau of Standards develops Electron-Optic Tool

As the result of a series of electron-microscope experiments at the National Bureau of Standards, Dr. L. L. Marton of the Electron Physics Laboratory has developed an electron-optical shadow technique which provides a valuable tool for the quantitative study of electrostatic and magnetic fields of extremely small dimensions. An electron-optical shadow is produced by using a fine wire mesh placed in the path of the electron beam. From the distortion in the shadow network caused by the deflection of the electrons as they pass through the field under study, accurate values of the field strength are computed.

According to Dr. Marton, it is thus possible to investigate quantitatively fields that have not been susceptible to other methods of investigation; for example, the fringe fields from the small domains of spontaneous magnetization in ferromagnetic materials. The technique, which is based on extensive theoretical analysis, should provide a powerful means for broadening present knowledge concerning space-charge fields, fields produced by contact potentials, patch fields in thermionic emission, charge distribution in a gaseous plasma, waveguide problems, and the basic magnetic properties of metals. Though similar in some respects to the electron-optical Schlieren method previously developed at the Bureau, the shadow method is better adapted to precise determinations of field intensity.

According to the Bureau, perhaps the greatest value of the electron-optical shadow method lies in its utility for exploring complex electric and magnetic fields of extremely small dimensions in which a probe of size greater than the electron would disturb the field under study. It provides data for accurate calculation of the absolute value of the intensity in the neighborhood of a specimen of any size or shape without altering or disturbing the field.

Television News

The FCC has issued proposals for expanding television broadcasting into the uhf band and has scheduled a broad television hearing. The proposals are based on proposed allocations and consideration of color video proposals as the first step toward lifting the current “freeze” on new television stations. The FCC simultaneously has reaffirmed previous uhf allocations with only three assignment changes in as many cities. The television allocations provide for 12 uhf channels, plus retention of the present 12 vhf bands, both on present tv standards.

They would permit construction of a nationwide television system embodying a total of 2,245 television stations in 1,400 cities and communities. Uhf channels alone allow for 1,702 stations in 1,179 areas in addition to any rural coverage. Uhf channel allocations, in general, are in virgin territory with a minimum of overlap between uhf and vhf stations, as recommended early this year by the FCC Conference on Broadcasts in Color, FEMA. For instance, no channels were assigned to New York City, Chicago, or Los Angeles, and only one uhf channel was allocated to Philadelphia and Washington. Baltimore got two uhf channels. Uhf channels will be numbered 141 to 164. They will be allocated for metropolitan television stations and for community stations. The uhf assignments will begin at 470 or 500 Mc, depending on the outcome of a Bell Laboratories petition for assignment of the 470-500 Mc band to multichannel mobile radio communications. Vhf channel assignments are identical with previous FCC allocations except in three cities, Cleveland, Syracuse, and Rochester, where channel changes within vhf were proposed. With the view of encouraging the adoption by foreign countries of the American Standards Association, the U.S. Delegation attending an international meeting on Television standards at Zurich, Switzerland were authorized to invite representatives of the study group to attend a television demonstration in this country.

The position of the various delegations with respect to the number of picture frames and lines was as follows: U. S. 30 frames, all others 25; British and French, 405 and 819 lines, respectively; U. S. 525; Denmark, Hungary, Netherlands, Sweden, Switzerland, and Czechoslovakia, 625 lines. The Conference agreed that for this study was required to determine the maximum brightness usable with 50 and 60 cycles per second.

...A dispatch from the American Consul General in the Netherlands to the U. S. Department of Commerce indicates that experimental television broadcasting stations are operating in that country. The government has no official plans for regular broadcasts. An experimental station at Eindhoven, uses 567-lines interlaced 2x25 (synchronized on 50-cycle line frequency). There is also an experimental television transmitter at Gronigen, built by amateurs, which broadcasts C B S a rgomter images to 4 amateurs, using a 250-line system. The Eindhoven station uses 67.75 Mc for video and 63.75 for sound. Following receipt of complaints from television receiver owners that diathermy equipment and some other radiation devices have been the cause of considerable interference to television broadcasts, the FCC has proposed to amend its rules to clarify the responsibilities of operators of these devices. It has announced the following statement: “an analysis of these complaints shows that in many instances the interfering signal is the fundamental signal emitted by the diathermy or other equipment in the channel centering on 27.120 Mc prescribed for such use in Part 18 of the Commission’s rules.” The interference involved results from the fact that most TV receiving sets and some other receiving sets have receivers tuned to the fundamental frequency of the broadcast channel. More channels may be available in the future to the ability of these crystals to hold precisely to an assigned frequency. Chances of overlapping or interfering point-to-point by another on an adjacent channel will be greatly reduced, the Signal Corps said. The discovery involves the super-heating to approximately 900 degrees Fahrenheit, followed by slow cooling for 24 hours.

FCC Actions

Two stations, one in Baltimore and one in New York, have been authorized by the FCC to conduct experimental television broadcasts in color with the view of further studying compatibility of color standards and the present black-and-white standards and the problem of converting present television receivers to color. The FCC has granted authority to begin broadcasting in color during the hours when its regular television programs are not being broadcast in order to gather information for the FCC television hearings. The other action authorized station WMAR-TV in Baltimore to broadcast color television programs at Johns Hopkins University Hospital in connection with the American Medical Association convention in Washington, and also to gather information for the television hearings. WMAR utilized equipment furnished by a Philadelphia pharmaceutical concern which previously demonstrated color television operations at Philadelphia and Atlantic City. The program was microwaved from Johns Hopkins to WMAR-TV and then broadcast to Washington. Letters have been sent to all experimental television stations asking for reports on their research and experimentation in both the uhf and vhf bands. Full details on the operation and findings of these stations, including questions on color if that type of transmission has been undertaken, have been asked. With respect to color, the FCC requested information concerning color break-up, flicker, color fringing, image registration, color fidelity, picture brightness, camera light efficiency, definitions, field tests and details with respect to modification of transmitters and receivers to provide the degree of compatibility contemplated by the recent television notice of the Commission.

Bound volumes of decisions and orders of the FCC, exclusive of annual reports, covering the period July 1, 1945, through June 30, 1947, are now on sale at the Superintendent of Documents, Government Printing Office, Washington, D. C., at $3.75 a copy, according an announcement by the RMA.

Signal Corps Crystal Discovery May “Revolutionize” Industry

Industry is expected to benefit from a new discovery in the processing of quartz crystals, according to an announcement by the U. S. Army Signal Corps. The development which may revolutionize the quartz crystal industry “promises proportionate savings in the future for all types of modern communications services, including television, radio broadcasting, point-to-point communications, and in hundreds of types of military equipments which depend on crystal control for operation.”

More channels may be available in the future due to the ability of these crystals to hold precisely to an assigned frequency. Chances of overlapping or interfering point-to-point by another on an adjacent channel will be greatly reduced, the Signal Corps said. The discovery involves the super-heating to approximately 900 degrees Fahrenheit, followed by slow cooling for 24 hours.
IRE People

Ralph A. Hackbusch (A'26-M'30-F'37), formerly vice-president and general manager of Stromberg-Carlson Co., Limited of Canada, has been named president and managing director of that company.

Born in Hamilton, Ontario, on September 18, 1900, he was educated at Hamilton and later specialized in electrical engineering. Prior to his employment with Stromberg-Carlson Co., he was associated with the Canadian Westinghouse Company and with the Canadian Brandes Co. (Kolster Radio).

In 1930 he became chief engineer and factory manager of the Stromberg-Carlson Telephone Manufacturing Company in Toronto, a position he held for a period of ten years. In 1940 he was named vice-president and general manager. During the war Mr. Hackbusch was requisitioned by the Canadian government and placed in charge of the radio division of the government's Research Enterprises Ltd., later being elected its vice-president in charge of radio and director of the radio division. He returned to Stromberg-Carlson as managing director, a position he held until his recent election by the board of directors for office as president.

Mr. Hackbusch, who was elected vice-president of the IRE in 1944, served as Chairman of the Toronto Section of the Institute in 1933, a member of the Sections committee in 1936, and on the Board of Directors in 1938, 1944-1946. He was the director of engineering for the RMA of Canada for ten years, member of the Canadian electrical code committee, the main committee of the Canadian Engineering Standards Association, and the RMA's general standards committee. In addition he has been a guest member of the joint coordination committee of the Edison Electric Institute, the National Electric Manufacturers Association, and the RMA. He has for many years been a guest member of the general standards committee of the Radio Manufacturers Association of the United States.

A professional engineer in the province of Ontario, he has also served as the official observer on the main RTPIB committee for the RMA of Canada, as vice-president of the Canadian Radio Technical Planning Board, and as chairman of the Canadian Building Fund Committee.

Frederick E. Llewellyn (A'23-F'38), consulting engineer of Bell Telephone Laboratories, has been elected a member of the executive committee of the United States National Committee of the International Electrotechnical Commission to serve for the coming year.

Dr. Llewellyn, who served as President of IRE in 1946, has been concerned with radio and circuit research and analysis of the electronic behavior of vacuum tubes at high frequencies. He was awarded the Morris Liebmann prize for his outstanding original work on constant-frequency oscillators and on vacuum-tube electronics at high frequencies in 1935.

Ernst F. W. Alexanderson (A'13-M'13-F'15), consultant to the General Electric Engineering and Consulting Laboratory, was recently named "Man of the Year" at Sweden Day ceremonies attended by 5,000 Swedes and Swedish-Americans.

Dr. Alexanderson was born in Upsala, Sweden, on January 25, 1878. He was graduated from the Royal Technical Institute in Stockholm and did post-graduate work at the Royal Technical Institute at Charlottenburg. Prior to his recent retirement, Dr. Alexanderson had completed forty-six years of service with General Electric.

During that time he was awarded 314 patents, or an average of one patent every seven weeks. The Alexanderson high-frequency alternator made possible reliable trans-oceanic radio transmission. His other patents cover such fields as radio telephony, radio telegraphy, television, vacuum tubes for radio and power rectifiers, and electronic circuits.

Dr. Alexanderson was the recipient of the IRE's Medal of Honor in 1919 and served as the Institute's President in 1921. He has published numerous papers in the PROCEEDINGS OF THE I.R.E. and in the Journal of the AIEE. He is also a Fellow of the AIEE.

Frederick Emmons Terman (A'25-F'37), Dean of engineering at Stanford University, has been elected vice-president in charge of the Engineering College Administrative Council of the American Society of Engineering Education as the result of the Society's annual convention held this summer at Rensselaer Polytechnic Institute. More than 1,700 engineering educators, administrators of engineering colleges and industrial leaders attended the session.

Born on June 7, 1900, in Engle-h, Inl., Dr. Terman received the B.S. degree in 1920 and the degree of engineer in 1922 from Stanford University, the S.E. degree from Massachusetts Institute of Technology in 1924. From 1925 to 1937 he served as instructor, assistant professor, and associate professor of Stanford's electrical engineering department. In 1937 he was named professor and head of the department.

During the war Dr. Terman assumed the directorship of the Radio Research Laboratory at Harvard University which developed counter measures to nullify radar equipment. He was Vice-President of the IRE in 1940 and President in 1941.

Herald S. Osborne (A'14-M'29-SM'34-F'15), chief engineer for the American Telephone and Telegraph Corporation, has been elected by the United States National Committee of the International Electrotechnical Commission to serve as its chairman for the coming year. The IEC is the electrical division of the International Organization for Standardization.

Dr. Osborne spent the summer in Europe serving as one of the United States delegates to IEC meetings held in June at Stresa, Italy.

A former vice-president and treasurer of the USNC, Dr. Osborne also is affiliated with the American Institute of Electrical Engineers, Acoustical Society of America, and American Physical Society. He is a member of the Joint Conference Committee on Standards, U. S. Department of Commerce, and the Industry Advisory Council, Federal Specifications Board. He is also vice-president and member of the Board of Directors of the American Standards Association.

Edwin A. Speakman (A'43-M'44-SM'48), branch head of the National Radio Laboratory's Countermeasures Section, has been appointed executive director of the Committee on Electronics, Research, and Development, National Military Establishment.

Associated with the National Radio Laboratory since 1940, Mr. Speakman has held his present post since 1943. In 1947 he received the Navy's Meritorious Civilian Service award for his work in radar.

Mr. Speakman was born in Graz, Pa., on August 14, 1909, and has been a radio amateur since 1925, with the call letters W3AUR and W8RSJ. While attending Haverford College in Haverford, Pa., from which he was graduated in 1931 with the bachelor of science degree in physics, he invented the first automatic photoelectric timing system for use in timing sports events. He served as instructor in radio at Haverford for three years following his graduation; then became a radio engineer with the Philco Corp. in Philadelphia and Detroit, remaining with that company until 1939.

Subsequently he was employed for a year as physicist by the Curtis Publishing Co. in Philadelphia, where his work consisted of research in electron-physics as applied to television receiving processes. His specific duties were to investigate special electronic devices for the improvement of high-speed four-color printing.

The inventor of the telescopic rod antenna system, Mr. Speakman is a member of the American Physical Society, the U. S. Naval Institute, and the Board of Civil Service Examiners for the Potomac River Naval Command.

Charles N. Kimball (A'34-M'40-SM'43) was recently appointed technical director of the Bendix Aviation Corp.'s research laboratories in Detroit, Mich.

Born in Boston, Mass., on April 21, 1911, Dr. Kimball received the B.E.E. degree from Northeastern University in 1931, and the S.M. and Sc.D. degrees in communications from Harvard University in 1932 and 1934, respectively. After graduation he was employed by the National Union Radio Corp. as a tube development engineer until 1935, when he joined the research staff of
the Radio Corporation of America’s license division laboratory in New York City. There he worked on television, frequency modulation, and special communication studies.

From 1940 through 1946 Dr. Kimball was vice-president in charge of engineering at the Aireon Manufacturing Corp. in Kansas City, as well as a member of the Board of Directors. His duties included supervision of work on a variety of projects, including radio countermeasure equipment and underwater sound techniques. He was one of the original workers in the field of high-frequency communications for railroads. In 1947 and 1948 he worked privately on instrumentation problems in the food processing industries.

The author of several papers on television and electronic instrumentation, Dr. Kimball holds patents in these fields. He is a member of Tau Beta Pi and the Harvard Engineering Alumni. He is also a former officer of the Kansas City Section of the Institute.

Books

Principles of Radar, by Denis Taylor and C. H. Wescott

Published (1949) by the Cambridge University Press, 51 Madison Ave., New York, N. Y. 136 pages, 8-page index-rex pages, $6. figures. $5.50.

Radar, a specific application of the principles of radio, was responsible for a large-scale development of new radio techniques. Much of modern radio technique is consequently erroneously classified in current literature as radar. As one turns weakly from the resulting congested volumes under that title, the first reaction to Taylor and Wescott is that it can’t be done in 136 easily readable pages.

The authors, however, have very nearly succeeded in doing exactly that! Each fundamental of the “Principles of Radar” is lucidly set before the reader in a volume that will delight any technical student. In fact, so condensed and so well classified is the material that the work could easily pass as an engineering handbook on radar. However, it is by no interpretation a mere tedium of facts. Based on systems engineering concepts, the quantitative facts mathematically expressed are backed up by qualitative logic clearly identifying the physical significance of every term, and further clarified by numerical examples based on historical equipments.

In only rare exceptions is clarity sacrificed to brevity. It is suggested that the lack of exact correspondence between text and figure 4.5 on “Electronic Range Marker” and omission of a statement on axial orientation of the microwave variable elevation beam equipment on page 94, be corrected in future editions. In both cases, the intent is clear to the expert, but possibly confusing to the student.

The question of what to exclude ultimately taxes the judgment of any author who attempts to put so much information in so little space. The omission of any treatment of signal storage, radar camouflage, and corner reflectors might be debated.

To one familiar with the development and application of radar on both sides of the Atlantic, this text displays a disappointing national limitation. The historical introduction, the examples, the application problems, the techniques themselves, are all limited to British experience. Were a similar work based on American experience, some statements of fact would appear in disagreement with Taylor and Wescott, simply because American and British radar developments, while similar, were not identical. Historically, original conceptions and significant observations would be earlier in the American Version, with development more rapid and military application earlier in the British version. Technically, American experience in common antenna working, rapid scan, IFF, and shipborne radar would lead to rather different treatment of these subjects, and there might be disagreement on the relationship between bandwidth and range accuracy due to employment of different techniques. This restriction renders the book a bit thin, both historically and technically, in those areas where the contribution was significantly American.

In spite of the above-noted limitation, since no American equivalent is known to this reviewer Taylor and Wescott’s “Principles of Radar” is recommended as a text in radar engineering for those who prefer less detail than is found in Ridenour’s 735-page microwave version of “Radar Systems Engineering.” It is also recommended as a handbook for radar systems engineering for practicing engineers. It is excellent for the technically inclined reader who seriously wants to know about radar. The merely curious will find it intellectually too strenuous.

Transformations on Lattices and Structures of Logic, by Stephen A. Kiss

Published (1948) by Stephen A. Kiss, 11 E. 92 St., New York 28, N. Y. 315 pages + 4-page index + 4-page bibliography + 4 pages. 6 figures, 7 X 10.

Mathematics develops by successive abstraction and generalization. Modern algebra is particularly subject to this tendency, with many of its concepts, such as those of group and lattice, now permeating the entire range of mathematical thought. In this book Dr. Kiss develops a generalization of Boolean algebra based upon its lattice theoretic properties. Naturally such a work will be of principal interest to the logician and the pure mathematician.

The book falls into three main parts of about the same length. The first consists of an orderly exposition of the pertinent results from modern algebra and lattice theory. In the second part the author defines his generalized Boolean algebras with 4, 8, 16 or generally 2^n elements, and develops their algebraic and lattice theoretic properties. These algebras have a natural and elegant mathematical structure. The third part of the book is an exposition of the calculus of classes and of propositions, and an application of the preceding algebraic theory to an extension of these fields.

Classical Boolean algebra and the calculus of propositions have been applied in such fields as switching theory and the study of nerve networks, and it is possible that the author’s extensions of these disciplines may find similar applications. Dr. Kiss conjectures that an adequate description of physical and biological phenomena require 4 and 16 class logics, without, however, developing this thesis. At any rate, considered as a purely algebraic theory, the author has added a significant contribution to mathematical literature.

Institute News and Radio Notes

ROBERT M. PAGE
Naval Research Laboratory
Washington 25, D. C.

Claude E. Shannon
Bell Telephone Laboratories, Inc.
Murray Hill, N. J.

1949

Robert D. Teasdale (S’44–A’46–M’49) has joined the teaching staff of the School of Electrical Engineering, Georgia Institute of Technology. Dr. Teasdale has been appointed associate professor of electrical engineering. Recently he earned the Ph.D. degree from Illinois Institute of Technology, where he has done some part-time teaching.

Dr. Teasdale, who is a native of Pennsylvania, received the bachelor’s degree at Carnegie Institute of Technology under a George Westinghouse Scholarship. He held a Gerard Swope Fellowship for 1947-1948, and an RCA Fellowship in Electronics for 1948-1949. He is a member of the American Institute of Electrical Engineers, Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.
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What Kind of Engineers?∗

J. D. RYDER†

The evolutionary theory teaches that modern man developed from some one-celled organism, through successive stages of improvement and specialization as water-living organisms, crustaceans and fishes, as amphibians, as land-living reptiles, and on into the present stage as a land-living mammal. It has been said that before birth a human embryo passes through a number of stages, each indicative of a phase in human evolution. While inaccurate, this statement implies a gern of truth—a human being, as a physical and mental assemblage of relatively worthless materials, is a product of all that has gone before in the history of life on this earth. It builds on experience and experiments and, it is to be hoped, profits by her mistakes! However, some samples seem to indicate that only a small profit has been taken.

In the field of science we also build and profit on the lessons and experiences of the past. The difference between the accumulated experiences of the human animal and those of the human scientist may be proportional to the difference between one million years and 300 years, as the probable respective ages of the two. In considering the human animal, it is customary to state that we have come a long way from the early cave man; but perhaps in speaking of the human scientist we should say we have not yet come very far.

In such a discussion it is necessary to use the words “science” and “engineering” interchangeably, since, in the early days, what is now engineering was science, and what is now science will later be engineering. In order to decide where science or engineering now is and, therefore, what kind of engineers we should be turning out of our colleges at the present time, it is necessary to go back and scan the scientific progress.

It has been said that in the whole history of human endeavor going back about one million years there have been only four inventions, or great developments which have radically altered the course of human history. These are considered to be: The art of making fire; the invention of the wheel; the invention of the alphabet; and the harnessing of electricity. It is interesting to note that we are now living in the period of time in which the fourth of these great changes is occurring. Is it any wonder that strife and turmoil exist? Is it not possible that those having the secret of fire were alarmed, troubled, and attacked by the hordes of that era? One can readily imagine the labor troubles arising if thousands of slaves carrying burdens on their backs were suddenly displaced by use of the roller or by carts with wheels. We talk of our modern technological and labor-saving advances, but is mankind in a situation really much different from that existing at the time of the invention of the wheel? We continue to live as tribes, to regard nonmembers of our tribe as outsiders and possible enemies. We are working with much greater forces but our social problems continue to be much the same, and it appears that our social thinking processes have changed very little.

A visit to the cliff dwellings of Mesa Verde National Park may illustrate the point for anyone caring to follow it through. Those Stone-Age apartment dwellers of 800 years ago had a social organization and system very little different from ours—it was apparent that the social thinking in their village groups of 50 to 200 was much the same as now. They had their Elks and Masons and Knights of Pythias as men’s clubs, with both religious and social aspects, and the women had their place for gossip at the community center. They were banded together for mutual protection from marauders, and for a certain amount of labor saving and specialization in agriculture and the building trades. They had an “ever-normal” granary for beans and corn. It is reported that they finally departed from the area because of a 23-year drought, showing that their politicians were unable to make rain—even as ours.

The important difference between their civilization and ours was not social, but scientific; add science, research, and their results to that Late Stone-Age civilization, and you have US.

However, it should be noted that the added quantity, technology, has been acting on civilization for less than 300 years, perhaps most of us would say for only 100 years. We take the human mechanism after 1,000-000 years of social living, add a new ingredient, and in 100 years expect that mechanism to adapt itself to the new conditions.

About one million years ago man learned to kill a fire. Ten thousand years ago he used the wheel. Five thousand years ago iron was discovered, three thousand years ago the alphabet was invented and we began to write, five hundred years ago movable type was invented and we could turn out mass propaganda and the comic book, and finally only 70 years ago we began to harness electricity!

The engineer and scientist have been severely criticized because we have failed to change the social attitudes of the human race, as an accompaniment to our technological advances, and the engineer’s best answer to such criticism may lie in the time scale, since man’s social reactions and attitudes have developed for a million years, and we are given one hundred years to change him! However, our scientific ability is such that I do not believe another million years will be required to make any meaningful change. In fact, good and noticeable progress may appear in the next thousand years.

As a result of this apparent disregard of the social implications of our scientific work, the engineering colleges have been severely criticized by engineers and noneengineers, and urged to include in their curricula more “humanistic-social” courses. This action is apparently based on the theory that with more humanistic-social courses the engineer becomes more socially conscious; he develops a new machine to lay brick, but only after considering the full and total social implications of forcing 500 bricklayers to find other jobs (where they will not be able to earn $2.75 an hour), and civilization will be much improved because the engineers are socially conscious! The doctors, lawyers, and sociologists should study a little more of their own field of human history to realize that mankind has not changed and will not be changed in any short period; and that as we go through one of the great change eras in the history of the world, the harnessing of the electrical forces, there abound to be dislocations and disruptions. This is not part of the old Industrial Revolution, this is a new Era of the Electron or of the Atom.

If study of humanistic-social subjects would make the engineer better able to converse and discuss social matters with the doctors and lawyers, that would be a gain, but it seems desirable that the doctors and lawyers come half-way and study a little science and engineering in order to discuss such subjects with us, and to understand better many of their own problems.

A very real trouble with engineering colleges is that they have alumni. These alumni graduated from college 10, 20, 30 or more years ago. Perhaps, as an example, a particular alumnus did not take accounting when he was in college. In his daily work since graduation he has found that it would have been advantageous to know the fundamentals of accounting. He immediately joins a chorus urging the addition of accounting courses, or economics, speech, history, psychology, or doorbell pushing to the engineering curricula. Does he look at a college catalog to see if our institution has more courses required? No, that is too much trouble and besides nothing ever changes at Old Siwash! When we went back for Homecoming last fall the ivy was still growing all over old Kempenphueher Hall, wasn’t it? And doesn’t Professor Hoffenflesser still have his morning coffee at exactly 10:02 every morning? Nothing changes!

Actually, of course, he would discover the Old Siwash engineering program had radically changed, and he would find already included courses in most of the subjects he now feels would be desirable. The difficulty with the move for more liberalization of engineering curricula is that the urge comes from men who base their criticism on what was taught twenty or more years ago. Our present curriculums may be good, but we will have to admit that our alumni tell us how bad they are twenty years hence!

In attempting to broaden the curricula, if we are to meet at least partially the criticism that has been directed at the engineering schools, it is extremely important to take
Ryder: What Kind of Engineers?

covered other materials besides amber which would attract light particles.

Then along came Ampere, Volta, Oersted, and finally Faraday, to discover the law of electromagnetic induction in 1831. In 1865 Maxwell told us that radio was possible, and Hertz proved in 1886 that Maxwell was right.

There are other fields in which the world has moved; in 1775 Washington took 12 days going by horse from New York to Boston to accept command of the Continental Army, and in 1963 astronauts landed on the moon. Bell had invented the telephone in 1875 and we now have telephone communication with all parts of the world. Edison, in 1875, used five 125-horse power machines at Pearl Street to light his electric lamp, and today we have 250,000 horsepower in a single machine.

Technical progress has been fast. Most of our electrical progress dates from Faraday in a total of 118 years. It must be remembered that each new development contributed to the total of knowledge, each further development was in turn built on a foundation of what had been achieved. At the unproven premise of those who would increase the number of years of a student in college: there is so much to learn that the time taken to learn it must be increased.

The increased accent on research is a second fundamental factor which must be considered in planning the college work of a young engineer. In order to develop an interest in our young graduate for such fundamental investigation, and to give him the basic tools with which to work, our engineering college courses need to provide considerable of the theoretical physical and mathematical bases of engineering.

The need is present because it can no longer be said that necessity is the mother of invention. Even if it were true, it seems a rather slipshod, unplanned, and unscientific way to get new devices invented. From an engineering standpoint it can be shown that very often invention is the mother of necessity, that one invention makes necessary a great many more inventions.

To refuse the old saying, consider the invention of the first incandescent lamp by De La Rue in 1820. It used a platinum filament in a semi-evacuated space, much like the Edison lamp of 1900. It was not needed or necessary, in fact, the world was not yet ready for it. Using wet batteries as the only available electric power of the day, it was not practical or economical. The invention of the electric motor by Davenport in 1834 is another example of an invention ahead of the need, there being as yet no great mechanization of industry, no machines waiting to be driven, and the only electric power, that from batteries. Fifty years later, after invention of the incandescent lamp and the electric motor became practical and needed. The invention of the scanning disk by Nikprow was hardly a necessary invention in its day, but it made necessary other inventions that have followed and must still be made, to make television fully successful.

In fact, it is the fundamental scientific progress of a continuous nature the basic invention must invariably be ahead of the need. Our present-day pure research is all of that nature, and more and more of our engineers are going into that work or into closely allied lines. Our colleges have not always trained engineers with that type of work as their intended vocation. It can be suspected that the goal of a great many of our engineering courses of 25 years ago was to train a man to the calibre of a drafting board, in a design department where he could then learn to be an engineer, or to eventually place him as a salesman of electrical products. A rather superficial form of electrical engineering teaching had grown up at that time around the one-frequency idea of 60 cycles only. This superficiality was so firmly rooted, so few real fundamentals were taught or stressed, that the resulting graduate was not too satisfactory in really technical positions. In fact, the roots were so firm and thick that the tree could not be seen and when radio made its entrance in the late 1920's electrical engineers were accused of over-specializing in the "narrow" field of radio.

It can now be seen in retrospect that the opposite was true, as radio and electronics have led to a return to true fundamentals, and the 60-cycle emphasis was an act of the treatment of electrical engineering in a very narrow and specialized field. As a result, most men who came through our college electrical engineering departments, in the 60-cycle days, were not fully fitted for work under our present-day emphasis on research or in fundamentals. It is frequently said that you can make a power man out of a radio man, but not the reverse. The implications of that statement should be carefully thought out.

Why were the men not given more fundamental college work? The answer can be found in certain statistics, first, on the growth of research in industry in the United States as seen in Table I. In 1920, research was very small business in the United States, but it is not today, when it represents a $750,000,000 business annually. In 1920 industry did not demand training of graduates to fit them for research, but during the last two decades the picture has changed. Additional statistics, from a survey by E.C.P.D., support the argument that the present demand is for better fundamental training than was formerly given. This information is presented in Table II, and does not fully include the radio engineering group which would probably raise further the higher percentages. Other comparative figures in Table III further illustrate the trend and show the tendency toward more advanced degrees—these degrees being in almost all cases indicative of highly technical work.

Another factor which has increased the demand for higher technical excellence of our engineering graduates is that some of our engineering fields are approaching a point of increased difficulty. That is, they have
The first clock was probably the simple rising and the setting of the sun, a mere yes-no sort of accuracy. There are evidences that crude forms of sundials existed about 6,000 years ago for interpolating the yes-no answers. These were of use only when the sun shone and were, of course, in error at different times of the year, unless properly installed. Thirty-five hundred years ago there came into use water clocks of many designs. These involved the realization that a leaky bucket, always originally filled to the same level, took practically uniform increments of time to empty. If a slave were added to refill the bucket each time it became empty, and mark the fact by ringing a gong, there appears the stone-age Swiss movement. Someone then realized that fluids other than water might be used, and in the desert where they had little water, sand was used, the result being the common egg-timer or sand glass. Burning candles or oil lamps were also forms of hour markers and such devices were the best to be had, with an ultimate accuracy of 5 per cent, if the slave stayed awake. Such a condition persisted until 1360.

In that year the escapement was invented, using power derived from falling weights. Galileo described the pendulum in 1578, but since he spent a great part of his time in jail, he was not enough interested in the measurement of time to build the pendulum into a clock, and left that for Huygens to do in 1657. The pendulum was good, so good that after much development it is still used in our very best clocks. It has been superseded for laboratory use by the quartz crystal clock, and in fact the quartz crystal has been used to show up previously undiscovered variations in the motion of pendulum clocks. These are due to the daily lunar cycle or tidal effects and amount to a few tenths of a millisecond.

Now to summarize the point. The clock has been under development for 6,000 years. The present best clock is an electrical or electronic device, and of interest to radio engineers. What sort of limits are we facing in attempts to develop further improvements in the measurement of time? The sand glass was accurate to 5 per cent, or an error of about one second in every twenty. The escapement at the limit of its development had reduced this error to one second in 8 days. The modern best pendulum chronometer is accurate to one second in two years. The quartz crystal clock may have an error of one second in fifteen years. All this while the earth itself is only consistent to one second in three years. Certainly in the measurement of time, as in many other fields, we are at the point where each new step forward becomes harder to take and entails a higher level of excellence in our work, but good engineers prefer it that way.

Now what have the colleges done? How have they reconciled the conflicting demands on a student’s time from those wishing to increase the humanistic-social content of the courses, from those insisting that the student know “more,” and from those with demands for a higher technical level of accomplishment? Various schools have come to different answers. Certain colleges, only a few, have gone to so-called “five-year plans,” to gain time. Others are thinking about the matter, but it can readily be said there will be no stampede.

It appears possible to state a philosophy which might be followed in the development of college engineering curricula. Simply stated, this could be: To meet modern needs, the engineering college should teach fundamentals only, and should teach those fundamentals thoroughly and well. As long as we remain assured that it is also possible to state the belief: If the fundamentals are well taught, then the applications encountered in professional work will follow easily.

Such a program is in line with the desires of certain industries which prefer to do further teaching in the application field in their own programs. Of course, there is room for discussion here as to where the dividing line comes between fundamentals and applications, but experience seems to indicate that the colleges usually are the better judges of this matter.

To carry out such a philosophy it is desirable that existing curricula be scrutinized closely. All curricula develop fringes, and dark corners with musty odors. Some dean, feeling that the course which he formerly taught was very good and extremely important, orders it included for all students. Then the dean becomes busy and no longer teaches the course, the teaching level and interest drops, the course deteriorates. The dean moves on, but the course remains. Other courses are attached to the curriculum because at one time they were thought important—electrical design, electric traction, telephones—to mention a few—and they carry on long past their prime. These should be carefully examined to determine if they are really fundamental. One enlightening question which may be asked concerning any such course is: How many men in a given graduating class will use that material in their future work? Then it is advisable to ask: Can a course be justified for that number?

Wartime curricula and teaching gave engineering teachers an excellent opportunity for research in details of curricular problems if advantage were taken of the situation thrust upon the colleges, and if questions were asked and studies were made. From such experience in research covering several such programs at the collegiate level, two basic principles were found to lead to more efficient use of a student’s time, namely:

1. Four, but never more than five, major courses in any one quarter or semester.
2. Courses should be given in concentrated doses, six credit hours in one semester being more efficient than three hours in each of two semesters.

In one of the wartime programs, students were assigned seven or eight courses in a given semester, resulting in a scattering of student interests and failure to do good work. Courses which are spread too thinly, meeting possibly twice a week, lose much of the value of a lecture and class discussion occurring on one day, because of the time elapsing before the student takes the book out of the corner to study for the second meeting of the week. By daily meetings, today’s class discussion and tomorrow’s study are brought together.

By dropping nonessential and outdated courses, sufficient time may be found to include a number of humanistic-social courses such as additional English, psychology, history, economics, and biology. At the same time, because of the increased efficiency obtained through teaching courses in concentrated doses, it should be possible to teach more technical fundamental material in the same number of technical hours.

Graduates are now better paid than when they leave college than was true twenty years ago, and they are worth more. It follows that their time in college is more valuable. Thus a college must increase the efficiency with which it utilizes a student’s time in accordance with the increased value of this time, or it is falling in its duty to the student and to those who pay the college bills. Increases in utilization efficiency will require careful investigation by the faculty of their methods, and of the course contents, and this should stimulate the faculty along the line of wholesome advancement.

It has been said that a college is a collection of buildings with ivy creeping around on the outside and with the faculty creeping around on the inside. In a modern engineering school that cannot be true.
Electronics Applied to the Betatron*  
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Summary—There are several phases of operation of the betatron, used as a source of X rays, which require the application of electronic circuits. This paper reviews the problems of electron injection, electron ejection, and X-ray monitoring. It reviews a set of electronic circuits which have been found satisfactory in solving these problems.

I. INTRODUCTION

It is the purpose of this paper to review the electronic circuit components of the betatron, or induction electron accelerator, used as a source of X radiation.

The theory and development of the betatron have been discussed by others, and its use as an X-ray equipment described. These will not be treated in detail here.

The betatron accelerates electrons around a circular path or equilibrium orbit by means of magnetic induction. The electrons are accelerated by a radially symmetrical time-varying magnetic flux existing between two poles of a magnet. The space distribution of the flux, as determined by the profile of the magnet poles, is such that the electrons are both accelerated by the potential gradient due to the time variation of the flux, and constrained to move in a circular orbit of constant radius by the flux's force on them due to their motion. That portion of the flux which links the equilibrium orbit is called the accelerating flux, and that in the immediate vicinity of the orbit, the guiding flux. The relation between these fluxes has been shown to be

$$\Delta \phi = 2\pi r \Delta B, \quad (1)$$

in which

$$\Delta \phi = \text{change in accelerating flux}$$

$$\Delta B = \text{change in flux density at the equilibrium orbit of radius}\ r,$$

regardless of the shape of the flux versus time wave. Therefore, for constant orbit radius, the proportionality of \(\Delta \phi\) and \(\Delta B\) must remain constant. The time variation of the flux is usually sinusoidal, and the electrons are accelerated to their maximum energy once each cycle during a certain fraction of the increasing half-cycle of the flux. The flux may have a steady or dc component, in which case the betatron is said to be biased.\(^*\)

\(^*\) Decimal classification: 621.352.5.623. Original manuscript received by the Institute, June 3, 1949. Tulane University, New Orleans, Louisiana, and General Electric Company, Schenectady, N. Y.


II. ELECTRON INJECTION EQUIPMENT

The components in this group are:  
a. Electron source, or gun  
b. Gun filament supply  
c. Gun accelerating voltage supply (Injection pulse generator, pulse transformer, and crest voltmeter)  
d. Timing-pulse control circuit.

These act to inject electrons into the equilibrium orbit each cycle near the time the guiding flux passes through zero in an increasing direction (at A in Fig. 2). A timing reference is obtained from a "peaker strip" placed across the gap between the magnet pole pieces. The peaker strip is a laminated permalloy bar of small cross section on which a coil is closely wound. It is saturated by the magnetomotive force across the gap, except for a short interval of time near guiding flux zero. During this short interval, the rapid flux reversal in the peaker strip as it unsaturates and saturates in the opposite direction induces a voltage pulse of short duration in the coil. Two such pulses are produced for each cycle of guiding flux and, since they are of opposite sign, the one produced at the zero-increasing guiding flux point can be selected. This pulse is applied to the timing-pulse control circuit, which then generates a triggering pulse delayed the proper interval after guiding flux zero. This triggering pulse fires the injection pulse generator which, acting through the pulse transformer, applies the high accelerating injection voltage pulse to the elements of the electron gun.

Fig. 1—Block diagram of biased open-core betatron.

The electrons are accelerated in a toroidal evacuated tube and, after reaching a desired energy, are caused to deflect or shift from their equilibrium orbit to strike a metal target. The rapid deceleration the electrons undergo as they strike the target gives rise to a fairly sharply defined X-ray beam projected in the direction the electrons were traveling at the instant they struck the target.

There are three phases of betatron operation which require the use of specialized electronic circuits.

Referring to Fig. 1, a block diagram showing a biased open-core betatron, these three phases of betatron operation are:  
(1) those concerned with injecting electrons into the equilibrium orbit;  
(2) those concerned with deflecting or ejecting electrons from the equilibrium orbit;  and  
(3) those concerned with monitoring the resulting X-ray output.

Fig. 2—Guiding flux density in a biased betatron.
In practice, the anode of the gun is connected to ground and the filamentary cathode is pulsed negatively. To supply the heating power for the filament, the secondary of the pulse transformer consists of a bifilar winding, the low potential end of which is connected to the filament supply and the high potential end through a step-down transformer to the filament.

There are also provided a voltage-regulated gun filament supply and an injection crest voltmeter for measuring the negative pulse supplied from the pulse transformer to the gun filament.

The relation between guiding flux density and the energy of an electron in an equilibrium orbit of given radius determines fundamentally the operating constants of the electron injection components. This relation may be written

\[ B_r = \frac{10^4}{3} \sqrt{V(V + 1.02)}, \]  

in which

- \( B \) = guiding flux density at equilibrium orbit in gauss
- \( r \) = equilibrium orbit radius in centimeters
- \( V \) = electron energy in million electron volts (MeV).

As shown in Fig. 2 for a biased betatron, \( B \) comprises a sinusoidal component of peak value \( B_s \) and a steady component \( B_{dc} \) of magnitude less than that of \( B \). From the figure, it can be written that

\[ B = B_s [\sin \delta + \sin (\omega t - \theta)]. \]

Substitution of (3) into (2) yields upon solving for \( V \)

\[ V = [0.264 - (3B_r)^2 [\sin \delta + \sin (\omega t - \theta)] 10^{-3}]^{1/2} - 0.51, \]

an expression which correlates energy of the electron in the equilibrium orbit, radius of the equilibrium orbit, guiding flux density, and time after guiding flux zero.

In order to enter and remain in the equilibrium orbit, electrons injected a time \( t \) after guiding flux zero must have the energy (corresponding to the flux at that instant) given by this expression. This energy is referred to as the "matching voltage." If from then on (1) is satisfied, the electrons will continue in the equilibrium orbit.

It has been found, in practice, that a good value for \( \delta \) is \( \pi/3 \). Fig. 3 shows a plot of (4) with \( \delta = \pi/3, \omega = 1.130 \) (for a 180-cycle supply), and \( B_{max} = 1.68 \times 10^3 \) gauss-centimeters (which corresponds to a maximum electron energy of 50 Mev).

The following points may be observed:

1. If excessive voltages on the electron gun are to be avoided, electron injection must be completed within a few microseconds after guiding flux passes through zero.

2. At a given instant only electrons of a certain energy match the equilibrium conditions exactly. For ideal matching conditions, the energy of the electrons coming from the gun should increase with time during the injection pulse.

To these may be added the following:

3. Because there exist restoring forces on electrons which are out of the equilibrium orbit, there is a small range of energy about the ideal value given by (4) which an injected electron may have at a given time after guiding flux zero and still match the equilibrium orbit. This range is of the order of \( \pm 1 \) per cent of the ideal value given by (4) and as the magnitude of the restoring forces increases with \( t \), it is larger for large values of \( t \) and smaller for small values. This puts definite limitations on the amount of jitter which can be present in the injecting pulse.

4. It is desirable to capture in the equilibrium orbit as many electrons as possible with a minimum of "lost" electrons, since these lost electrons increase space charge and upon hitting the tube walls cause overheating.

Keeping these observations in mind, the properties of the components of the electron injection equipment can be discussed.
A. Electron Gun

A 50-Mev betatron tube and electron gun are shown in Fig. 4. In order to achieve high gun emission and high percentage of electron capture in the equilibrium orbit, injection voltages should be high. This requirement has to be balanced against the spacing of the gun elements which is limited by the space available for the gun structure. For guns of size commensurate with available space, the maximum accelerating voltage pulse which can be applied to gun elements is around 70 kv peak. Reduction of the number of stray electrons hitting the tube wall can be helped in design of the gun by shaping the elements to produce a ribbon-like stream of electrons tangent to the equilibrium orbit. To avoid the necessity of close alignment of the gun structure and to insure tangency after it is mounted in the tube, a deflecting electrode is incorporated into the gun. In the gun shown, the filament and focusing electrode are connected together and pulsed negative with respect to the anode, while an adjustable proportional positive pulse is simultaneously applied to the deflector electrode. To prevent stray electrons from accumulating on the tube inside wall and building up a disturbing electrostatic field, the wall is covered with a grounded conducting coating of sufficiently high resistance to prevent excessive heating from the eddy currents induced by the alternating guiding flux.

The gun filament must be reasonably long-lived, capable of supplying about 2 amperes peak and 2 milliamperes average emission, and must be mechanically strong enough to withstand rather severe vibration.

B. Gun Filament Supply

For steady X-ray output of the betatron, it is necessary that average gun emission be held quite closely to a constant value. This can be achieved by regulating the voltage of the gun filament supply, which, because the filament power is carried through the pulse transformer, is most conveniently an ac supply allowing use of a step-down transformer at the high end of the pulse transformer to keep the current through its bifilar winding low. The magnetic field resulting from the current flowing in the filament affects the output of the gun, and if the supply for the filament is not of the same frequency as the magnet flux, modulation of the X-ray output will result. Since there are not presently available non-electronic constant voltage supplies for frequencies other than 60 cps, an electronically regulated gun filament supply is required.

In Fig. 5 is shown the circuit employed for the gun filament supply. It is an adaptation of a circuit described by Ridenour and Lampson. The primary of transformer T1 is supplied through capacitor C which has an appreciable impedance at the supply frequency. Diode V1 is operated emission limited such that anode current varies considerably with small changes in filament voltage. The anode of V1 is connected to the grid of V2, whose cathode is at a slightly higher positive potential. The anode of V2 is connected to the grids of the push-pull tubes V3 and V4 whose cathodes also are connected to a source of higher positive potential than that of their grids. The current drawn by these tubes through secondary S2, together with that drawn by the gun filament from secondary S3, cause a reduction in transformer voltage, and a rise in the impedance drop through capacitor C. As supply voltage increases the increased emission in V1 causes considerable drop in its anode potential, which is amplified by V2 to produce a rise in grid potential of V3 and V4. The resulting increase in their anode current loads the transformer more heavily, and the increased impedance drop through C maintains the transformer voltages essentially constant.

Variable ratio auto transformer T2 allows adjustment of the gun filament voltage to any desired proportion of the regulated output voltage.

C. Gun Accelerating Voltage Supply

Electrons will be captured in the equilibrium orbit only if they have at the instant of injection energies within about 1 per cent of that given by (4). Thus, recalling Fig. 3, the ideal wave shape of the pulse of injection voltage should be as shown in Fig. 6(a). It is not practically possible to obtain a pulse of exactly this shape; however, it has been found that by generating a square-wave pulse and applying it to the gun through a high turns-ratio pulse transformer, the resulting distorted secondary voltage pulse (which is roughly

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A typical injection pulse generator is shown in Fig. 7. It consists of a pair of 872 phantrons 1' and 1'2 connected in a voltage-doubling circuit to charge a 3-microsecond, 50-ohm lumped-constant line to 8,000 volts. When the line is discharged through the 5C22 hydrogen thyratron 1'3 into a 50-ohm resistive load, a square wave of voltage of 4,000 volts and 3 microseconds duration is produced. A commutating voltage for the thyratron is provided by a tap on the supply transformer T1.

The 5C22 thyratron is chosen because of its high peak current rating and its short deionization time, allowing its use with betatrons operating at frequencies as high as 1,920 cps.

The pulse from the injection pulse generator is fed, as shown in Fig. 1, to a pulse transformer having a turns-ratio of 17 1 to 1. The secondary is loaded by the electron gun and its voltage divider to produce an equivalent load referred to the primary of 30 ohms. Ideally, the resulting 20-kv voltage appearing across the secondary terminals would be a square wave, but design limitations and unavoidable stray load capacitances produce the fortunately more desirable wave shape shown in Fig. 6(b).

The gun voltage divider is tapped at the proper points for supplying the gun elements and at 50 ohms from ground for supplying a low proportional voltage used in observing and measuring gun voltage with the injection crest voltmeter. This latter uses well-known circuits for measuring crest voltages, employing a diode rectifier which charges a capacitor to the crest voltage and a two-tube bridge voltmeter for measuring the capacitor voltage. An over-current relay is included in the output line to protect against arc-over of the gun.

According to (4), matching voltage is a function of guiding flux density; hence, in order to maintain correspondence between injection voltage and the matching voltage required by (4), the supply voltage for the injection pulse generator must not differ substantially from that supplying magnet excitation. In some circuits, such as the one shown in Fig. 1, the magnet exciting voltage is not constant and does not vary in the same manner as that of the magnet power supply. Therefore, best results are obtained by supplying the injection pulse generator from a source of voltage which is proportional to magnet flux. Such an auxiliary voltage supply may be had from several turns placed around a convenient section of the magnet core.

D. Timing-Pulse Control Circuit

The essential function of the timing-pulse control circuit is to control the time phase of the gun acceleration voltage pulse with respect to magnet guiding flux zero. Three functions are required: (1) manual phase control to adjust initially the gun acceleration voltage pulse to the matching voltage; (2) automatic phase control to adjust for random variations of matching voltage; and (3) suppression of jitter in the gun acceleration voltage. These are provided by a circuit such as that shown in Fig. 8.

Manual phase control is achieved by the piper and variable delay circuits, the details of which are given in Figs. 9 and 10. The piper strip is biased by means of current flow resulting from the two sources of voltage as shown, such that its flux passes through zero before the magnet guiding flux. The voltage pulses from the peaker strip are applied to the piper, Fig. 9, the input rectifiers of which are arranged to prevent a net direct current from being drawn through the peaker strip and to allow only negative pulses, Fig. 9(a), to reach CI.

This capacitor, acting with the grid resistor R3 of tube 1'1, causes a voltage which is the derivative of the negative peaker pulse to appear at the grid of that tube (see Fig. 9(b)), 1'1 is a class-A amplifier which applies the amplified pulse shown by Fig. 9(c) to the grid of 1'2. The grid of this tube is over-driven so that the amplified pulse appears on the anode as shown by Fig. 9(d), which in turn is differentiated as shown in Fig. 9(e) by means of a small capacitor C3 and grid resistor R3. 1'3 is biased well below cutoff, so that only the narrow top of the differentiated wave causes plate current flow, and the resulting plate voltage pulse is as shown by Fig. 9(f). This pulse is applied to the positively biased 1'4 and is of sufficient value to drive this tube well below cutoff, producing the square topped wave shown by Fig. 9(g). 1'4 is a 6AG7, chosen for its high transconductance, which allows a small plate load resistor to be used. This keeps the time constant of the plate circuit low permitting a rapid rise of voltage upon cutoff which produces the desired steep wave front.

Refer now to the variable delay circuit, Fig. 10. In the normal condition, 1'2 is conducting because of its positive bias and 1'1 is held nonconducting by the negative bias developed across the common cathode resistor R4. The square-topped pulse from the
Pipper is applied to the input as shown in Fig. 10(a). This drives the grid of \( V_1 \) positive, causing current to flow in the anode load potentiometer \( P_1 \) and resulting in a fall of anode potential as shown by Fig. 10(b). Since the voltage across capacitor \( C_4 \) cannot change instantly, the grid of \( V_2 \) is driven negative as shown by Fig. 10(c) and the potential of its anode rises to supply voltage as shown by Fig. 10(d). The grid of \( V_1 \) is then held positive by voltage divider action between \( R_14 \) and the grid resistor \( R_2 \). The voltage on \( C_4 \) will slowly decay as shown by Fig. 10(c), as capacitor \( C_4 \) discharges through resistor \( R_8 \), allowing the grid voltage of \( V_2 \) to rise to cutoff, at which point this tube starts to conduct. The fall of the anode voltage of \( V_2 \) causes a fall of voltage on the grid of \( V_1 \), causing a rise in the anode voltage of \( V_1 \), which assists in the rise of grid voltage of \( V_2 \). This results in a very rapid return to the original condition. The rapid fall of anode potential of \( V_2 \) is differentiated by the small coupling capacitor \( C_5 \) and grid resistor \( R_5 \), producing a pulse at the end of the timing cycle on the grid of \( V_3 \) as shown by Fig. 10(e). Since the grid is driven well below cutoff, the resulting pulse across the anode resistor \( R_{13} \) is square-topped as shown by Fig. 10(f). This pulse is applied to the grid of \( V_1 \), a 657T tube whose two sections are connected in parallel and biased to cutoff, and a 50-volt, 1-ampere pulse appears across the 50-ohm load at the end of a 50-ohm coaxial line in the cathode circuit. It is this pulse which fires the injection pulse generator. Manual control of the time delay between the input pulse and output pulses is determined by the setting of the potentiometer \( P_1 \) in the anode of \( V_1 \). This determines how far negative the grid of \( V_2 \) is driven, which in turn determines the time required to rise to cutoff. The minimum time delay obtainable is equal to the width of the triggering pulse, and can readily be made only a few microseconds. In practice, the peaker strip bias is usually set ahead to allow full range of manual adjustment of the variable delay circuit. The variable delay circuit is then adjusted until gun accelerating voltage coincides sufficiently with the required matching voltage.

Automatic phase control is required because supplying the injection pulse generator from a source whose voltage varies linearly with the excitation voltage of the magnet does not eliminate entirely deviation of the gun accelerating voltage pulse from the value required for correspondence with the matching voltage. It may be observed from (2) that, for a given short time after guiding flux zero, matching voltage varies as the square of magnet excitation voltage. Thus if magnet excitation voltage increases slightly, the gun accelerating voltage should increase as its square, but, since the injection pulse generator supply voltage varies linearly with magnet excitation voltage, only a linear increase occurs. This means that gun accelerating voltage is too low. To correct for it, the time of injection must be advanced.

This is accomplished, as shown in Fig. 8, by that portion of the peaker strip bias current derived from the magnet voltage. As magnet voltage increases, more bias current is applied to the peaker strip, causing its flux to go through zero and produces the reference voltage pulse at an earlier time. The value of voltage supplied is such that the rate of change of voltage produces the correct rate of change in timing for the particular injection voltage being used. This voltage would normally result in an appreciable current through the peaker strip which would overheat the windings. Therefore, since it is only the change in current which is required for compensation, the initial current can be balanced out by a fixed opposing potential from a regulated source as shown. In the peaker bias circuit, the rf choke prevents the high-frequency peak from being shunted out. The resistor \( R \) (Fig. 8) is sufficiently large that the current flow from the ac bias supply is in phase with the voltage, such that its maximum value occurs at guiding flux zero when it is needed.

In order to suppress the jitter which appears in the gun acceleration voltage, it is necessary to eliminate the jitter at its source. One of the main sources of jitter may be due to the tolerance in input voltage which will fire the injection pulse generator thyatron. Consequently, for a given input pulse there will be a corresponding variation in the time at which the thyatron will fire. To minimize this time variation, it is necessary to use a large pulse of short build-up time to trigger the thyatron. Such a pulse is provided in the variable delay circuit by use of the high transconductance tubes \( V_1, V_2, V_3 \) in Fig. 10) working into low load resistors. Another source of jitter is pickup, which, if not suppressed, may trigger the thyatron before the desired time. Pickup can be minimized by careful shielding and the use of coaxial cables. A final source of jitter is variations in the timing circuit. This is eliminated by the use of high-quality components for \( R_8 \) and \( C_4 \) in the timing circuit (Fig. 10), and by using regulated voltage supplies.

### III. Electron Ejection Equipment

These circuits act to shift or eject the electrons from their equilibrium orbit and cause them to strike the metal target. As may be observed from (1), for electrons to remain in an orbit of constant radius during acceleration, the rate of change of accelerating flux with respect to guiding flux \( \left( \frac{\Delta \Phi}{\Delta \Psi} \right) \) must be constant. A deviation from this condition will cause a change in orbit radius; therefore, when it is desired to expand or contract the orbit to cause the electrons to strike the target, it is merely necessary to change one of either guiding or accelerating flux without changing the other. In one form of the betatron, the core is so designed that its central portion saturates at the crest of the accelerating flux wave. This results in a deficiency of \( \Delta \Phi \), causing the orbit radius to reduce to maintain equilibrium, and the electrons spiral inward to strike the target. This simplified design produces X rays of only one maximum energy. A more flexible design incorporates "orbit-shift" coils mounted on the magnet.
pole faces in the vicinity of the equilibrium orbit such that when current is passed through them, the guiding flux is strengthened or weakened and the orbit radius contracts or expands until the electrons hit the target. Current may thus be passed through these coils in a short pulse at any time after injection causing the electrons to strike the target with the production of X rays of any desired maximum energy up to the maximum for which the betatron is designed. Such a system is indicated in Fig. 1. The components in this group, designed to contract the orbit, are:

a. Orbit-shift coils
b. Orbit-shift coil supply (ejection pulse generator and peak current indicator).

d. Orbit-Shift Coils: The orbit-shift coils are mounted on the upper and lower magnet pole faces as shown in Fig. 1. At a point somewhere between guiding flux zero and guiding flux maximum, such as C in Fig. 2, a pulse of current from the ejection pulse generator is sent through the coils in such a direction as to strengthen the guiding flux in the space between them.

The increase in guiding flux so obtained decreases the radius of the equilibrium orbit, and the electrons spiral inward striking the target with essentially the energy they had at C. Control of the time delay between guiding flux zero and the firing of the ejection pulse generator gives control of the location of point C, or, in other words, control of the maximum energy of the X-rays produced by the betatron.

Design of the orbit-shift equipment is based on the positioning of the orbit-shift coils and the change in equilibrium orbit radius required to cause the electrons to strike the target. Assuming that the electron orbit is contracted rapidly without significant change in electron energy from one of radius \( r_0 \) to one of radius \( r_f \), corresponding to the location of the target, the guiding flux density \( B_f' \) at the target at the end of the orbit-shift pulse must be

\[
B_f' = B_0 \frac{r_f}{r_f} \quad (5)
\]

in which \( B_0 \) is the guiding flux density at the equilibrium orbit of radius \( r_0 \) at the beginning of the orbit-shift pulse.

In a betatron, the guiding flux density is a function of radial distance from the magnet pole piece center such that when no orbit-shift pulse is present, the guiding flux density \( B_1 \) at the radius \( r_f \) is

\[
B_1 = B_0 \frac{r_f}{r_f} \quad (6)
\]

in which \( n \) is a number between zero and one, in present designs usually \( 3/4 \). Subtracting (6) from (5), the change \( \Delta B_1 \) in guiding flux density required to shift the orbit from one of radius \( r_0 \) to one of radius \( r_f \) is:

\[
\Delta B_1 = B_0 (A - A^n) \quad (7)
\]

in which \( A \) is the ratio \( r_0/r_f \).

In order to shift the orbit from one of radius \( r_0 \) to one of radius \( r_f \), it is necessary to change the flux through the entire area between \( r_0 \) and \( r_f \). The coil configuration shown in Fig. 1 does this. For design purposes, it can be assumed that the orbit-shifting flux appears only in the volume between the upper and lower coils, and that the reluctance of the flux path in the magnet iron is negligibly small compared to that in air. Under these assumptions, the peak current required to shift the orbit is then

\[
I_p = 0.795B_0 \cdot d \quad (8)
\]

in which

\[ I_p = \text{peak current in amperes} \]
\[ \Delta B_1 = \text{orbit-shift flux pulse in gauss required by equation (7)} \]
\[ d = \text{average distance in centimeters between upper and lower orbit-shift coils} \]
\[ n = \text{total number of orbit-shift coil turns} \]

The inductance \( L \) of the shift-coil configuration can be calculated or measured, and the peak energy \( W_p \) required by the orbit-shifting flux pulse calculated from

\[
W_p = \frac{1}{2}LI^2 \quad \text{joules.} \quad (9)
\]

b. Orbit-Shift Coil Supply: A relatively simple way to obtain the flux pulse is to discharge a capacitor through the orbit-shift coils, the capacitor energy storage required being equal to that given by (9), or,

\[
\frac{1}{2}CV^2 = \frac{1}{2}LI^2. \quad (10)
\]

The value of capacitance chosen should be such that the time of rise of the current through the orbit-shift coils is small, compared to that of a cycle of the guiding flux. The practical limit is that of insulation for the higher voltages required to achieve shorter rise times.

A typical ejection pulse generator for supplying the orbit-shift pulse to a betatron is shown in Fig. 11.

Power for the orbit-shift coils is supplied through transformer \( T1 \) from turns on the magnet core. Capacitor \( C1 \) is charged through phosphotron \( V1 \) to an energy given by (9) on the half-cycle of voltage preceding that during which the electrons are accelerated. When thyatron \( V2 \) is fired, \( C1 \) is discharged through the orbit-shift coils. \( V2 \) must be able to supply a peak current given by (8); further, it must have a deionization time which is short enough to enable the point of firing \( C \) in Fig. 2 to be moved close to guiding flux maximum, obtaining maximum energy ejected electrons while still allowing \( V2 \) to be deionized before \( V1 \) begins to conduct on the following charging half-cycle.

Bias for the thyatron is obtained from rectification of the voltage from one winding of \( T2 \) and the firing pulse from peaking transformer \( T3 \). \( T3 \) is fed from the phase-shifting network comprising \( R1 \) and \( C2 \), which is supplied from a second winding of \( T2 \). Since \( T2 \) is supplied from magnet voltage, this network determines the point \( C \) in Fig. 2, at which the ejection pulse occurs and hence the maximum energy of the X-ray output.

A voltage pulse for use in the megavoltmeter (to be discussed in the next section), and a peak current indicator is obtained from a current transformer \( T4 \) placed in series with the cathode of \( V2 \). The current transformer is loaded with a resistor \( R2 \) of sufficient value to produce a proportional voltage within its saturation rating. This voltage is applied to a crest voltmeter similar to that previously described, and serves to monitor the orbit-shift coil current.

---

**Figure 11**: Ejection pulse generator for a betatron.
IV. X-Ray Monitoring Equipment

There are two components of monitoring equipment necessary:

a. X-ray energy monitor (megavoltmeter)
b. X-ray intensity monitor (ionization chamber and associated amplifiers).

1. X-Ray Energy Monitor: As has been discussed previously, changing the phase relative to guiding flux zero of the orbit-shifting current pulse changes the energy with which the electrons strike the target, and hence changes the maximum energy of the X-rays produced by the betatron. It is necessary in many applications to know what the maximum X-ray energy is, and desirable to have a more or less direct way of measuring it, rather than depending on an indirect method such as calculating the position of phase shifting control Ri in the ejection pulse generator (Fig. 11).

As electron injection always occurs at essentially zero guiding flux density (A, Fig. 9), equation (1) may be rewritten

\[ \Delta \phi = (2\pi)Br. \]  

(11)

can be seen from (11) that the product Br is proportional to the total change in accelerating flux over the time of acceleration; that is:

\[ Br = \int_{t_1}^{t_2} \Delta \phi dt. \]  

(12)

Since the instantaneous voltage E induced around a loop enclosing the accelerating flux is proportional to the rate of change of that flux, it may be written

\[ E = \frac{\Delta \phi}{dt}. \]  

(13)

Substituting (14) into (13):

\[ Br = \int_{t_1}^{t_2} E dt. \]  

(15)

The voltage E may be integrated over the time of acceleration (A to C on Fig. 2) to obtain a quantity proportional to the value of Br at the instant of ejection. From (2) the corresponding value of electron energy V can then be obtained.

A circuit (called a megavoltmeter) to accomplish the integration of E over the time of acceleration has been described, and is shown in Fig. 12. A voltage proportional to that induced within the equilibrium orbit (E in (14)) is obtained from the auxiliary winding around the magnet core (Fig. 1). This is applied to thyatron V1, the anode of which is biased positively by an amount equal to the tube drop of approximately 15 volts. Thus, instantaneous current flow through the tube is proportional to E divided by the series resistance R1 plus R5.

At guiding flux zero, the thyatron V1 is fired by the peaker strip voltage pulse and conduction starts. At the instant of ejection, a pulse of voltage from the orbit-shifter circuit fires thyatron V2 whose cathode is connected to a dc potential more negative than the cathode of V1. This lowers the anode of V1 below its cathode potential causing the tube to fire. The ammeter A, in series with the anode of V1, measures the average current through that thyatron. Since this current is proportional to the integrated value of E, it is also proportional to the product Br. Knowing the constants of the betatron, it is then possible to substitute in (2) and obtain corresponding values of V for each value of Br, and thus mark the scale of ammeter A to be direct reading in maximum X-ray energy. The variable resistor R5 is provided to calibrate the ammeter initially to agree with the scale markings.

b. X-Ray Intensity Monitor: The unit used in measuring the amount of X radiation passing by a point in space is called the roentgen (after the discoverer of X-rays) and is symbolized by the letter R. It is defined by the ionization produced in a unit volume of air by the radiation at the point in question. Numerically, it is that quantity of radiation which will produce ions bearing one electrostatic unit of charge, of one sign, per cubic centimeter of dry air at standard temperature and pressure. In the practical system of units, then

\[ 1R = 3.33 \times 10^{-18} \text{ coulombs/cc}. \]  

(16)

The unit most commonly used in measuring intensity of X radiation is the roentgen per minute. Thus, X-ray intensity at a point may be measured by measuring the charge per minute produced in a unit volume of air at that point.

A practical form of intensity monitor consists of a thin walled metallic chamber filled with air in which is placed an insulated electrode connected to a source of direct potential. When an X-ray beam passes through the chamber, the resulting ionization allows current to flow between the electrode and the chamber walls. This current is a measure of the time of ionization produced in the chamber and hence, with the volume of the chamber known, is a direct measure of the average intensity of X radiation through it.

There are several factors dictating the design of an "ionization chamber" which have been discussed elsewhere. Since the movement of ions through the air to the electrode is relatively slow (of the order of 1 centimeter per second per volt per centimeter gradient), it is desirable to have the spacings small and the potentials high to speed up the collection of ions, and thus minimize the occurrence of recombinations which will reduce the collected current. On the other hand, if the potential gradient is too high, the velocity of the simultaneously produced electrons becomes sufficiently great to cause ionization of additional gas molecules, causing an increase in collected current which can lead to a runaway condition once ionization is started. Therefore, in the design of an ionization chamber for direct reading of the relatively intense radiations of a betatron, the spacings are kept small in order that high speed of collection may be attained without excessive electron velocities.

One form of ionization chamber consists of three parallel circular plates spaced a small distance apart and placed directly in and perpendicular to the X-ray beam (see Fig. 13). These plates are made of a thin lightweight material, such as aluminum, and offer small absorption to the beam. The outer plates are connected to a source of potential of the order of 1,000 volts and the
central plate through a high resistance to ground. The potential drop in this resistor caused by the current flow due to the ions produced by the X-ray beam is measured on a tube voltmeter, and the indication of the voltmeter taken as a measure of the intensity of X radiation through the chamber.

In a typical case, an ionization chamber having a 200 cubic centimeter volume placed in an X-ray beam of 60 rpm (1 rpm, a minimum usable betatron output) will produce according to (16) a current of

\[(3.33 \times 10^{-18}) \times 200 \times 1 = 6.67 \times 10^{-14}\]
coulombs per second, or amperes, which through the 10-megohm resistor will produce 0.67 volt, which offers no problem in measurement.

In some cases where it is undesirable to remove the low-energy component of the X-ray beam by the filtration of an "in-the-beam" ionization chamber, it is necessary to place the chamber to the side of the beam and pick up the "stray" radiation. By means of comparisons with measurements obtained by instruments placed in the beam, a calibration is obtained. Such positioning receives radiation of intensity many orders of magnitude less than one in the beam and requires a higher electrode resistor and a dc amplifier to operate an indicating instrument. Such a circuit is shown in Fig. 14. It is usually necessary to place the first stages of amplification at the location of the chamber, which is often a considerable distance from the point at which readings may be observed. The circuits must be shielded against relatively strong stray magnetic fields of guiding flux frequency and against vibration of frequency the first several integral multiples of guiding flux frequency.

Referring to Fig. 14, to minimize drift and increase stability in general, the preamplifier and amplifier are arranged for 100 per cent negative feedback and supplied from a closely regulated power supply. To avoid high voltages to ground, the power supply provides both a positive and negative bus.

The input signal from the ionization chamber is developed across a 5,000-megohm input resistor; the output signal is developed from the cathode of the last stage of the amplifier to ground. A switch for checking the output instrument reading and no current flowing through the input resistor and balance controls (acting on the cathode potential of the first stage of the preamplifier) for adjusting the output meter to read zero with zero input are provided.

The first stage of the preamplifier, which contains three stages, is a 954-pentode connected as a space-charge grid electrometer type. This tube is operated at reduced filament and anode potentials to minimize grid current. It is followed by one stage of amplification and a cathode follower stage. All filament currents come from the regulated supply.

The amplifier contains sufficient stages to give a net feedback loop gain of about 50,000. The final stage is a cathode follower, the output of which energizes the output instrument.

V. Conclusion

In this paper we have endeavored to present in some detail the electronic circuit components of the betatron. All electronic aspects of the device have not been touched upon, for a complete treatment would include design of the betatron tube, the magnet, and power control circuits; however, solutions to the main electronic circuit problems have been given.

CORRECTION

M. G. Morgan, author of the paper, "A Modulator Producing Pulses of 10⁻⁷ Second Duration at a 1-Mc Recurrence Frequency," which appeared on pages 505-509 of the May, 1949, issue of the Proceedings of the I.R.E., has called the following error to the attention of the editors.

In the computation on the lower left of page 507, the two plus signs should be multiplication signs, to read:

\[e = E(1 - e^{-t/RC})\]

\[R = \frac{t}{C \ln \frac{E}{E - e}} = \frac{\frac{1}{2} \times 10^{-7}}{125 \times 10^{-12} \ln \frac{5,000}{5,000 - 3,000} = 437 \text{ ohms}}\]
The Influence of UHF Allocations on Receiver Design*

JOHN D. REID†, FELLOW, IRE

The appearance of this paper in the PROCEEDINGS OF THE I.R.E. is the result of the activities of the IRE Broadcast Engineers Group.

Summary—The principle of affording protection to receivers from adjacent channel interference by alternate channel assignments has been universally accepted. It is shown that approximately 20 db added protection between receivers from local oscillator radiation can be obtained by alternate channel assignments and receiver intermediate frequency standardization. There is also developed a square assignment plan in which added protection is provided for the picture image. It is shown that 41.25 is the optimum value of intermediate frequency for joint vhf-uhf usage.

INTRODUCTION

IN CONSIDERING the allocation of a new television band, it would seem justified to consider receiver design problems which affect the performance and cost of receivers in the allocation problem. It is proposed that an intermediate frequency be standardized that is satisfactory for both very-high frequency and ultra-high frequency bands and that the uhf assignments be made in a manner to afford a degree of protection to receivers from oscillator radiation and image response, and so that local oscillator frequencies will come out in whole numbers divisible by a common integer.

1. IF REQUIREMENTS FOR RECEIVER PROTECTION FROM LOCAL OSCILLATOR RADIATION

Table I shows where the local oscillator frequency will fall as the if frequency is varied. From this Table it can be seen that an if between 36 and 42 Mc will place the local oscillator an odd number of channels above (assuming the local oscillator is above the signal frequency). The higher the beat that is produced with the picture the better, but caution should also be observed that the oscillator interference does not fall too close to a sound carrier.

In Table I where the oscillator interference is such as to interfere with sound or picture carrier an even number of channels above, the beat that is produced is italicized. Fig. 1 shows the preferred location for local oscillator interference to fall. The row top curves show the receiver susceptibility to interference which is a maximum at the picture carrier frequency, and falls off sharply on the low side and less sharply on the high-frequency side.

II. IF REQUIREMENTS IN RELATION TO IMAGE RESPONSE

As we go higher in signal frequency, radio frequency preselection becomes increasingly difficult to obtain, and the use of uhf rf stages ahead of crystal converters will generally result in a reduction in signal-to-noise ratio. The if selectivity which can be obtained from two circuits in the uhf band is most unsatisfactory when considering the image rejection requirement predicated upon using if's below the uhf band. As if in the neighborhood of 400 Mc would minimize the image problem, but such an if is not practical today from gain and selectivity standpoint. It is, of course, highly desirable that a common if be used for combined vhf-uhf receivers.

Further consideration of this problem indicates that by careful design and the use of two tuned circuits in the preselector, an average image attenuation of 10 to 15 db can be expected in the uhf band in conjunction with a 41-megacycle if. This average figure would be approximately double at the low end of the band and approximately half at the high end of the band.

Table II shows the interference which would result in respect to if's between 36 and 42 Mc. Figs. 2, 3, and 4 diagram the information contained in this Table. The curves of receiver susceptibility are drawn in an inverted position in Fig. 2, and the distance they are placed above the base line indicates the amount of receiver image attenuation required. In this diagram, the assumption is made that the image interference will be down 15 db in respect to the local station. These figures illustrate the

<table>
<thead>
<tr>
<th>IF</th>
<th>Osc. Fall</th>
<th>Beat with Pic.</th>
<th>Beat with Sound</th>
<th>Beat with Adj. Pic. or Sound</th>
</tr>
</thead>
<tbody>
<tr>
<td>21.25</td>
<td>4 Channels Above</td>
<td>1,750 Above Pic.</td>
<td>OK</td>
<td>OK</td>
</tr>
<tr>
<td>21.9</td>
<td>4 Channels Above</td>
<td>2,300 Above Pic.</td>
<td>OK</td>
<td>400 Above Sound</td>
</tr>
<tr>
<td>30.4</td>
<td>6 Channels Above</td>
<td>1,100 Below Pic.</td>
<td>OK</td>
<td>OK</td>
</tr>
<tr>
<td>36.4</td>
<td>7 Channels Above</td>
<td>1,100 Below Pic.</td>
<td>OK</td>
<td>OK</td>
</tr>
<tr>
<td>36.6</td>
<td>7 Channels Above</td>
<td>900 Below Pic.</td>
<td>400 Above Sound</td>
<td>OK</td>
</tr>
<tr>
<td>36.75</td>
<td>7 Channels Above</td>
<td>725 Below Pic.</td>
<td>OK</td>
<td>OK</td>
</tr>
<tr>
<td>41.20</td>
<td>7 Channels Above</td>
<td>3,700 Above Pic.</td>
<td>OK</td>
<td>OK</td>
</tr>
<tr>
<td>41.25</td>
<td>7 Channels Above</td>
<td>3,750 Above Pic.</td>
<td>OK</td>
<td>OK</td>
</tr>
<tr>
<td>41.75</td>
<td>7 Channels Above</td>
<td>4,250 Above Pic.</td>
<td>250 Below Sound</td>
<td>1,500 Below Pic.</td>
</tr>
</tbody>
</table>

| 475 Mc | 559 Mc |

Table I

Oscillator Radiation

* Decimal classification: R583.5X R007.1. Original manuscript received by the Institute, June 13, 1949. Presented, Third Annual Spring Conference, Cincinnati, Ohio, April 23, 1949. 1 Crosby Division, Avo Manufacturing Corporation, Cincinnati, Ohio.
case for the first channel of the uhf band (channel 14).

In Fig. 2 it will be seen that an if of 36.775 Mc would cause the image to fall 13 channels above channel 14, or in channel 27 with the picture image falling exactly on the sound frequency of channel 27 and the sound image falling exactly on the picture frequency of channel 27. In this instance, assume the carriers in channel 27 were down 15 db from the carriers of channel 14, or from the carriers of channel 28 which might be another local station. Then, in order to obtain a 40-db favorable ratio for the desired picture, the receiver image attenuation required would be 25 db. The inverted receiver susceptibility curve shows that, for equal protection from channel 28 picture carrier, 30-db receiver attenuation would be required.

Similarly, Fig. 3 shows that if an if of 41.25 Mc were used, the image attenuation required of the receiver would be between 18 and 24 db. Eighteen db would take care of the channel-29 picture carrier, while 24-db attenuation would be required for the channel-28 sound. If, as is likely, channel 28 is in an adjacent area to channel 14, then an additional 6-db protection or more would be afforded and channel 29 picture would be the limiting interference. Fig. 4 illustrates that an if of 41.65 looks to be the optimum in that the receiver attenuation required would be between 12.5 and 18.5 db.

III. Optimum IF for Combined VHIF-UHIF Usage

The RMA R-4 Committee considered various ifs between 21 and 42 Mc for uhf usage and those 41.2 Mc as first choice for proposed standardization. Re-examining the factors entering into a choice of 41.2, we find that the second harmonic of the sound it falls 850 kc below channel-6 picture, 41.65, as indicated by image requirements for the uhf band, would put this second harmonic within 50 cycles of channel-6 picture, and is therefore undesirable. 41.3, as indicated by uhf local oscillator radiation requirements, would put this second harmonic 650 below channel-6 picture, which could probably be tolerated. However, there is another consideration which should enter into the final choice of the exact if frequency. It would be highly advantageous in respect to future developments where local oscillator frequencies might be obtained by multiplication and/or addition, that the local oscillator frequency be even integers. With this requirement in mind, 41.25 Mc seems to be the best compromise in respect to all of the preceding factors, and in that it gives us a series of local oscillator frequencies starting with 522 Mc and increasing in 6-Mc steps, all of which are divisible by 6.

IV. Theoretical Allocation Plans

The ideal allocation plan, from a standpoint of conservation of spectrum, would be one which used adjacent channels in the same area. If all local areas had the same requirements for channels, we could use an alternate channel assignment basis with the same efficiency. Since areas do not have equal channel requirements, there is a sacrifice in spectrum usage from an alternate channel assignment basis, but this is recognized as necessary in order that receiver design is not unduly complicated. Present regulations provide at least 6-db protection to a transmitter's service area in respect to adjacent channel interference. This requires approximately a 100-mile separation between adjacent channel stations. Cochannel interference requires some 40 db protection, and may be obtained by spacing cochannel stations approximately 200 miles apart. Image interference in the uhf band will be a serious problem, and certainly one of equal if not greater importance than adjacent channel interference, in respect to receiver design.

TABLE II

<table>
<thead>
<tr>
<th>IF</th>
<th>Pic. Image</th>
<th>Sound Image</th>
</tr>
</thead>
<tbody>
<tr>
<td>36.4</td>
<td>13 Channels Above</td>
<td>13 Channels Above</td>
</tr>
<tr>
<td></td>
<td>700 Below Sound</td>
<td>700 Below Pic.</td>
</tr>
<tr>
<td></td>
<td>1,900 Below Pic.</td>
<td></td>
</tr>
<tr>
<td></td>
<td>3,800 Above Pic.</td>
<td></td>
</tr>
<tr>
<td>36.6</td>
<td>13 Channels Above</td>
<td>13 Channels Above</td>
</tr>
<tr>
<td></td>
<td>300 Below Sound</td>
<td>300 Below Pic.</td>
</tr>
<tr>
<td></td>
<td>1,500 Below Pic.</td>
<td></td>
</tr>
<tr>
<td></td>
<td>4,200 Above Pic.</td>
<td></td>
</tr>
<tr>
<td>36.775</td>
<td>13 Channels Above</td>
<td>13 Channels Above</td>
</tr>
<tr>
<td></td>
<td>0 Beat with Sound</td>
<td>0 Beat with Pic.</td>
</tr>
<tr>
<td></td>
<td>1,250 Below Pic.</td>
<td></td>
</tr>
<tr>
<td></td>
<td>4,500 Above Pic.</td>
<td></td>
</tr>
<tr>
<td>41.25</td>
<td>15 Channels Above</td>
<td>14 Channels Above</td>
</tr>
<tr>
<td></td>
<td>3,000 Above Sound</td>
<td>3,000 Above Pic.</td>
</tr>
<tr>
<td></td>
<td>1,500 Above Pic.</td>
<td>1,500 Below Sound</td>
</tr>
<tr>
<td></td>
<td>3,000 Below Sound</td>
<td></td>
</tr>
<tr>
<td>41.65</td>
<td>15 Channels Above</td>
<td>14 Channels Above</td>
</tr>
<tr>
<td></td>
<td>3,800 Above Sound</td>
<td>700 Below Sound</td>
</tr>
<tr>
<td></td>
<td>2,300 Above Pic.</td>
<td></td>
</tr>
<tr>
<td></td>
<td>2,200 Below Sound</td>
<td></td>
</tr>
<tr>
<td>475 Mc</td>
<td></td>
<td></td>
</tr>
<tr>
<td>559 Mc</td>
<td></td>
<td></td>
</tr>
<tr>
<td>643 Mc</td>
<td></td>
<td></td>
</tr>
<tr>
<td>14</td>
<td></td>
<td></td>
</tr>
<tr>
<td>16</td>
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<tr>
<td>18</td>
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<tr>
<td>20</td>
<td></td>
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<tr>
<td>22</td>
<td></td>
<td></td>
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<td>24</td>
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<td>26</td>
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<td>28</td>
<td></td>
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<tr>
<td>30</td>
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<tr>
<td>32</td>
<td></td>
<td></td>
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<tr>
<td>34</td>
<td></td>
<td></td>
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<tr>
<td>36</td>
<td>(36)</td>
<td></td>
</tr>
<tr>
<td>38</td>
<td></td>
<td></td>
</tr>
<tr>
<td>40</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Fig. 2— Image of channel 14 with 36.775-Mc intermediate frequency.

Fig. 3— Image of channel 14 with 41.25-Mc intermediate frequency.

Fig. 4— Image of channel 14 with 41.65-Mc intermediate frequency.
Therefore it is logical that we should obtain as much or more protection for image interference in the allocation plan as we are now providing for adjacent channel interference. To do this, the image channel stations should be spaced in excess of the 100-mile separation provided for adjacent channel stations, and, if spaced by 150 miles, would afford approximately 15-db protection.

Circular, equilateral, triangular, and square allocation plans were investigated, and the square plan seemed to offer the greatest promise of providing image protection without increasing the number of channels required. Making the sides of the square equal to the adjacent channel spacing gives the diagonal of the square equal to 1.4 times the side or for 105 miles adjacent channel spacing, the diagonal will be 148 miles. Therefore, if the image channels are kept on the diagonals, the desired image channel separation will be obtained. The cochannel spacing will automatically be twice the adjacent channel spacings.

Fig. 5 shows a block of 28 6-Mc channels. This block is divided into half giving two groups of 14 channels. In the first group, alternate channels are designated A and B, and in the second group, C and D. Considering an if of 41.25 Mc, this will place the image of channels A in channels D, and the image of channels B in channels C. This gives us four types of stations to fit our square allocation plan. (There are only two basic types, as channels C are an extension of A, and channels D an extension of B.)

Fig. 6 illustrates the square assignment plan with channels A, B, C, and D at the corners of a square and the image channels separated by the diagonals of the square.

A transparent grid of this basic pattern can be laid over the map as a guide in allocation. Consideration of the basic square assignment shows a large middle area without any guidance as to which type channel would be best to assign. This can be rectified by breaking the block assignment into 8 blocks rather than 4. Each type channel is split into two giving A1, B1, A2, B2, C1, D1, C2, D2, as illustrated in Fig. 5.

Channels A1 will have their image (pic) in channels D2.
Channels A2 will have their image (pic) in channels D4.
Channels B1 will have their image (pic) in channels C1.
Channels B2 will have their image (pic) in channels C2.

The above blocks allow 3 or 4 stations of one type in a given locality, whereas the former plan permitted 7 of each type. We can then fill in the middle areas of our squares by displacing A2, B2, C2, D2 uniformly from A, B, C, and D, as illustrated in Figure 6.

This plan still maintains our 210-mile cochannel, 105-mile adjacent channel, and 148-mile image channel separation.

This plan as illustrated for 28 channels could be repeated if double the number of stations per area were required.

Wherever possible, however, all stations in one area should be grouped. The assignment of stations in blocks for a given area is highly desirable from the standpoint of receiver performance and user convenience. The receiver performance will be improved by increased antenna efficiency and the possibility of improving the tracking of rf circuits over a limited portion of the band. The use of bandpass acceptor or rejector circuits would be facilitated. The user would find it convenient to have his local stations groups within a small area of the tuning dial.

**CORRECTION**

J. H. Mulligan, Jr., author of the paper, "The Effect of Pole and Zero Locations on the Transient Response of Linear Dynamic Systems," which appeared on pages 516–530 in the May, 1949, issue of the PROCEEDINGS OF THE I.R.E., has brought to the attention of the editors the following errors:

Page 517, equation (6): The upper limit g in the second term of the denominator should be replaced by q. Page 521, last line, left-hand side: "sin (β4 - λ)" should be "sin (β4 + λ)." Page 521, third line from bottom of page, right-hand side: "B3," should be "β3." Page 522, first line, right-hand side: "B4," should be "β4." Page 522, line 14, right-hand side: \(D_{11}, D_{12}^2\) should be \(D_{11}^1, D_{12}^2\).
Electrodes for Vacuum Tubes by Photogravure*

MARSHALL P. WILDER†, ASSOCIATE, IRE

Summary—An improved method of photogravure is described which is particularly suited for making up small lots of electrode parts for vacuum tubes. Complicated designs and configurations can be cut with a minimum of effort, and frequent changes prior to final tooling involve little time and expense. Its principal utility is in the flexibility introduced by this improved process enabling precise parts to be made for new experiments in vacuum-tube structures.

INTRODUCTION

THE WELL-KNOWN photomechanical process of photogravure has been used for many years to transfer variations in tone of a photographic image to variations in depth or area on a metal plate. This plate, properly inked, is used to reproduce the original image by all varieties of the printing art. Photogravure provides an inexpensive method whereby complicated vacuum-tube electrode of 0.002 to 0.015 inch thick sheet, such as nickel, iron, or stainless alloys, may be produced in an easily reproducible form and in a wide variety of shapes. Resolution of the order of 40 lines per millimeter has been done with fair results; however, for such fine work, thin copper 0.002 inch thick is required. In general, vacuum-tube electrodes do not require definition much greater than 15 lines per millimeter. With reasonable care, this is readily accomplished. Electrolytic etching is used for all tube parts except for very fine screens 500 to 1,000 holes per linear inch. This method is faster and undercut much less than more conventional ferric chloride etching procedures. (See Fig. 1.)

The photoengraving process consists of applying to the metal a light sensitive coating which becomes less water soluble when exposed to light. This coating, therefore, is referred to as a “resist.” On exposure to light of the plate covered by a negative, the areas protected by the dark lines on the plate remain soluble and can be washed off with water, leaving the light struck portions of the coating intact. The resist is then soaked in a solution of ammonium dichromate, dried, and baked to further insolubilize the resist.

The unprotected portions of the metal can then be etched. It is necessary to completely etch through the metal plate and preferable to etch through simultaneously from both sides; therefore, exposure of both sides of the metal plate is accomplished by using an envelope negative. Duplicate negatives of the design required are placed together so that the images match and three edges are sealed together, while the fourth is left open for insertion of the metal plate. (The term “negative,” as used in this paper, refers to what is called a positive transparency in the usual photographic terminology.) Considerable difficulty is encountered in etching the metal by ferric chloride because material formed in the cut slows down the etching process and the time required to completely etch through results in lifting of the resist. Electrolytic etching was developed as a procedure, using solutions of acids and acid salts, and resulted in reducing the etching time to less than a tenth of that otherwise required.

Detailed Procedure

Preparation of Negatives

The lines in the art work should be approximately \( \frac{1}{4} \) inch wide when the work is to be reduced photographically approximately 5 to 1. The work is reduced to actual size by photography and sufficient positives made so that an envelope can be formed registering the pieces carefully under a microscope if necessary and cementing three sides of the envelope. (See Fig. 2.)

When it is desired to etch out areas more than about 0.040 inch across, it is usually preferable to prepare the original drawing in such a way that only the boundary of the area has to be etched, allowing the inside to drop out rather than etching the entire area. It is also desirable to avoid extremely fine lines on the envelope negative such as those under 0.002 inch.

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The resist is prepared in two steps. A stock albumen solution is made up first, and this is combined with glue and more ammonium dichromate as detailed below:

**Stock Albumen Solution:**
- 57 gms. Egg Albumen
- 240 ml. Water
- 0.1 ml. Chloroform
- 4 drops Ammonium Hydroxide (Sp. Gr. 0.90)

This is allowed to stand overnight and filtered through absorbent cotton. After addition of 30 ml. of ammonium dichromate solution (22 gms. per 100 ml. of solution), the stock solution is diluted to 280 ml. with water.

**Resist Solution:**
- 54 gms. Photoengravers Glue (Rogers)
- 22.4 ml. Albumen Stock Solution
- 41 ml. Ammonium Dichromate Solution (22 gms. per 100 ml. solution)
- 100 ml. Water

The ammonium dichromate solution is diluted with the water before addition of the glue and albumen. The entire solution is diluted to 300 ml. with water.

The solutions should be protected from light during storage and may not work satisfactorily if more than a week old. The protein material becomes less soluble in water on aging, with the result that the unexposed parts of the image are not completely free from resist after development. Higher temperatures of storage increase the rate of aging.

The resist solution is filtered through absorbent cotton in a funnel into the container to be used for dipping immediately before use. The end of the stem of the funnel should touch the side of the container to eliminate bubbles. The plates are dipped in the resist held horizontal for five seconds, then immediately redipped and held horizontal for 20 seconds. They are then hung vertically by one corner to drain and partially dry. After about one hour, they are again dipped in the same manner as before. After draining, the drop of solution at the lower corner should be wiped off. The plates are allowed to dry until no longer tacky. The plates should be dipped and dried where they will be exposed to a minimum of light, particularly direct light. The coated plates can be kept in the dark for 24 hours without harm, but if kept for longer periods, particularly where warm, they may be difficult to develop satisfactorily.

**Exposure and Development**

The dried plates are inserted in the envelope negative previously described and exposed to light in a printing frame. The simple type of printing frame can be used for very flat, flexible plates, but a photoengraver’s vacuum printing frame is preferable. This frame consists of a ribbed rubber blanket with a connection for attaching a vacuum line, and a piece of ¼-inch thick plate glass, as shown in Fig. 3. These frames are available from photoengraving supply houses. One No. 2 photoflood bulb in a 10-inch aluminum reflector held about 5½ inches above the work produces sufficient light on a 30-second exposure. The usual printing cabinets do not give nearly as much light, with the result that excessively long exposures are required. This has the disadvantage of overheating the negative and plate, as well as being time consuming. After the exposure, the negative with the plate is reversed, being careful not to move the plate in the envelope, and the reverse side exposed similarly.
The exposed plate is developed by sweeping it back and forth in water at room temperature (20 to 30°C) for one minute and rinsing in clean water. The developed plate is then immersed for three minutes in a 3 per cent ammonium dichromate solution in water. After a rinse in clean water, it is now ready for drying and baking. This is accomplished by heating to about 300°C for 8 minutes to harden and further insolubilize the resist. The image should be readily visible, but the resist should be a very pale golden color. The plate can now be stored for extended periods of time without deterioration. Before etching, the edges should be coated with a lacquer such as a 5 per cent nitrocellulose solution to reinforce the resist. The times and temperatures given above are not critical.

Etching

Ferric chloride solutions of 32 to 44 per cent concentrations are commonly used for commercial photoengraving, but are slow in action and not particularly satisfactory in hand operations. Electrolytic etching with acid salts, such as ferric chloride and sodium acid sulfate, or with acids such as sulfuric acid, were found much more rapid and gave much better results than simple immersion in ferric chloride. Sulfuric acid solution containing 50 per cent acid by weight is particularly satisfactory. This solution can be used several times and has been discarded only when it became so dark from dissolved salts that it was difficult to see the work.

The electrolytic cell consists of a suitable container in which two flat iron or steel electrodes exceeding the length and width of the plates to be etched by at least \( \frac{3}{4} \) are held vertically and parallel. The distance between them may be from 2 to 4 inches for plates up to 3 inches square. Make the work the anode. A current density of the order of 15 amperes per square inch of actual surface to be etched has been found satisfactory. The area to be etched will, of course, be the surface area not protected by the resist, and is independent of the total area of the plate. A voltage of 3 to 4 volts measured between the anode and cathode of the cell is sufficient for etching small areas totaling about ½ square inch. (See Fig. 4.)

The temperature of the etching solution is of importance primarily because of its effect on the resist coating. The coating is less rapidly affected by the acid solution at temperatures below 20°C than at 30°C. Plates of Nichrome V with a thickness up to 0.006 inches are rapidly etched, requiring only 5 to 10 minutes at temperatures up to 20°C. Nickel of the same thickness requires about 50 per cent more time, and is best carried out at lower temperatures, under 20°C. Temperature has little or no effect on the rate of etching.

Removal of Resist

The resist coating can be readily removed from the etched work by boiling in the same caustic solution used for cleaning the metal plates originally.

Conclusions

The photogravure method is suitable for making up small lots of electrode parts for vacuum tubes. Rather complicated shapes and designs can be cut out and, later, formed by drawing and bending into a desired shape. The possibility of saving time and money, particularly where frequent changes in design may become necessary prior to the final tooling, makes this method a valuable tool in vacuum-tube research and production.

Acknowledgment

The writer wishes to acknowledge the very valuable help and contributions made by Earl S. Bidgood of this Laboratory. Electrolytic etching and the additional hardening of the developed resist coating by soaking in ammonium dichromate were suggested and developed by him. Fine mesh screens (Fig. 4) were prepared employing similar procedures by Buckbee Mears Company of St. Paul, Minn.
Short Antenna Characteristics—Theoretical*

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Summary—The experimental data obtained by Smith and Johnson on short antennas top-loaded by an umbrella were analyzed mathematically. The theory developed therein is within practical agreement with the measurements. It shows that, for single-tower operation, the horizontally polarized radiation is negligible. In addition, it is shown that with the optimum length of umbrella, the vertical radiation characteristic is the same as it would be from the radiator without top loading. The paper provides a tool for further investigation of other top-loading arrangements.

INTRODUCTION

DURING THE WAR, the performance of short antennas, on the order of 36° or \( \frac{\lambda}{10} \) high, was scrutinized. Smith and Johnson investigated experimentally in Cleveland the performance of a vertical tower top-loaded by an umbrella of various lengths. This paper was prepared in an endeavor to explain theoretically the measurements made at Cleveland and to suggest a possible better arrangement.

DISCUSSION

The first arrangement tried by Smith and Johnson was an "umbrella" loading such as shown in Fig. 1(a). The angle \( \Delta \) used was \(-40^\circ\). In order to simplify computation, it was assumed that the top-load consisted of a perfectly conducting cone as shown in Fig. 1(b).

This will be called "cone" loading to distinguish the theoretical from the measured.

In the Appendix is shown the derivation of the formulas used to compute the radiation from the cone. As may be seen from the formulas in the Appendix, computation is greatly simplified if the angle in Fig. 1(b) is chosen as \(-45^\circ\). Therefore, this angle was used rather than the \(-40^\circ\)-angle used in Cleveland. In Fig. 1(b), the current was assumed to be zero at the periphery of the cone. Radiation from the vertical lead to the apex of the cone was computed from the conventional formula for a vertical antenna with a nonradiating top load.

It was assumed that the cone was opaque to the radiation from the vertical. Mechanical integration and comparison with a "standard" \( \frac{\lambda}{4} \) antenna was used to obtain the resistances of the various arrangements.

The second arrangement tried in Cleveland had a skirt at the end of the umbrella wires referred to as "umbrella with skirt" as shown in Fig. 2(a). Again in order to compute the performance of this arrangement, a cone was substituted for the wires and a nonradiating load was placed at the periphery of the cone to simulate the loading due to the skirt as shown in Fig. 2(b). The theoretical arrangement will be called "loaded-cone" top loading.

A third arrangement, called "folded umbrella" that was not tried at Cleveland, suggested itself to the author subsequently and is shown in Fig. 3(a). In this arrangement, the umbrella wires are connected at the outer periphery by a skirt but the alternate wires have an insulator at the top of the antenna. The simulated

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§ The negative sign is used because the angle is opposite in sign to that used in the development shown in the Appendix.

Fig. 1—Top-loaded vertical antenna; (a) umbrella loading and (b) cone loading.
theoretical arrangement is shown in Fig. 3(b) and is called "double-cone" top loading.

The computed base resistance for the three arrangements is shown in Fig. 4 plotted for various radii of the cone. Also plotted in Fig. 4 is the Smith-Johnson experimental data for the condition where the Cleveland tower was $\lambda/10$ high. It is to be noted that the base resistance at the zero radius is somewhat different for the measured and the computed. Two factors contribute mainly to this. The first is that the Cleveland tower had a considerable cross section compared to the wavelength and, therefore, the velocity of the propagation on it was lower than the speed of light assumed in the computations. This would lower the measured base resistance. Secondly, the large base shunt capacitance of the tower would tend to lower the measured resistance.

As can be seen from Fig. 4, as the radius of the cone was expanded out from zero, the measured resistance increased more rapidly with increase of radius than did the computed values. Again there are two principal reasons for this.

In the Smith-Johnson experiments, the first increment of top loading added included the "hat" on the WHK tower. It can be seen from Figs. 1, 2, and 3 that there is radiation from the "umbrella" and the "umbrella with skirt" that is out of phase with the radiation from the vertical. Mechanically, the "hat" is substantially horizontal and, therefore, contributes practically nothing to the field at the same time producing a large "capacitance" top load. This would then
make the measured resistance rise more rapidly than the computed resistance. Secondly, the Cleveland umbrella was at a flatter angle and, therefore, for any given length of wires the vertical currents which are out of phase with the current in the vertical thus producing an out of phase field would be less than in the computed arrangements. Therefore, as the radius of the top-load was increased from zero, the measured resistance should increase more rapidly than the computed.

It is to be noted that the "umbrella with skirt" came up to the same maximum base resistance as the "umbrella" but at a lesser radius. The computed arrangements show the same thing. It may be that, in practice, the skirt wires are undesirable mechanically in view of the fact that the same effect can be obtained with an "umbrella" of slightly greater radius. The performance of a "double cone" (in practice a "folded umbrella") top-load as shown in Fig. 4 indicates that this arrangement may have enough merit to warrant using a skirt. It is expected that a considerably higher base resistance may be obtained at a somewhat smaller radius.

In the Appendix, it is shown that the radiation from the cone is entirely vertically polarized. Referring to Fig. 2(a), if the system of wires is viewed from any point in a plane perpendicular to the ground and the plane containing any wire and its complement wire at azimuth 180° or viewed from any point in a plane drawn through the apex of the umbrella and bisecting the angle between two of the wires, that the viewer is looking at a symmetrical mechanical arrangement of wires such that each horizontal component of current on any wire has an equal but opposite component elsewhere at the same distance from the viewer. Therefore, in these six planes, there is no horizontal radiation. If all the possible planes are inserted there will be an azimuth angle of 22.5° between planes. The angle of greatest dissymmetry will be midway between planes; i.e., 11.25° to any given plane. Viewed from any point in a plane placed thusly, there is very little dissymmetry and, therefore, the horizontal radiation is extremely small compared to the vertical radiation. A similar reasoning applied to the "folded umbrella" shown in Fig. 3 leads to the conclusion that the horizontally polarized radiation is also small in this arrangement. Therefore, in many applications, the top loadings discussed herein are satisfactory on the score of magnitude of horizontal radiation.

The vertical form factors for the individual components as well as the total system of a "double-cone" top-loaded antenna of δ/10 height were computed for a radius of 0.04 δ for the cone and are shown in Fig. 5. For comparison purposes, the vertical form factors, for a zero height, a δ/4 height, and for the δ/10 height top-loaded, assuming the top loading were nonradiating antennas, are shown in Fig. 5. The form factor for a δ/10 height without any top loading conforms almost exactly with the radiation from the total antenna and is, therefore, not plotted. This indicates that the improvement in radiated field, as observed by Smith and Johnson, was due almost entirely to reduction in ground losses and not to an improvement in the vertical form factor.

Loop-radiation resistances were computed for the various elements and for the total system of a "double-cone" top-loaded antenna and are shown in Fig. 6. This shows that if the optimum top loading could be effected without it contributing to the radiation field, a δ/10 antenna would have a loop-radiation resistance of 14.9 Ω as compared with 1.46 Ω for the same height without any top loading. The base resistance in each case would be higher. Unfortunately, in the arrange-
ments under discussion as the radius is increased, the radiation from the top-load which is 180° out of phase with the radiation from the vertical, increased rapidly. This is shown by the increase in radiation resistance of cone “a” with increase in radius of the cone, the radiation from which is out of phase with the radiation from the vertical as shown in Fig. 6.

As a side light to the theme of this paper, Fig. 6 is interesting in regard to the radiation from “hats” as they have been used on broadcast station towers. Cone “a” at a radius of 0.044 λ or 15.84° has a loop resistance of 0.1 Ω. In the case of a “hat,” the mechanical arrangement in the vertical plane is flatter than the computed cone and, therefore, would have a still lower loop-radiation resistance. On the other hand, in broadcast applications the “hat” is at a greater height, usually on the order of 120° which would raise the resistance. It is, therefore, reasonable to assume that the radiation resistance of the “hat” as used in broadcasting, is only a fraction of an ohm. Therefore, the intrinsic radiation from the “hat” is negligible in view of the much higher loop-radiation resistance of the tower. Due to symmetry of the mechanical arrangement, as viewed from any distant point, there can be no horizontally polarized radiation from a “hat” as used in broadcasting.

CONCLUSIONS

The observations made by Smith and Johnson have been explained theoretically. Certain assumptions were used in the analysis, however, it is believed that they have been justified in the paper. A refinement of the Smith-Johnson arrangements has been suggested which promises to effect a worthwhile improvement.

APPENDIX

Radiation from a Cone Antenna

Introduction

The vertical plane component, for all zenith angles, of the electric-field intensity is computed for a sinusoidally varying current flowing from the periphery to the apex of a perfectly conducting cone antenna over perfectly conducting earth.

The first step in determining the electric-field intensity from the cone antenna is to divide the area of the cone into infinitesimal areas. The current passing through each infinitesimal area is next determined and then the electric-field intensity due to each element of current. Finally, the total electric-field intensity due to the cone is determined by adding, vectorially, all the individual components.

General Radiation Formula

The increment of the electric-field intensity due to current flowing in an infinitesimal antenna in vacuum and sufficiently accurate in air is

\[ \delta(E_0) = \frac{\pi^4}{r^2} I \sin \tau e^{-ikr} f(\delta r). \]  
(1)

The mks system of practical units is used where

- \( E_0 \) = the electric-field intensity in volts per meter at a distant point \( P \)
- \( \delta r \) = the infinitesimal length of the infinitesimal antenna
- \( f(\delta r) \) = the scalar component of \( \delta r \) whose normal component to \( r' \) contributes to the field at point \( P \)
- \( I_0 \) = the current in amperes flowing in the infinitesimal element \( \delta r \)
- \( r' \) = the distance in meters from \( \delta r \) to the point \( P \)
- \( k = \frac{2\pi}{\lambda} \)
- \( \lambda \) = the operating wavelength in meters
- \( \tau = \) the angle between \( f(\delta r) \) and \( r' \)
- \( e^{-ikr} = \) the retarded vector
- \( j = \) the operator \( \sqrt{-1} \)
- \( \eta = 120\pi \).

Ramo and Whinnery derived \( (1) \) by taking the curl of the vector potential due to a sinusoidally varying current flowing in an infinitesimal element. A similar development is used by others, such as Skilling.

Infinitesimal Areas on Cone

Referring to Fig. 7, the surface of the cone is divided radially into rings of infinitesimal width \( \delta r \) such as shown at a distance \( r \) in meters from the apex of the cone or at a radius \( r \) in meters.

\[ \text{S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, New York, N. Y., 1944; p. 431. Note that their } \theta \text{ has been changed to } r; r \text{ to } r', \text{ and } h \text{ to } f(\delta r). \text{ This is for convenience.}
\]

Each ring is, in turn, divided into infinitesimal segments. At any angle \( \phi \); an increment of angle \( \delta \phi \), is added such as shown in Fig. 7. The length of an arc on a circle equals the radius times the angle of arc in radians. The arc length of the segment of the ring is then \( r \delta \phi \) long. By selecting successively the proper values of \( r \) and \( \phi \) and adding the increments \( \delta r \) and \( \delta \phi \), the area of the cone is divided into infinitesimal areas of size \((\delta r)(r\delta\phi)\).

**Current Distribution**

The rigorous solution for the current distribution on a cone can be obtained by the use of principles set forth by Hallen\(^4\) or Schelkunoff.\(^5\)-\(^7\)

The rigorous solution is involved and requires a great deal of labor even for a single case. Therefore, an approximation was sought which would be acceptable. It is known that as the diameter of a cylindrical or the diameter of the base of a conical antenna is increased, the phase velocity of propagation decreases. King and Harrison\(^8\) have made solutions for the current distribution on cylindrical antennas of practical diameters.

Schelkunoff\(^9\) also has made semirigorous solutions for the current distribution on cylindrical antennas of practical diameters. On page 504 of footnote reference 6, Schelkunoff said, "In practice, precise knowledge of current distribution is of lesser importance than knowledge of the input impedance. This is because the directive gain of antennas and their radiation patterns are not very sensitive to the changes in the current distribution. Radiation patterns will be affected seriously only in those directions in which radiation is small."

Harrison and King\(^9\) said, "The results of the present analysis indicate that for purposes of computing the shape of the pattern of the distant field, even for relatively thick antennas, a simple sinusoidal distribution is entirely adequate if it is understood that sharp zeros are rounded off."

Both of these comments are applicable to cylindrical antennas of practical dimensions on the order of height to radius ratios of 50 to 1 or greater. To how much smaller ratios they can be trusted will be left to the mathematicians.

Where cones are treated in this paper, practically a limited number of wires may replace the solid cone surface in many applications. Therefore, carrying the solution for the cone beyond the first approximation, as is done in this paper, may not be justified.

At the apex of a cone, the diameter of the conductor is zero and, therefore, the velocity of propagation is the velocity in free space. As the diameter increases toward the periphery, the velocity gradually decreases.

Schelkunoff\(^10\) has given the following approximate formula for the current on a cylindrical antenna\(^11\)

\[
\frac{I_0(r)}{I_0} = \left( \sin \beta (l - r) + \frac{F(L)}{K} \cos \beta (l - r) \right) - j \frac{G(L)}{K} \cos \beta (l - r)
\]

where

- \( l = \) total length of the antenna
- \( r = \) distance from the base to a point of interest
- \( I_0(r) = \) current at point \( r \)
- \( I_0 = \) loop current
- \( \beta = 2\pi/\lambda \)
- \( j = \sqrt{-1} \)

\( l, r, \text{ and } \lambda \) are measured in the same units.

\( F(L), G(L), \text{ and } K \) as defined in Schelkunoff's article.\(^8\)


\(^3\) See p. 504 of footnote reference 6.

\(^4\) The symbols used in the text from the beginning of (2) to the beginning of (6) are not necessarily the same as in the rest of the paper because Schelkunoff's symbols have been used.
If the differential of (2) in respect to \( r \) is taken and set equal to zero, the points of maximum and minimum current may be determined by solving for \( r \). Performing this operation

\[
\tan 2\beta (l - r) = \frac{2KF(L)}{F^2(L) + G^2(L) - K^2}.
\] (3)

Referring to Fig. 8, the distance from the periphery to the first maximum \((l - r_1)\) would be \( \lambda/4 \) if the velocity were that in free space. The apparent velocity is then proportional to

\[
V_{\text{max}} = \frac{4(l - r_1)}{\lambda}.
\] (4)

Similarly, the distance from the periphery to the first minimum \((l - r_2)\) would be \( \lambda/2 \) if the velocity were that in free space. The apparent velocity is then proportional to

\[
V_{\text{min}} = \frac{2(l - r_2)}{\lambda}.
\] (5)

Using a length equal to \( \lambda/2 \) and varying the radius \( R \) of the cone, the apparent velocity factor was computed from (3), (4), and (5), and is shown in Fig. 9.

Curve \( V_{\text{max}} \) is the apparent velocity factor using (4), and curve \( V_{\text{min}} \) is the apparent velocity factor using (5).

In all cases, the free-space distance from the maximum to the minimum, turns out to be \( \lambda/4 \). This is due to the approximations used in deriving (3) and indicates that the development is not rigorous.

This current distribution development has been fraught with assumptions to the extent that \( V_{\text{max}} \) showed a supposedly unreasonable inclination to approach zero at a radius of 0.5\( \lambda \).

It is believed that the shape of the curves in Fig. 9 should have a uniform trend rather than three trends.

The inverse of \( V_{\text{min}} \) is plotted in Fig. 10. Also plotted is a straight line labeled \( k_1 \) that approximates the inverse of \( V_{\text{min}} \) at small values.

Arbitrarily "velocity factors" in accordance with \( k_1 \) of Fig. 10 will be used. The formula is

\[
k_1 = \text{velocity factor} = 1.0 + R/\lambda
\] (6)

where

- \( R \) = the radius of the cone
- \( \lambda \) = the free-space operating wavelength.

The way \( k_1 \) is used in the subsequent development, it is assumed that the velocity of propagation is the same over all of the cone.

No mention has been made of the fact that the current has a complex character or have the mutual effects been commented on except as they bear on the comments quoted earlier from Schelkunoff, and King and Harrison.

In this development, the current distribution will be assumed to be a sinusoidal distribution modified by the velocity factor \( k_1 \).

**Current Passing through an Infinitesimal Arc \( r\phi \)**

Let

\( I_r \) = the total current in amperes flowing through any given ring such as shown in Fig. 7.

\( I_\phi \) = the current in amperes flowing through the infinitesimal arc segment \( r\phi \).

Then

\[
I_\phi = \frac{I_r \delta\phi}{2\pi}.
\] (7)
Referring to Fig. 11, it is assumed that the current $I$, has a sinusoidal distribution. In setting up the formula for the current distribution, the cone dimensions such as $R_1$ and $r_1$ will be multiplied by the velocity factor $k_1$ as given by (6). The electrical length of the cone will be greater than the physical length by the factor $k_1$.

It is also assumed that there is a nonradiating top-load $L$ in meters at the apex of the cone as shown in Fig. 11.

Referring to Fig. 11, at any distance $r_1$ from the apex of the cone the current $I$ is

$$I_r = I \sin \frac{2\pi}{\lambda} \left( L + \frac{k_1r}{\cos \Delta} \right)$$

(8)

where $I$ is the loop current in amperes.

Substituting (8) in (7)

$$I_\theta = I \sin \frac{2\pi}{\lambda} \left( L + \frac{k_1r}{\cos \Delta} \right) \frac{\phi}{2\pi}$$

(9)

**DETERMINATION OF $f(\delta r_1)$**

Referring to Fig. 12 a set of co-ordinates $X'$, $Y'$, and $Z'$ have been selected such that the $X'Y''$ plane is parallel to the ground plane and the $X'Z'$ plane is perpendicular to the ground plane.

Because the cone is symmetrical, the electric-field intensity from the cone is the same in all directions at any given distance $r'$ and at any given zenith angle $\theta$. It is, therefore, only necessary to derive the electric-field intensity in one direction. The direction will be chosen such that a line drawn from any current element of interest to the distant point $P$ will lie in the $X'Z'$ plane or a colinear plane. $P$ is a point at sufficient distance from the cone so that only the radiation field is important and so that lines drawn from all points on the cone to $P$ may be considered parallel.

$\delta r_1$ is a current element lying in the cone at distance $r_1$ from the apex. Let $\delta r_1$ be expressed as a vector in terms of three vectors in the $X'$, the $Y'$, and the $Z'$ directions as shown in Fig. 12.

Then

$$\delta r_1 = i'A_x + j'A_y + k'A_z$$

(10)

where

$i' = a$ unit vector in the $X'$ direction,

$j' = a$ unit vector in the $Y'$ direction,

$k' = a$ unit vector in the $Z'$ direction.

An examination of Fig. 12 reveals that each $A_y$ component has an equal and opposite component elsewhere on the cone at the same distance from $P$. The retarded vector potentials for the two components are, therefore, equal and opposite at $P$ and, hence, this component of the electric field is zero. The $A_y$ components are, therefore, of no interest. This is not true of the $A_x$ and the $A_z$ components. The components of $\delta r_1$ of interest, therefore, lie in the $X'Z'$ plane and the colinear planes.

Let $\delta r_{x'z'}$ represent the $X'Z'$ component of $\delta r_1$. Then from (10)

$$\delta r_{x'z'} = i'A_x + k'A_z$$

(11)

From Fig. 12, it is seen that

$$i'A_x = O_1P = i'\delta r_1 \cos \Delta \cos \phi$$

(12)
and

\[ k'Z' = O'Q' = k'd_1 \sin \Delta = k'd_1 \cos \Delta \tan \Delta. \quad (13) \]

Substituting (12) and (13) in (11)

\[ |\delta x'| = d_1 \cos \Delta \sqrt{\cos^2 \phi + \tan^2 \Delta}. \quad (14) \]

Since (14) expresses the \( X'Z' \) scalar component of the current element length \( d_1 \) and since only the \( X'Z' \) components are of interest, then (14) equals \( f(d_1) \) in (1).

Therefore

\[ f(d_1) = d_1 \sqrt{\cos^2 \phi + \tan^2 \Delta}. \quad (15) \]

**Sin \( \tau \) for "Physical-Cone" Element**

The total electric-field intensity at \( P \) is due to the contribution made by the "physical cone" and its image referred to as "image cone." The discussion so far applies equally to both the "physical cone" and the "image cone."

The radiation due to the "physical cone" will now be determined. In (1), the only two parameters undefined so far are \( \sin \tau \) and \( e^{-ikr'} \).

Referring to Fig. 12, the component of \( f(d_1) \) normal to \( r' \) is the only component contributing to the field at \( P \). The term \( \sin \tau \), therefore, appears in (1). \( \tau \) is the angle between \( f(d_1) \) and \( r' \).

Referring to Fig. 12, for the "physical cone," keeping in mind that the current \( I_0 \) flows from the periphery to the apex

\[ \tau_1 = \theta + 90^\circ + \beta. \]

Then

\[ \sin \tau_1 = \sin (\theta + 90^\circ + \beta) = \cos \theta \cos \beta - \sin \theta \sin \beta. \]

After some manipulation

\[ \sin \tau_1 = \frac{\cos \theta \cos \phi - \sin \theta \tan \Delta}{\sqrt{\cos^2 \phi + \tan^2 \Delta}}. \quad (17) \]

**"Physical-Cone" Element Retarded Vector \( e^{-ikr'} \)**

The retarded vector \( e^{-ikr'} \) will now be determined for the "physical cone" element. This vector indicates the effect on the field at the distant point \( P \) due to the time difference between the flow of current in \( d_1 \) and the effect at \( P \). The phase of all the infinitesimal antennas will be referred to a common point \( O \) as shown in Fig. 13. The distance from \( O \) to \( P \) is designated as \( r_0 \).

From Fig. 13 for the "physical cone"

\[ r_1' = r_0 - W_1 - Y_1 - X_1 \quad (18) \]

where

\[ W_1 = (H - H_2) \cos \theta = (H - R \tan \Delta) \cos \theta \]
\[ Y_1 = H_2 \cos \theta = r \tan \Delta \cos \theta \]
\[ X_1 = (O'P'') \sin \theta = r \cos \phi \sin \theta. \]

Multiplying (18) by \( k \) and substituting

\[ k' = \frac{2\pi}{\lambda} \left[ r_0 - (H - R \tan \Delta) \cos \theta - r \tan \Delta \cos \theta - r \cos \phi \sin \theta \right] \quad (19) \]

where

\[ a = \frac{2\pi r_0}{\lambda} \]
\[ b = \frac{2\pi}{\lambda} (H - R \tan \Delta) \cos \theta \]
\[ c = \frac{2\pi}{\lambda} (r \tan \Delta \cos \theta) \]
\[ d = \frac{2\pi}{\lambda} (r \cos \phi \sin \theta). \]

Fig. 13—Determination of retarded vector from a cone antenna.

Then (19) becomes

\[ kr_1' = a - b - c - d. \quad (20) \]

Hence, the retarded vector is

\[ e^{-ikr_1'} = e^{i(-a-b+c+d)} = e^{-i\phi} e^{i(b+c+d)}. \quad (21) \]
INCREMENT OF THE ELECTRIC-FIELD INTENSITY
FOR THE "PHYSICAL-CONE" ELEMENT

All of the parameters have now been defined for the "physical-cone" element.

Let
\[ E_0' = \text{the electric-field intensity due to the "physical cone"} \]
\[ E_0'' = \text{the electric-field intensity due to the "image cone"} \]
\[ E_0 = E_0' + E_0'' = \text{the total electric-field intensity due to the cone antenna.} \]

Substituting from (9) for \( I_0 \); (17) for \( \sin \tau \); (21) for \( e^{-jkr} \) and (15) for \( f(\delta r) \) in (1), the result is

\[ \delta(E_0') = \frac{\eta}{4\pi r} \left[ \frac{2\pi}{\lambda} \left( I + \frac{k}{r} \right) \frac{\delta \phi}{2\pi} \right] \left[ \cos \theta \cos \phi - \sin \theta \tan \Delta \right] \sqrt{\cos^2 \phi + \tan^2 \Delta} \cdot \left[ e^{-jkr} \right] [\delta \rho \sqrt{\cos^2 \phi + \tan^2 \Delta}]. \]

Let
\[ e = \frac{2\pi}{\lambda} \]
\[ f = \frac{2\pi k}{\lambda \cos \Delta} \]

Then (22) becomes

\[ \delta(E_0') = \frac{30j e^{-i\theta} I}{r_1 \lambda} \cdot \sin (e + f) \cdot e^{i(b+c+d)} (\cos \theta \cos \phi - \sin \theta \tan \Delta) \delta \rho \phi. \]  

"IMAGE-CONE" GENERAL DISCUSSION

The vertical plane electric-field intensity distribution due to the "image cone" will now be determined.

It should be remembered that the ground was assumed to be a flat perfect conductor. The current term \( I_0 \) and the \( X'Z' \) component \( f(\delta r) \) are both the same for the "physical-cone" element and the "image-cone" element.

\( I_0 \) as given in (9) and \( f(\delta r) \) as given in (15) may, therefore, be used for these two terms in determining \( E_0'' \). The development for the "physical cone" may be used up through (15) for the "image cone."

\[ \sin \tau \text{ FOR "IMAGE-CONE" ELEMENT} \]

By a process similar to that used in deriving (17)

\[ \sin \tau = -\frac{\cos \theta \cos \phi + \sin \theta \tan \Delta}{\sqrt{\cos^2 \phi + \tan^2 \Delta}}. \]

"IMAGE-CONE" ELEMENT RETARDED VECTOR

\[ e^{-jk\tau'} = e^{i(b+c+d)} = e^{-j\delta \rho \phi}. \]

INCREMENT OF THE ELECTRIC-FIELD INTENSITY
FOR THE "IMAGE-CONE" ELEMENT

Substituting from (9) for \( I_0 \); (24) for \( \sin \tau \); (25) for \( e^{-jk\tau'} \) and (15) for \( f(\delta r) \) in (1) and after some manipulation

\[ \delta(E_0'') = \frac{30j e^{-i\theta} I}{r_2 \lambda} \cdot \sin (e + f) \cdot e^{i(b+c+d)} (\cos \theta \cos \phi - \sin \theta \tan \Delta) \delta \rho \phi. \]

INCREMENT OF THE ELECTRIC-FIELD INTENSITY
FOR THE CONE ELEMENT

The increment of the total electric-field intensity for the cone-antenna element is the sum of (23) and (26).

The distance \( r_1 \) in the denominator of (23) and \( r_2 \) in the denominator of (26) may be set equal to \( r_0 \) with negligible error.

The term \( e^{-i\theta} \) in (23) and (26) is a constant phase function and may be set equal to unity.

Therefore

\[ \frac{30j e^{-i\theta} I}{r_1 \lambda} \cdot \frac{30j I}{r_2 \lambda} = \frac{30j I}{r_0 \lambda}. \]

Let
\[ E = \frac{30j I}{r_0 \lambda}. \]  

Then combining (23) and (26), substituting (27) for its equivalent and rearranging

\[ \delta(E_0) = E \sin (e + f) \left\{ \cos \theta \cos \phi e^{i(b+c)} - e^{-i(b+c)} - \sin \theta \tan \Delta e^{i(b+c)} \right\} \delta \rho \phi. \]

Equation (28) is the increment of the electric-field intensity for the cone-antenna element.

THE INTERNAL EQUATION

Indicating the integration of (28) and setting the limits of integration

\[ E_0 = E \int_0^R \int_0^{2\pi} \sin (e + f) \left\{ \cos \theta \cos \phi e^{i(b+c)}(r \sin \phi \cos \phi) + e^{-i(b+c)} \right\} \cdot \left\{ e^{i(b+c)} - e^{-i(b+c)} \right\} \sin \theta \tan \Delta e^{i(b+c)}(r \sin \phi \cos \phi) \sin (e + f) \delta \phi dr. \]

INTEGRATION IN RESPECT TO \( \phi \)

The solution of the first term containing \( \phi \) is\[ e^{i(b+c)} \int_0^{2\pi} \cos \phi e^{i(b+c)}(r \sin \phi \cos \phi) d\phi = \int_0^{2\pi} \cos \phi e^{i(b+c)}(r \sin \phi \cos \phi) d\phi = j2\pi I_1 \left( \frac{2\pi}{r_0} \sin \theta \right) \]

The solution of the second term containing \( \phi \) is\[ e^{-i(b+c)} \int_0^{2\pi} \cos \phi e^{-i(b+c)}(r \sin \phi \cos \phi) d\phi = \int_0^{2\pi} \cos \phi e^{-i(b+c)}(r \sin \phi \cos \phi) d\phi = j2\pi I_1 \left( \frac{2\pi}{r_0} \sin \theta \right) \]


12 See example (19) p. 43 of footnote reference 12.
\[
\int_0^{2\pi} e^{j(2\pi\lambda)(r \sin \phi \cos \phi)} d\phi = 2\pi j_0 \left( \frac{2\pi}{\lambda} \sin \theta \right)
\]  
(31)

Substituting (30) and (31) in (29)

\[
E_0 = E \int_0^R \sin (\theta + \phi) \left\{ \cos \theta \left[ j 2\pi j_1 \left( \frac{2\pi}{\lambda} \sin \theta \right) \right] - \sin \theta \tan \Delta \left[ 2\pi J_0 \left( \frac{2\pi}{\lambda} \sin \theta \right) \right] \right\} d\rho.
\]

After manipulation and rearrangement

\[
E_0 = \frac{30ij}{r_0} \int_R^0 \left\{ -2\pi \cos \theta J_1(qr/\lambda) \right\} \sin A \sin p r/\lambda + \sin B \sin p r/\lambda - \cos A \sin p r/\lambda + \cos B \cos p r/\lambda + 2j \left[ j \pi \sin \theta \tan \Delta J_0(qr/\lambda) \right] \sin A \cos p r/\lambda - \sin B \cos p r/\lambda + \cos A \sin p r/\lambda + \cos B \sin p r/\lambda \left\} \frac{d\rho}{\lambda}
\]

(33)

where

\[
A = (b + e) = \frac{2\pi}{\lambda} \left[ (l_1 - K \tan \Delta) \cos \theta + L \right]
\]

\[
B = (b - e) = \frac{2\pi}{\lambda} \left[ (l_1 - K \tan \Delta) \cos \theta - L \right]
\]

\[
P_1 = \frac{f + c}{r/\lambda} = \frac{2\pi(k_1 + \sin \Delta \cos \theta)}{\cos \Delta}
\]

\[
P_2 = \frac{f - c}{r/\lambda} = \frac{2\pi(k_1 - \sin \Delta \cos \theta)}{\cos \Delta}
\]

\[
q = 2\pi \sin \theta.
\]

In (33) the electric-field strength \( E_0 \) is expressed in volts per meter at a distance of 1 meter for 1 loop-ampere.

Converting to millivolts per meter at 1 mile for 1 loop-ampere

\[
E_0 = -\frac{30ij}{r_0} = -18.64j.
\]

(34)

For convenience the variable will be changed. Let

\[
z = \frac{qr}{\lambda} = \frac{2\pi \sin \theta}{\lambda}.
\]

Substituting

\[
E_0 = -j18.64j \int_0^Z \left\{ \cot \theta \sin A \sin n z J_1(z) + \sin B \sin n z J_1(z) - \cos A \sin n z J_1(z) + \cos B \cos n z J_1(z) \right\} dz \text{ mv/m/mi/loop-ampere} (35)
\]

where

\[
n_1 = \frac{k_1 + \sin \Delta \cos \theta}{\sin \theta \cos \Delta}
\]

and

\[
n_2 = \frac{k_1 - \sin \Delta \cos \theta}{\sin \theta \cos \Delta}
\]

The expression "mv/m/mi/loop ampere" is used to denote millivolts per meter at a distance of 1 mile for 1 loop-ampere.

The expressions containing \( z \) in (35) may be integrated by several methods. For each expression, either the circular or the Bessel function or both may be expanded and the result integrated term by term. The expressions may also be integrated by parts, considering either the circular or Bessel function first. The methods all lead to an answer that is a series of series and, therefore, numerical computation is long and tedious.

Solutions derived for the author by Smith\(^\text{14}\) yield answers that make the burden of numerical computation much less than it is with any of the author's derivations.

Following Smith's solutions

\[
\int \sin n z J_0(z) dz = \frac{-2}{\sqrt{\lambda^2 - 1}} \left\{ F(n, z) \cos nz + \phi(n, z) \sin nz \right\}
\]

(36)

\[
\int \cos n z J_0(z) dz = \frac{2}{\sqrt{\lambda^2 - 1}} \left\{ F(n, z) \sin nz - \phi(n, z) \cos nz \right\}
\]

(37)

\[
\int \sin n z J_1(z) dz = \frac{2n}{\sqrt{\lambda^2 - 1}} \left\{ F(n, z) \sin nz - \phi(n, z) \cos nz - \sin nz J_0(z) \right\}
\]

(38)

\[
\int \cos n z J_1(z) dz = \frac{2n}{\sqrt{\lambda^2 - 1}} \left\{ F(n, z) \cos nz + \phi(n, z) \sin nz \right\} - \cos nz J_0(z)
\]

(39)

where

\[
F(n, z) = 1/2 J_0(z) - k_n^2 J_2(z) + k_n^4 J_4(z) \cdots
\]

\[
\phi(n, z) = k_n J_1(z) - k_n^3 J_3(z) + k_n^5 J_5(z) \cdots
\]

\[
k_n = n - \sqrt{n^2 - 1}.
\]

Taking \( n \) equal to \( n_1 \) or \( n_2 \), as the case may be, (35) may be evaluated by using (36) through (39).

\(^\text{14}\) The author is indebted to Prof. V. G. Smith, Department of Electrical Engineering, University of Toronto, Toronto, Ontario, Canada, for the derivation of the integrals in (35).
Diffraction of High-Frequency Radio Waves Around the Earth*

M. D. ROCCO† AND J. B. SMYTH‡

Summary—This paper reports nonoptical height-gain measurements on a remarkably uniform desert link at frequencies of 25, 63, 170, 520, 1,000, 3,300, and 9,375 Mc using horizontal polarization. It has been shown that when the index of refraction of the atmosphere is a linear function of elevation, the problem of refraction and diffraction may be represented approximately by a problem in diffraction alone. Data taken under linear and slightly nonlinear index of refraction gradients are compared with fields predicted by standard diffraction theory. A mechanism other than diffraction is required to explain the higher-frequency fields observed under standard meteorological conditions.

I. Introduction

THERE ARE FEW PROBLEMS which have received the theoretical attention that has been given the investigation of the propagation of radio waves around the earth, and yet, quantitative data sufficiently accurate to compare with theory are very meager at the higher frequencies. It is not expected that the ionosphere will affect the higher-frequency fields near the earth, only diffraction around the earth together with reflection, refraction, absorption, and scattering in the lower troposphere must be considered in explaining nonoptical fields. This paper compares experimental data with diffraction theory, particular interest being centered on investigating the validity of replacing the problem in diffraction and refraction by one in diffraction alone for: (a) standard meteorological conditions, and (b) nonstandard meteorological conditions. The data presented are for horizontal polarization. However, numerous checks showed identical results for vertical polarization.

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† U.S. Navy Electronics Laboratory, San Diego 52, Calif.
¶ If the index of refraction near the earth is a linear function of elevation, decreasing at the rate of 1.18 × 10⁻⁴ per foot, the refracting property of the atmosphere is defined as standard. Refraction is taken into account in the diffraction problem by increasing the earth's radius by a factor of k. For the standard atmosphere, k = 1.33.

II. Experimental Site

The data presented in this paper were obtained at the Navy Electronics Laboratory's experimental station located in the Gila Valley in southern Arizona. Fig. 1 is a contour map of the region and shows the location of
the three 200-foot towers on which elevator cabs were installed for hoisting radio antennas and associated equipment. The insert in Fig. 1 is a profile of the radio link, the vertical scale is indicated by the 200-foot tower markers. Transmitters were located at Gila Bend airfield and receivers at Datelan, 47 miles west along the valley. Fig. 2 is a photographic view of the path taken from atop the Gila Bend tower.

The receiver cab made continuous excursions measuring the vertical field distribution associated with successive transmitter cab elevations of 1, 25, 50, 100, and 190 feet. About 15 minutes was required to take a complete set of data which included a height-gain run for each of the 5 transmitter cab locations. Measurements were made over a number of 24-hour periods. The meteorology of the lower atmosphere in the region is characterized by a large diurnal variation in the temperature of the air near the ground. Fig. 3 gives a time sequence of index of refraction\(^a\) versus elevation covering a typical 24-hour period. Each curve is the average of data taken at 6 stations approximately equally spaced along the propagation path. In the daytime, the index of refraction increases with elevation through the lower 6 feet of the atmosphere, producing a substandard condition in this region.

\section*{III. Standard Propagation}\(^b\)

The measured fields are plotted in decibels versus the receiver height multiplied by the wavelength raised to the minus two-thirds power. This is a convenient representation which allows the data to be shown on one graph for direct comparison with diffraction theory. Fig. 4 is a typical height-gain run for the transmitter at 190 feet. The 4 plots show the same data compared with theory where different values of \(k\) have been used. This does not affect the shape of the height-gain; the only change is the reference level for the various frequencies. If a sphere is drawn through the base of the towers at Gila Bend and Datelan, and the highest point at Sentinel, the value of \(k\) appropriate to this geometry and the standard atmosphere would be 1.10. It will be noted that the majority of the data agree best with theory where a value of \(k = 1.33\) is chosen,\(^b\) possibly showing


\(^{b}\) Free-space fields (i.e., attenuation as inverse distance) for the transmitter at 190 feet are obtained by subtracting the following decibels:

\begin{center}
<table>
<thead>
<tr>
<th>frequency</th>
<th>170</th>
<th>520</th>
<th>1,000</th>
<th>3,000</th>
<th>9,375</th>
</tr>
</thead>
<tbody>
<tr>
<td>db</td>
<td>57.4</td>
<td>64.7</td>
<td>71.7</td>
<td>89.2</td>
<td>113.2</td>
</tr>
</tbody>
</table>
\end{center}
Fig. 4—Observed fields compared with theory (parameter $k$ varied as specified).

Fig. 5—The effect of local terrain near the link terminals.

Fig. 6—The effect of transmitter elevation on nonoptical fields.

Fig. 7—25- and 63-Mc fields compared with theory.
that the earth's curvature; i.e., the local topography, near the transmitter and receiver is more important than the small hill at Sentinel, which is the only portion of the path that deviates much from a sphere with the earth's radius (Fig. 1).

Fig. 5 shows the local effect in another way. Height-gain data obtained by placing the receiver at 190 feet and lowering the transmitter cab are compared with data taken in the usual manner. At the higher frequencies, there is a marked difference between the data taken by the two methods.

Height-gain data for the various transmitter heights are shown in Fig. 6. As the antenna elevations are decreased, the agreement with theory becomes poorer the higher the frequency. When both the receiving and transmitting antennas approach the ground, the higher frequency fields fluctuate and the gain in field with height disappears. Both of these facts are at variance with predictions based on standard diffraction theory and will be discussed more fully in the last section.

IV. Nonstandard Propagation

In reality, the so-called standard index gradient is the anomaly. The meteorological situation which is standard for one frequency may markedly affect nonoptical fields\(^\text{10,11}\) on other frequencies. A comparison of the lowest with the highest frequency signals in Fig. 6 clearly shows this effect. The 25- and 63-Mc data presented in Fig. 7 show no consistent time variation which can be attributed to the low-level modification of the air.\(^\text{12}\)

Fig. 8 shows how the nonoptical fields on the higher frequencies are affected by various degrees of modification of the refractive index gradient near the earth's surface (Fig. 3). The data presented here might be interpreted as confirmation for the waveguide type of propagation.\(^\text{13,14}\) Such an interpretation of the data is being investigated; the present paper is limited to a description of the data by considering the problem as one in diffraction alone. If a value of \(k=1.81\) is chosen, the data taken at 1808 and 2313 with the transmitter at 190 feet are moderately well predicted by diffraction alone, as demonstrated in Figs. 9 and 10.\(^\text{15}\) This value of \(k\)


\(^\text{12}\) Free-space fields for a transmitter elevation of 190 feet are obtained by subtracting 52.4 db and 54.6 db, respectively, from the 25- and 63-Mc levels.


\(^\text{15}\) Free-space fields for a transmitter elevation at 190 feet are obtained by subtracting the following decibels:

<table>
<thead>
<tr>
<th>Frequency</th>
<th>170</th>
<th>200</th>
<th>1000</th>
<th>3,300</th>
<th>9,375</th>
</tr>
</thead>
<tbody>
<tr>
<td>db</td>
<td>53.4</td>
<td>58.7</td>
<td>61.5</td>
<td>78.2</td>
<td>94.2</td>
</tr>
</tbody>
</table>

Fig. 8—The effect of low-level modification of the air on nonoptical fields. Transmitter cab height = 190 feet.

Fig. 9—Diffraction fields under nonstandard meteorological conditions \(k=1.81\).
makes the 1,000-Mc height-gain data taken at 2313, with the transmitter cab at 190 feet, approach closely the theoretical curve. The comparison of experience with theory under these conditions is certainly no less favorable than the similar comparison under meteorological conditions approaching standard.

Fig. 11 is an actual height-gain record taken under a meteorological condition approaching standard. Fig. 12 is a similar record obtained during nonstandard conditions.

VI. CONCLUSIONS

The lower frequency fields received on the nonoptical link under the meteorological conditions encountered, are amply described by diffraction alone.

The fields observed at the higher frequencies for low terminal heights were considerably stronger than standard diffraction theory would predict, even under meteorological conditions which were definitely substandard near the ground. These strong fields were characterized by rapid time variations and no apparent variation with height. The frequency distribution of the signal fluctuations agrees with a Rayleigh distribution.

Fig. 15—Frequency distribution of 3,300-Mc pulse signal levels.
Fig. 11—Height-gain record under standard meteorological conditions.
Fig. 12—Height-gain record under nonstandard meteorological conditions.
that the earth's curvature; i.e., the local topography, near the transmitter and receiver is more important than the small hill at Sentinel, which is the only portion of the path that deviates much from a sphere with the earth's radius (Fig. 1).

Fig. 5 shows the local effect in another way. Height-gain data obtained by placing the receiver at 190 feet and lowering the transmitter cab are compared with data taken in the usual manner. At the higher frequencies, there is a marked difference between the data taken by the two methods.

Height-gain data for the various transmitter heights are shown in Fig. 6. As the antenna elevations are decreased, the agreement with theory becomes poorer the higher the frequency. When both the receiving and transmitting antennas approach the ground, the higher frequency fields fluctuate and the gain in field with height disappears. Both of these facts are at variance with predictions based on standard diffraction theory and will be discussed more fully in the last section.

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In reality, the so-called standard index gradient is the anomaly. The meteorological situation which is standard for one frequency may markedly affect nonoptical fields19,20 on other frequencies. A comparison of the lowest with the highest frequency signals in Fig. 6 clearly shows this effect. The 25- and 63-Mc data presented in Fig. 7 show no consistent time variation which can be attributed to the low-level modification of the air.12

Fig. 8 shows how the nonoptical fields on the higher frequencies are affected by various degrees of modification of the refractive index gradient near the earth's surface (Fig. 3). The data presented here might be interpreted as confirmation for the waveguide type of propagation.13,14 Such an interpretation of the data is being investigated; the present paper is limited to a description of the data by considering the problem as one in diffraction alone. If a value of \( k = 1.81 \) is chosen, the data taken at 1808 and 2313 with the transmitter at 190 feet are moderately well predicted by diffraction alone, as demonstrated in Figs. 9 and 10.15 This value of \( k \)

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23 Free-space fields for a transmitter elevation at 190 feet are obtained by subtracting the following decibels:

<table>
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<tr>
<th>Frequency</th>
<th>170</th>
<th>520</th>
<th>1,000</th>
<th>3,300</th>
<th>9.375</th>
</tr>
</thead>
<tbody>
<tr>
<td>Decibel</td>
<td>53.4</td>
<td>58.7</td>
<td>62.7</td>
<td>78.2</td>
<td>94.2</td>
</tr>
</tbody>
</table>

Fig. 9—Diffracted fields under nonstandard meteorological conditions (\( k = 1.81 \).
makes the 1,000-Mc height-gain data taken at 2313, with the transmitter cab at 190 feet, approach closely the theoretical curve. The comparison of experience with theory under these conditions is certainly no less favorable than the similar comparison under meteorological conditions approaching standard.

Fig. 11 is an actual height-gain record taken under a meteorological condition approaching standard. Fig. 12 is a similar record obtained during nonstandard conditions.

V. Signal Fluctuations

Nonoptical fields recorded during the daytime (meteorological conditions approaching standard) yield moderately steady signals for the lower frequencies at all heights and for the higher frequencies when the antennas are high. Figs. 13 and 14 are photographs of the recording tape showing the signal fluctuations at various receiver elevations under standard and nonstandard conditions. During the daytime, the maximum signals observed at the frequencies of 9,375 and 3,300 Mc for low antenna heights are well above the fields associated with diffraction under standard conditions. With the transmitter at 1 foot, the 9,375-Mc field is stronger, relative to the free-space level, than the 3,300-Mc field; the index of refraction gradient during this time is substandard near the earth.

Scattering in the lower atmosphere would appear to be a plausible process that might produce variable nonoptical fields on the higher frequencies which are stronger than can be accounted for by diffraction. In order to make a detailed investigation of the signal fluctuations, the 3,300-Mc pulses displayed on an oscilloscope were recorded photographically. Fig. 15 is a frequency distribution of received pulses compared with a Rayleigh distribution. The chi-squared test for goodness of fit was near the 20 per cent level, possibly indicating that the observed nonoptical signal might be produced by incoherent scattering in the lower atmosphere.

The authors are indebted to W. C. Hoffman and R. F. Arenz for the analysis leading to Fig. 15.

Fig. 15—Frequency distribution of 3,300-Mc pulse signal levels.

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The lower frequency fields received on the nonoptical link under the meteorological conditions encountered, are amply described by diffraction alone.

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Fig. 11—Height-gain record under standard meteorological conditions.
Fig. 12—Height-gain record under nonstandard meteorological conditions.
Fig. 13—Time variation in received signal (at elevation indicated) under standard meteorological conditions. Elevation of transmitting antennas = 190 feet. Time: 5:30 p.m.
Fig. 14—Time variation in received signal (at elevation indicated) under nonstandard meteorological conditions. Elevation of transmitting antennas = 190 feet. Time: 7:30 a.m.
Calculator and Chart for Feedback Problems

JEAN H. FELKER†, ASSOCIATE, IRE

Summary—This paper describes a chart and a mechanization of the chart in the form of a calculator that will simplify the calculation of the effects of feedback. When the magnitude and phase of the feedback are put into the calculator, figures are obtained at each frequency point for the magnitude and phase shift of the transfer characteristic with the feedback loop closed.

1. The Mu-Beta Effect

The chart and calculator are based on the "mu-beta" effect, which is defined by use of the fundamental feedback equation:

\[ G = \frac{\mu}{1 - \mu\beta} \]  (1)

where

- \( \mu \) = the voltage amplification of the amplifier with the feedback loop open
- \( \beta \) = the fraction of the output voltage of the amplifier that is fed back to the input
- \( G \) = voltage amplification of the amplifier with the feedback loop closed.

Equation (1) can be rewritten in the form

\[ G = \frac{1}{\beta} \left( \frac{-\mu\beta}{1 - \mu\beta} \right) \]  (2)

from which it is seen that, when \( \mu\beta \) is large, the response is given approximately by

\[ G \approx -\frac{1}{\beta} \]

and is independent of the \( \mu \) or gain circuit. Since amplifying devices always have a finite bandwidth, there will be frequencies at which \( \mu\beta \) is not large, and at these frequencies account must be taken of the second factor of (2), which can be anything from 0 to \( \infty \).

The departure of the amplifier response from \( -1/\beta \) has been described as the "mu-beta" effect, because it results from the noninfiniteness of \( \mu\beta \). Accordingly, the \( \mu\beta \) effect is defined as a complex number

\[ \gamma \phi = \frac{-\mu\beta}{1 - \mu\beta} \]

\[ \gamma = \text{modulus or magnitude of mu-beta effect} \]
\[ \phi = \text{argument or phase of mu-beta effect} \]

Substitution of (3) in (2) gives

\[ G = \frac{1}{\beta} \left[ \frac{\gamma + \phi - \psi}{180^\circ} \right] \]  (4)

from which it is seen that the magnitude of \( G \) in decibels \( \gamma \) is \( \gamma \) in decibels minus the magnitude of \( \beta \) in decibels, while the phase of \( G \) is \( 180^\circ \) plus \( \phi \) minus \( \psi \) (the phase of \( \beta \)).

A plane can be visualized in which the vector from the origin to any point represents the \( \mu\beta \) effect at that point. For a given \( \mu\beta \) effect there can correspond only one value of \( \mu\beta \). One might label each point with the corresponding values of \( |\mu\beta| \) and \( \theta \), or one might plot contours of constant \( |\mu\beta| \) and \( \theta \) on such a plane. Such a representation would be useful when \( \gamma < 1 \), since this region is confined within a unit circle. The lower half of such a plane is shown in the lower half of Fig. 1.

For values of \( \gamma \) between 1 and \( \infty \), an inverse plane might be imagined in which the distance from the origin represents the inverse of the \( \mu\beta \) effect. The mirror image of such a plane is shown in the top half of Fig. 1.

II. Use of the Calculator

The chart of Fig. 1 can be made into a calculator by cutting out the scale at the bottom of the figure and attaching it at the center of the chart as a rotatable arm. To find the \( \mu\beta \) effect corresponding to a particular pair of values for \( \gamma \) and \( \theta \), first locate the intersection of the appropriate \( |\mu\beta| \) and \( \theta \) contours, and then rotate the calculator arm until the edge of the scale on it crosses the intersection. Read the magnitude \( (\gamma) \) of the \( \mu\beta \) effect off the scale on the arm, and the phase \( (\phi) \) at the point where the arm crosses the semicircular phase scale at the top of the calculator.

Since the calculator shows only one-half of the direct and inverse \( \mu\beta \)-effect planes, only angles between 0 and \( 180^\circ \) can be represented. This represents no real difficulty, since any angle greater than \( 180^\circ \) can be represented as a negative angle less than \( 180^\circ \). To permit the use of the calculator for both negative and positive angles, no signs have been attached to the \( \theta \) contours. The only rule that need be remembered to avoid angle

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

The term "in decibels" is employed herein to refer to 20 log of the absolute value of a quantity. (The quantities dealt with are ratios of linear quantities, voltages or currents, rather than power.)

3 This calculator is commercially available as advertised in this issue.
ambiguity is that the sign of $\phi$ will always be the negative of $\theta$. If, for example, $|\mu| = 10$ db and $\theta$ is $30^\circ$, $\gamma$ is $+2.6$ db and $\phi$ is $-12^\circ$, while if $\theta$ were $-30^\circ$ (or $330^\circ$), $\phi$ would be $+12^\circ$.

Note that the positive half of the calculator arm has been made longer than the negative half, with the result that the phase of the $\mu\beta$ effect can be determined only when the positive half of the arm is in the top half of the calculator. Thus no special effort is required to ensure that the correct sign is attached to the decibel value of $\gamma$.

### III. A Sample Problem

The characteristic of the $\mu\beta$ loop of a servo is shown in Fig. 2. Since $\beta$ in a servo is $-1$, the $\mu\beta$ effect is the...
The High-Frequency Response of Cylindrical Diodes*

EDWARD H. GAMBLE†

This abstract includes the results of an investigation into the response of cylindrical diodes to the application of voltages which contain high-frequency components. In particular, the response to "large" signals has been determined, where the term "large signal" refers to the condition that the amplitudes of the time-varying components are comparable to that of the constant polarizing voltage.

The problem has been studied as a quasi-stationary one since the applied signal may be assumed to change only slightly in the time required for an electromagnetic wave to propagate the interelectrode space. The three basic postulates upon which this theory is built are Poisson's equation, Newton's equation, and the continuity equation.

The complete solution of the cylindrical diode under general space-charge conditions is found to be more complicated than the associated planar case. The method of integral equations has been adopted. The solution for the trajectory of the electrons or the radial distance \( R(t, r) \) is the most difficult problem since the integral equation for \( R(t, r) \) involves the reciprocal of \( R(t, r) \) under the integral sign. A method of successive approximations has been adapted for each specific form of applied voltage. In general, only approximate solutions can be found for the cylindrical diode under various space-charge conditions. Nevertheless, the derived relationships between the fundamental physical quantities do afford one insight into the inner workings of the cylindrical thermionic diode.

The simplifying assumptions which must be made to reduce the general space-charge relations to those which describe the space-charge-limited operation and temperature-limited operation have been pointed out.

Numerical solutions have been determined for a wide variety of signal voltages for both space-charge-limited and temperature-limited operations. These should prove of considerable help to the tube designer. The variations in the important physical variables which occur as the amplitude of the voltage signal increases from that referred to as "small" to the "large" signals are determined and graphically shown. These numerical solutions check the integral equation relationships to an accuracy of about 2 per cent.

Under large signal conditions, considerable modulation of the electric field, and the velocity and trajectory of the electrons occur. In some cases, the electrons are returned to the cathode or caused to oscillate radially in the interelectrode space.

When the radii of both the cathode and anode are large, the solutions reduce to those which are found for the planar diode configuration.

The first approximation for the numerical solutions was determined by setting the nonlinear differential equation of motion of the electrons into an electronic analogue computer. Analytic formulas which matched the numerical solutions were determined. These analytic expressions were substituted for the variables in the integral equations and found to satisfy them within the accuracy of the numerical solution.
Response of $RC$ Circuits to Multiple Pulses*  

Daniel Levine†, Member, IRE

Summary—The voltage across the capacitor in a pulsed $RC$ circuit can be found without difficulty, if the change in the base-line is known as a function of the number of pulses. This paper derives the necessary equations for this transient effect, using only the well-known charge discharge functions for a direct-current circuit. The results can be extended immediately to the voltage across the resistor.

The voltage variation across a component in a resistor-capacitor circuit upon applying a rectangular voltage wave is treated adequately in many textbooks; however, the effect of applying several pulses in succession generally is left as an exercise for step-by-step solution. This paper generalizes the step-by-step process for any number of pulses.

The case for $n$ pulses, where $n$ may have any value, is of practical concern in some circuit design, and may be solved exactly by finding the voltage before the start of the $n$th pulse. An equivalent statement is that we want to find functions that describe the maximum and minimum envelopes for a finite number of pulses.

Initially, we shall consider the voltage variation across the capacitor in Fig. 1. Starting the rectangular wave at time $t=0$, when the capacitor has a voltage denoted by $e_c(0)$, causes a change in voltage of

$$\Delta e_c(t) = [E_m - e_c(0)](1 - e^{-1/(T_1)},$$

where $T_1 = R_1C_1$, and $E_m$ is the magnitude of the rectangular wave. The total voltage across the capacitor, therefore, is

$$e_c(t) = e_c(0) + [E_m - e_c(0)](1 - e^{-1/(T_1)}). \tag{1a}$$

At the end of the pulse the time is equal to the pulse duration $T_1$, so that the voltage is

$$e_c(T_1) = e_c(0) + [E_m - e_c(0)](1 - e^{-1/(T_1)}), \tag{2}$$

where

$$e_c(T_1) = e_c(0) + [E_m - e_c(0)](1 - e^{-1/(T_1)}), \tag{3}$$

where

$$e_c(T_1) = e_c(0) + [E_m - e_c(0)](1 - e^{-1/(T_1)}), \tag{4}$$

where $E_m = e_c(0) + a_1[E_m - e_c(0)]$. The voltage variation for $t > T_1$ is given by

$$e_c(t) = \left[e_c(0) + a_1[E_m - e_c(0)]\right] e^{-(t - T_1)/T_2} \tag{5}$$

where $T_2 = R_2C_2$. (In many sweep generators the time constant during discharge is different from that during charge. In order to develop equations having the broadest application, this type of condition is covered by introducing $T_2$.)

The start of the second pulse is at time $t = T_2$, where $T_2$ is the pulse repetition period; the initial voltage is

$$e_c(T_2) = e_c(T_1)e^{-(T_2 - T_1)/T_2} = a_2e_c(T_1) \tag{6}$$

where

$$e_c(T_2) = e_c(T_1)e^{-(T_2 - T_1)/T_2} = a_2e_c(T_1) \tag{7}$$

In a similar manner we obtain for the second pulse interval

$$e_c(T_2 + r_1) = e_c(T_2) + a_1[E_m - e_c(T_2)] = a_1E_m + a_2[1 - a_1]e_c(T_1), \tag{8}$$

and

$$e_c(T_2 + r_2) = a_2e_c(T_2 + r_1).$$

When this procedure is extended to cover additional pulses, we find that the voltage expressions may be represented by the following progression:

<table>
<thead>
<tr>
<th>Pulse</th>
<th>Voltage at $t = (m - 1)T_2 + T_1$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>$b$</td>
</tr>
<tr>
<td>2.</td>
<td>$a + br$</td>
</tr>
<tr>
<td>3.</td>
<td>$a + (a + br)r$</td>
</tr>
<tr>
<td>4.</td>
<td>$a + (a + ar^2 + \ldots + ar^{m-1})$</td>
</tr>
<tr>
<td>5.</td>
<td>$a + ar + ar^2 + \ldots + ar^{m-1}$</td>
</tr>
</tbody>
</table>

Therefore, the voltage at the termination of the $m$th pulse is obtained by substituting the values for $a$, $b$, and $r$ (given in equations (6) and (7)) into equation (8):

$$e_c[(m - 1)T_2 + T_1] = \frac{a_1E_m}{1 - a_2[1 - a_1]} \left\{ 1 - \left[ a_2(1 - a_1) \right]^{m-1} \right\} \tag{9}$$

As $m$ becomes very large, the factor containing $r^{m-1}$ approaches zero, since the ratio factor $r$ is less than unity; therefore
Impedance Measurements with Directional Couplers and Supplementary Voltage Probe

B. PARZEN, ASSOCIATE, IRE

Summary—Impedance measurements may be made at very-high frequencies and above with the impedometer, consisting of two oppositely connected directional couplers and a voltage probe in a short transmission line. An experimental model for frequencies between 50 and 500 Mc is described.

INTRODUCTION

DIRECTIONAL COUPLERS have been used to determine the magnitude of the reflection coefficient (and hence SWR) of impedances referred to a given transmission line. The addition of a voltage probe permits the phase angle and magnitude of the reflection coefficient to be determined from which the real and imaginary components of the impedance may be obtained.

The essential components of the equipment are shown in Fig. 1. The impedance $Z_L$ to be measured is connected by a short length of transmission line to a generator, preferably through an attenuating pad. Two directional couplers are inserted anywhere in the line. One, giving an output $E_1$, is directed to the reflected wave and the other, of output $E_2$, to the forward wave. A voltage probe, giving an output $E_p$, is inserted in the line at electrical degrees $\theta_1$ from $Z_L$.

The attenuating pad maintains $E_1$ substantially independent of the load $Z_L$, and is not required if the internal impedance of the generator is equal to the characteristic impedance of the line.

\[ E_1 = \frac{a_1 E_m}{1 - a_2(1 - a_i)} \]

Since the amplitude of the first pulse is $a_1 [E_m - e_c(0)]$, the ratio of the first pulse amplitude to the steady-state amplitude is

\[ \frac{a_1 E_m - e_c(0)}{E_m(1 - a_2)} \]

The voltage across the resistor in the RC circuit does not require a separate analysis, since the sum of the resistor and the capacitor voltages is equal to the applied voltage.

The wave forms in Fig. 2 illustrate the application of these equations to the problem of a square wave applied to a circuit having no initial charges, and a time constant 0.721 times the period.
at any point \(\alpha\) on the line between the reflection coefficient \(K_\alpha\) and the normalized impedance \(z\), equal to the actual impedance \(Z_\alpha\) divided by the characteristic impedance of the line \(Z_0\), are given by

\[
z_\alpha = r_\alpha + jx_\alpha = \frac{1}{y_\alpha} = \frac{1}{g_\alpha + jb_\alpha} = \frac{1 + K_\alpha}{1 - K_\alpha},
\]

(1)

\[
K_\alpha = |K_\alpha e^{j\alpha}| = \frac{z_\alpha - 1}{z_\alpha + 1} = \frac{1 - y_\alpha}{1 + y_\alpha}.
\]

(2)

From well-known transmission-line and directional-coupler theory, we have

\[
E_1 = C_1E_2(1 = K_1),
\]

(3a)

\[
E_2 = C_2E_3K_2,
\]

(3b)

\[
E_3 = C_3E_4,
\]

where \(C_1\), \(C_2\), and \(C_3\) are constants and \(E_i\) indicates the forward components of the voltage wave in the transmission line.

After making \(C_1\), \(C_2\), and \(C_3\) equal to each other, we obtain from (2) and (3)

\[
|e_1| = \frac{|E_1|}{|E_3|} = |1 + K |
\]

\[
= |1 + |K| \cos \delta_1 + j|K| \sin \delta_1|,
\]

(4)

\[
|e_2| = \frac{|E_2|}{|E_3|} = |K|.
\]

(5)

From (1), (2), (4), and (5) result

\[
\cos \delta_1 = \frac{|e_1|^2 - |e_2|^2 - 1}{2|e_2|},
\]

(6)

\[
1 = y_\alpha = g_\alpha + jb_\alpha = \frac{1 - K}{1 + K} = \frac{1 - |K|^2 - 2j|K| \sin \delta_a}{1 + |K|^2 + 2|K| \cos \delta_a},
\]

(7)

\[
|e_1|^2 = |1 + K_1|^2 = \frac{4}{(1 + g_1)^2 + b_1^2},
\]

(8)

\[
|e_2|^2 = |K|^2 = 1 - \frac{4g_1}{(1 + g_1)^2 + b_1^2}.
\]

(9)

From (5) and (6), the magnitude \(|K|\) and phase angle \(\delta_\alpha\) of the reflection coefficient at the point of the voltage probe may be calculated. We can calculate the angle \(\delta_L\) at the load from the relation

\[
\delta_L = \delta_1 + 2\theta_1.
\]

(10)

If \(2\theta_1\) is not known, it may be found easily as follows:

1. Short the end of the line for which load \(|e_2|=1\) and \(\delta_L=180^\circ\), and read \(e_1\).
2. Calculate \(\delta_1\) from (6), and
3. \(2\theta_1=180-\delta_1\).

Equation (6) is incomplete in that it permits the determination of only the magnitude of \(\delta_1\) and not its sign since, in general, \(\cos \delta=\cos (-\delta)\). An additional component is, therefore, added to obtain the sign of \(\delta_1\). For convenience, the component consists of a capacitive susceptance \(jb\) at the point of the voltage probe. The susceptance is normally out of the line and is inserted into the line and its effect noted after \(E_1\), \(E_2\), and \(E_3\) have been measured. Equations (7), (8), and (9) show that \(|e_1|\) increases and \(|e_2|\) decreases when \(\delta_1\) is positive and conversely when \(\delta_1\) is negative. Thus, by noting the change in \(|e_1|\) and/or \(|e_2|\), the sign of \(\delta_1\) is determined. Equations (8) and (9) also indicate that the effective change in \(|e_1|\) is greater when the reflection coefficient is large and the change in \(|e_2|\) is greater when the reflection coefficient is small. In every case, the change in either \(|e_1|\) or \(|e_2|\) will be sufficient to fix the sign of \(\delta_1\) positively.

The procedure to be followed in measuring an impedance is (1) connect the unknown impedance to the equipment, (2) read \(E_1\), \(E_2\), and \(E_3\), (3) find the sign of \(\delta\), by inserting the susceptance and noting the change in readings, (4) calculate \(|K|\) from (5) and \(\delta_1\) from (6), (5) add \(\delta_1\) to \(2\theta_1\) and obtain \(\delta_L\) (if \(2\theta_1\) is not known, it is found as described above), (6) calculate \(e\) from (7), using Smith or equivalent impedance charts.

To minimize labor in calculating \(\delta_1\) from (6), that equation has been plotted in Fig. 2. It may be noted that (6) is but an expression of the cosine law for triangles.

This method is practical at all frequencies, since all the components (directional couplers, voltage probe, variable capacitive susceptance, and short section of transmission line) have been built at all radio frequen-
cies. The method is particularly useful for measurements between 50 and 500 Mc in view of the absence of other good methods for this frequency region.

**Experimental Model**

A cross-sectional view of an impedometer for use from 50 to 500 Mc is shown in Fig. 3 and a photograph in Fig. 4. It consists of a short section of 1\(\frac{1}{4}\)-inch coaxial line with adapters for type \(N\) fittings. In this line, are placed two wide-band directional couplers, a capacitive voltage probe, and a variable susceptance made up of a high-capacitance probe that can be inserted by a spring-returned push button. The length of the active portion is 5 inches and the total length, including the adapters, is 15 inches.

Each coupler consists of a resistor, capacitor, and loop. They differ only in the orientation of the loop with respect to the line. Their theory of operation has been fully covered in the referenced paper.

The coupler characteristics are as follows.

**Internal Impedance**

The internal impedance is approximately 50 Ω.

**Directivity**

The directivity is approximately 40 db. By directivity is meant the value of \(|e_1|^2\) expressed in db when the line is terminated by \(Z_0\). The directivity is important as it fixes the lowest value of reflection coefficient that can be measured. Thus, a directivity of 40 db corresponds to a minimum reflection coefficient of 0.01, which is equivalent to a VSWR of 1.02. For the type of couplers shown, a directivity of 35 db and better can be readily obtained.

**Coupling**

The coupling is approximately \(-40\) db at 100 Mc. The coupling decreases 6 db as the frequency decreases by an octave. By coupling is meant the ratio in db of the power obtained from the \(E_2\) coupler to the power absorbed by the impedance being measured, when the impedance is equal to \(Z_0\). The coupling should be as small as possible, for as it increases, so do the distortion of the field within the line and the measurement errors. It will be noted that the impedometer as presently constructed has a coupling of \(-25\) db at 500 Mc. However, the coupling can be easily decreased by reducing both the inductive and capacitive couplings.

**Voltage Probe**

The voltage probe consists of a small capacitance in series with a low resistance. The degree of coupling is adjusted by moving the probe in or out to vary the capacitance. When properly adjusted, the internal impedance and attenuation characteristics of the probe are identical to those of the couplers. Thus, correct probe adjustment at any frequency insures correct adjustment at all frequencies.

**Transmission Line**

The transmission line has a characteristic impedance of 53 Ω. For maximum accuracy, the impedometer should be used without the adapters. The VSWR of the impedometer with both adapters is 1.05.

**Adjustable Susceptance**

The adjustable susceptance consists of a grounded variable-capacitance probe, which can be inserted into the line by a spring-return mechanism. On inserting the probe, both \(E_1\) and \(E_2\) change. As previously explained, the change in \(E_2\) is greater when the VSWR is low and the change in \(E_1\) is greater when the VSWR is high. In every case, the change in either \(E_1\) or \(E_2\) is sufficient to fix positively the sign of \(\delta\).

On initial use of the impedometer, the following should be checked:

- \(E_1 = E_3\) when the load is equal to \(Z_0\).
- \(E_2 = 0\) when the load is equal to \(Z_0\).
- \(E_2 = E_3\) when the load is a short circuit.

When these conditions are established, we can be certain of the correct operation of the impedometer. The impedometer possesses fair accuracy when the reflection coefficient of the impedance being measured has a magnitude between 0.02 and 0.95 at any phase angle.

**Impedance Measurements**

Impedance measurements made with the impedometer at 100 Mc have been found to be within 5 per cent
of those obtained with a slotted line. At this frequency, the required minimum length of the slotted line is 60 inches compared with the 5-inch active length of the impedometer.

The equipment arrangement illustrated in Fig. 5 and the following procedure makes indicators having precise amplitude calibrations unnecessary.

(1) A standard-signal generator having a maximum calibrated output of 100,000 µv was used. The attenuator was accurately calibrated over a range from +40 to −10 db referred to 1,000 µv. The modulation was at 400 cps.

(2) The detector was a standard communications receiver, having sufficient selectivity so that generator harmonics are not troublesome.

(3) A 10-db resistive attenuator was inserted between the generator and the impedometer to insure constancy of $E_1$ regardless of the impedance being measured.

(4) The impedance to be measured was connected to the impedometer, and the attenuator was set to 1,000 µv. The output of the receiver was noted when it was connected to the $E_1$ probe. The values of the attenuator settings required to yield the identical receiver output when connected to the $E_1$ and $E_2$ probes determine the magnitude of $E_1$ and $E_2$.

(5) The impedance is then calculated in the manner previously described. The accuracy is thus determined almost entirely by the quality of the impedometer probes and by the calibration of the signal-generator attenuator; fairly accurate measurements are, therefore, not too difficult to achieve.

Although the impedometer has been specifically designed for the frequency band of 50 to 500 Mc, it is usable at much lower frequencies, being limited only by the attenuation of the coupler, which increases 6 db per octave decrease in frequency. Where sufficiently powerful signal generators and sufficiently sensitive receivers are available, this limitation is not serious. For example, at 3 Mc, a generator having a maximum output of 1 volt and a receiver sensitivity of 5 µv will be suitable.

To summarize, the basic principles of the impedometer are applicable at all frequencies. However, the availability of good bridges at frequencies below 50 Mc, restrict its application to the higher frequencies. It is particularly useful between 50 and 500 Mc as (1) it covers a very-wide frequency band, (2) it is a low-cost instrument, (3) except for the slotted line, no other simple instrument is available, and (4) slotted lines for this frequency region are too bulky and expensive.

**Bilinear Transformations Applied to the Tuning of the Output Network of a Transmitter**

K. S. KUNZ†, ASSOCIATE, IRE

*Summary—The literature on the proper tuning of the output network of a class-C power amplifier inductively coupled to a load is surprisingly meager1 and, in general, there seems to exist considerable confusion concerning the relation between the following three conditions:

(a) Resonance of the secondary circuit

(b) Maximum-minimum point of average plate current

(c) Maximum coupled circuit efficiency.

This state of affai rs is probably due in large part to the complexity of mathematics involved in a straightforward analysis of the circuits involved. In this paper we have employed the powerful theory of bilinear transformations in a complex plane, for which circles are transformed into circles, to determine the actual value of the secondary capacitance for a maximum-minimum of average plate current and for maximum coupled circuit efficiency.

The results of this investigation show that (b) and (c) differ widely from (a)—except for the case of a series load in the secondary circuit. Moreover, although (b) may differ considerably from (c), these two conditions are indistinguishable if the $Q$ of the primary coil is large, and the coefficient of coupling is very small.

**Introduction**

The output circuit of a class-C amplifier that acts as the final stage of a transmitter (see Fig. 1) is usually considered to be properly tuned if any of the secondary is considered. Moreover, even for this case, series resonance is assumed rather than derived.

change in the plate circuit capacitor $C_1$ increases the plate current, and any change in the secondary circuit capacitor $C_2$ decreases it. In other words, one often adjusts $C_1$ for minimum and $C_2$ for maximum plate current. It should be pointed out that in Fig. 1 the load is included in the four-terminal network $A$.

![Fig. 1—Class-C amplifier inductively coupled to a load.](image)

Parker, in investigating the use of such an inductively coupled class-C amplifier for rapid determination of an unknown load impedance in the secondary circuit, noted that, contrary to what is often implied, the maximum-minimum tuning referred to above did not coincide with resonance of the secondary circuit. In fact, for a load in shunt with $C_2$, this maximum-minimum tuning was obtained when the secondary was not even in the neighborhood of resonance.

The following paper is an outcome of the author's efforts to determine theoretically the conditions for a maximum-minimum of average plate current and for maximum coupled circuit efficiency. Ordinary methods of determining maxima and minima by differentiation proved inadequate, due to the complexity of the circuits. However, the theory of bilinear transformations as applied to networks turns out to be a very powerful tool for this purpose. When adapted to this problem, it makes possible the determination of the tuning of $C_1$ and $C_2$ for a maximum-minimum condition of plate current for even the most complex secondary circuit, one in which the network $A$ of Fig. 1 is any assemblage of passive linear elements.

Since the tuning of $C_1$ that makes $Z_L$ a pure resistance also makes this impedance a maximum, it is generally agreed that, for minimum plate current, $C_1$ is tuned for resonance, i.e., $Z_L = R_L$ (pure resistance). This tuning insures that the plate voltage is least when the pulse of plate current flows through the tube. It is then only the tuning of the capacitor $C_2$ for maximum plate current that leads to serious mathematical complexity, when treated by elementary methods. In fact, we shall assume that $C_1$ is at all times adjusted for resonance, and confine our attention to finding the value of $C_2$ for maximum plate current.

It is clear that, since $C_1$ is adjusted for resonance, the tube sees only the real part of the admittance $Y_b = G_b - jB_b$; hence, $C_2$ is to be adjusted for maximum value of $G_b$. If one considers the four-terminal network $B$ (see Fig. 1), the problem may be considered as follows. As $C_2$ is varied, the load impedance for network $B$ varies along a straight line, the imaginary axis of the complex plane. The input admittance $Y_b$ as $C_2$ is varied, will also describe some path in the complex plane. It remains only to find the value of load impedance for network $B$ that corresponds to the point in the locus of $Y_b$ that is farthest to the right.

That the locus of $Y_b$ is a circle results directly from the following general theorem of network theory:

**Theorem 1:**

*If a variable impedance is applied to the output of a linear four-terminal network and the locus of the points in the complex plane representing the variation of this impedance is either a circle or a straight line, then the locus of the points in the complex plane representing the input impedance (or admittance) is a circle (or in some cases a straight line).*

In part I we shall review briefly the subject of bilinear transformations and the basis of the general network theorem stated above, and then demonstrate how the maximum-minimum tuning can be expressed in terms of the constants of network $B$. In Part II, we shall consider the three important special choices for network $A$ given in Fig. 2. In each case $R$ is the load resistance. Thus, described in the ordinary way, the three cases considered are:

(a) Series load, neglecting distributed capacitance of the secondary
(b) Parallel or shunt load
(c) Series load as influenced by the distributed capacitance $C_t$ of the secondary

Part III will include an alternative graphical treatment of the problem that has considerable merit as a conceptual aid in visualizing the requirement for maximum-minimum tuning. In Part IV we shall determine the proper tuning of $C_2$ for maximum coupled circuit efficiency and relate this tuning to the maximum-minimum tuning.

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footnotes:
- Sam E. Parker, now at the U. S. Navy Electronics Laboratory, San Diego, Calif.
- Only for a series load (see Fig. 2(a)) was this method at all applicable.
- Previous use has been made of this method for solving other problems. For example, A. Hazeltine, "Current loci in the general linear A-C network," *Elec. Eng.*, vol. 56, pp. 325-330, 1937.

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10 The statement of this theorem is simplified if one considers a straight line as a special case of a circle, namely, one with an infinite radius.
I. Bilinear Transformations and the General Conditions for Maximum-Minimum Tuning

A complex variable \( w \) is said to be a bilinear function of another complex variable \( z \) if

\[
w = \frac{Az + B}{Cz + D},
\]

where \( A, B, C, \) and \( D \) are constants (in general complex). If \( C \neq 0 \), (1) can be written

\[
w = \frac{az + b}{z + d}.
\]

If the real and imaginary parts of \( z \) are represented by the \( x \)- and \( y \)-co-ordinates, respectively, of a point in a plane called the \( z \) plane, then the possible values of \( z \) are simply the two fold infinity of points in this plane. To each such point in the \( z \) plane there corresponds a value of \( w \) given by (1), and hence a point in a second plane, which we will call the \( w \) plane.

**Lemma 1.** Any set of points lying on a circle in the \( z \) plane is transformed by a bilinear transformation into a set of points lying on a circle in the \( w \) plane, it being noted, however, that a straight line is to be considered as a special case of a circle (one with infinite radius).

Since a general proof of this lemma may be found in many standard books on the functions of a complex variable, \(^{11}\) we shall not give this proof. However, a special case of the lemma, the one actually needed in our calculations, will be proved in the Appendix.

**Lemma 2.** The input impedance (or admittance) of a linear four-terminal network is a bilinear function of the impedance (or admittance) closing the other end.

The proof of this lemma is very simple and may be found in various books on network theory. \(^{12}\)

The combination of Lemmas 1 and 2 leads at once to the proof of Theorem I.

From the discussion in the Introduction we have deduced the following conditions for maximum-minimum tuning:

1. \( C_2 \) is tuned for the maximum value of \( G_b \), where

   \( Y_b = G_b - jB_b \).

2. \( C_1 \) is tuned to cancel the imaginary part of \( Y_b \), so that the tube seen a pure conductance \( G_b \); therefore, \( \omega C_1 = B_b \).

Also, we have shown that, as a consequence of Theorem I, the locus of \( Y_b \) in the complex plane, as \( C_2 \) is varied, is a circle (see Fig. 3). The problem thus reduces to finding the point \( P \) in the \( Y_b \) plane that lies farthest to the right and the corresponding point \( Q \) in the impedance plane of \( C_2 \) in terms of the constants of the network \( B \).

Since \( Y_b \) is a bilinear function of \( z = -j/\omega C_2 \) (see (2)),

\[
Y_b = \frac{az + b}{z + d},
\]

where the complex constants \( a, b, \) and \( d \) can be determined, once network \( B \) is specified.

It is shown in the Appendix that the tuning of \( C_2 \) for maximum \( G_b \) is given by the equation

\[
\frac{1}{\omega C_2} = \frac{\text{Re} (Nd^*)}{\text{Im} (N)} = \frac{\text{Real part of } Nd^*}{\text{Imaginary part of } N},
\]

where

\[
N = \sqrt{ad - b},
\]

and \( a, b, \) and \( d \) are the constants of the bilinear transformation (3). The complex conjugate of a quantity is indicated by an asterisk. The desired tuning of \( C_2 \) for the condition of a maximum-minimum of average plate current is therefore given by (4).

II. Special Cases

In this part we shall determine the maximum-minimum tuning of \( C_2 \) for the three special cases mentioned in the Introduction. These correspond to the three choices of network \( A \), of Fig. 1, given in Fig. 2.

A. Series Load Neglecting Distributed Capacitance of Secondary

Network \( B \) is then

\[
Y_b = \frac{Y_{12} + Y_1(Z_2 + R)}{z + (Z_2 + R + \omega^2 M^2 Y_1)}
\]

hence from the theory of coupled circuits

\[
Y_b = \frac{Y_{12} + Y_1(Z_2 + R)}{z + (Z_2 + R + \omega^2 M^2 Y_1)}
\]

where

\[
Z_1 = R_1 + j\omega L_1 = R_1(1 + j\omega)
\]

\[
Z_2 = R_2 + j\omega L_2 = R_2(1 + j\omega).
\]

The constants of this bilinear transformation are (see (3) and (5))

\[
d = Z_2 + R + \omega^2 M^2 Y_1
\]

\[
N = \omega M Y_1.
\]


Substituting these constants into (4), one has
\[ \frac{1}{\omega C_2} = \omega L_2 - \frac{1}{Q_1} \left[ R + R_2 + \omega^2 M^2 \right]. \tag{9} \]
This can be alternatively expressed as
\[ \frac{1}{\omega C_2} = \omega L_2 \left[ 1 - \frac{1}{Q_1\Gamma} - k^2 \right], \tag{10} \]
where
\[ Q_1 = \frac{\omega L_2}{R_2 + R}, \tag{11} \]
\[ k = \frac{M}{\sqrt{L_1L_2}} = \text{coefficient of coupling.} \tag{12} \]

Since at resonance \(1/\omega C_2 = \omega L_2\), it is clear that the maximum-minimum tuning for \(C_1\) given by (9) or (10) differs from resonance. Note that even when the coupling is reduced to a negligible amount, the secondary still differs from resonance by the term \(-(R + R_2)/Q_1\).

For actual circuits the latter term is apt to be very small, but the term depending on the coupling will usually be quite appreciable.

The condition for resonance of the secondary circuit is decidedly different from (14). For one thing, it clearly does not depend on \(R_2\), the resistance of the secondary coil, or on the coupling \(k\). If we take these two factors to be zero, then (14) becomes \(1/\omega C_2 = \omega L_2\), which is clearly independent of \(R\). On the other hand, for resonance to occur at all, \(R\) must be greater than the critical value \(R = 2\omega L_2\), and for \(R > 2\omega L_2\), there are two resonance tunings. If we choose the one that reduces to series resonance for \(R = \infty\), then \(1/\omega C_2 = \lambda \omega L_2\), where \(\lambda\) varies from 1 at \(R = \infty\) to 2 at the critical value of \(R\).

**C. Series Load Taking into Account the Distributed Capacitance of Secondary**

Network \(B\) for this case has the form

\[ Y_b = \frac{(Z_2 + Z_d)z + Z_d(Z_d + R) + Z_dR}{(Z_1(Z_2 + Z_d) + \omega^2 M^2)z + (Z_1Z_2 + \omega^2 M^2)(Z_d + R) + Z_1Z_dR}, \]

hence

\[ Y_b = \frac{(Z_2 + R)z + RZ_2}{[Z_1(Z_2 + R) + \omega^2 M^2]z + (\omega^2 M^2 + Z_1Z_2)R}. \tag{13} \]

This is a bilinear function of \(z\) expressed in the form of (1). By dividing through by the coefficient \(C\), it takes the form of (2). On applying (4) and simplifying, one obtains the equation.

\[ \frac{1}{\omega C_2} = \omega L_2 \left[ 1 - \frac{1}{Q_1\Gamma} - k^2 \right]. \tag{14} \]

This is the maximum-minimum tuning for \(C_2\). Note that as long as the load resistance \(R\) is large, compared to the resistance of the secondary coil, the tuning of \(C_2\) is practically independent of the load.

**III. Contours of Constant \(G_b\) in the \(Z_0\)-Plane**

An alternative treatment of the problem of maximum-minimum tuning can be had by determining how the contours of constant \(G_b\) and \(B_b\) appear in the \(Z_0\) plane \([Z_0\text{ being the impedance attached to the secondary of the transformer, (Fig. 1)}\]. In the \(Y_b\) plane, the input admittance plane, the contours of constant \(G_b\) and \(B_b\) are, of course, straight lines parallel to the axes. To obtain the contours of constant \(G_b\) and \(B_b\) in the \(Z_0\) plane as given in Fig. 4, it is merely necessary to determine how the bilinear transformation expressing \(Z_0\) as a function of \(Y_b\) transforms these orthogonal straight lines.

Since from coupled-circuit theory (see Fig. 1),
the bilinear transformation needed is

\[ Z_0 = -Z_1 + \frac{\omega^2 M^2}{1/\gamma - Z_1} \]

Now the contours of constant \( G_b \) and \( B_b \) in the \( Y_b \) plane \((Y_b = G_b - jB_b)\) all pass through the point at infinity, and since this point is transformed into the point \( Z_0 = -Z_1 - \omega^2 M^2/Z_1 \) by (16), the circles representing contours of constant \( G_b \) and \( B_b \) in the \( Z_0 \) plane must all pass through this point. Similar considerations will verify the fact that these latter contours form the system of orthogonal circles shown in Fig. 4.

As \( C_3 \) is tuned in Fig. 1, the impedance \( Z_0 \), which is the input impedance of network \( A \), must by Theorem I vary along a circle or straight line. Since the maximum-minimum tuning requires that \( C_3 \) be adjusted for maximum value of \( G_b \), it is clear that this required tuning is represented by the point of tangency of the circle or straight line representing the locus of \( Z_0 \) with the family of circles of constant \( G_b \) shown in Fig. 4.

To illustrate the use of Fig. 4, consider the case of a series load, where

\[ Z_0 = R - \frac{j}{\omega C_3} \]  

As \( G_b \) is varied the locus of \( Z_0 \) is the vertical straight line shown in Fig. 5. Since the circles of constant \( G_b \) have their centers on a line having a slope of \( 2 \tan^{-1} 1/Q_i \), it can easily be proved geometrically that the point of tangency \( M \) of the locus of \( Z_0 \) with these circles lies on a line passing through \( P \) and making an angle of \( \tan^{-1} 1/Q_i \) with the horizontal. This point \( M \), being the point on the locus of \( Z_0 \) yielding the maximum value of \( G_b \), is the value to which \( Z_0 \) is adjusted when \( C_1 \) and \( C_3 \) are tuned for maximum-minimum of plate current.

The imaginary part of \( Z_0 \) at the point \( H \) is equal to the imaginary part of \( Z_0 \) at \( P \), and hence

\[ Z_0 = H = R - j \text{Im} (Z_1 + \omega^2 M^2 Y_1) = R - j(\omega L_1 - \omega^2 M^2 B_1). \]

Since

\[ |MH| = \frac{1}{Q_i} |PH| = \frac{1}{Q_i} \left[ R + \text{Re} (Z_1 + \omega^2 M^2 Y_1) \right] \]

we have

\[ Z_0 = R - j \left[ \omega L_1 - \frac{R + R_1}{Q_i} - \omega^2 M^2 G_i \left( \frac{Q_i + 1}{Q_i} \right) \right]. \]

Therefore, by the use of (17) and the fact that

\[ G_i = \frac{1}{R_1(1 + Q_i^2)} \]

we obtain the formula for the tuning of \( C_3 \),

\[ \frac{1}{\omega C_3} = \omega L_1 - \frac{R + R_1}{Q_i} + \frac{\omega^2 M^2}{R_1}. \]

This is the same tuning as given by (9).

The reason for maximum-minimum tuning of \( C_3 \) given by (19) differing from resonance is here seen to be due to two factors. The first is due to coupling, and results in the term \(-\omega^2 M^2 B_1\) being added to the imaginary part of \( Z_0 \) at \( P \) (or at \( H \)). This is just the impedance that would be reflected into the secondary circuit if the pri-
mary circuit were short-circuited. The second factor is due to the resistance in the primary circuit and results in the contours of constant  being rotated through an angle of \(2 \tan^{-1} \frac{1}{Q_1}\) about the point \(P\). This necessitates a correction equal to \(|MI|\), given by (18), in the reactance of \(C_2\).

IV. Tuning for Maximum Coupled-Circuit Efficiency

Perhaps the most important consideration in the determination of the proper tuning of the secondary capacitor \(C_2\) of an inductively coupled output amplifier such as that diagrammed in Fig. 1 is that of coupled-circuit efficiency. This is defined as the fraction of the power that the tube delivers to its load impedance that reaches the secondary circuit. One should, therefore, investigate how well the maximum-minimum tuning agrees with the condition of maximum coupled-circuit efficiency.

It is convenient to take account of the secondary circuit, in the following investigation, by the impedance \(Z_e = \omega^2 M^2/(Z_2 + Z_0)\) that is reflected into the primary circuit. Thus the equivalent circuit for Fig. 1 is that given in Fig. 6. The power transferred to the secondary circuit is accounted for in this equivalent circuit by the power dissipated in \(R_e\), the resistive part of reflected impedance. The coupled-circuit efficiency \(\eta\) is thus seen to be

\[
\eta = \frac{R_e}{R_e + R_1}.
\]

Since \(R_1\) is fixed, the maximum value of \(\eta\) occurs when \(R_e\) is a maximum or the equivalent, \(R_b\) a maximum. Thus the condition for maximum coupled-circuit efficiency is that \(C_2\) be tuned for maximum value of the real part of \(Z_b\), the input impedance of the transformer and hence of network \(B\). This is completely parallel to the condition for maximum-minimum tuning, except that \(Z_b\) now represents \(Z'\). Accordingly, we need only apply formulas (4) and (5) to the bilinear transformation connection

\[
z = -j/\omega C_2\text{ to the impedance } Z_b, \text{ namely the transformation}
\]

\[
Z_b = \frac{A z + B}{C z + D} = \frac{a z + b}{z + d}.
\]

We shall consider the same three special cases treated in Part III of this paper.

A. Series Load

From (6),

\[
Z_b = \frac{Z_1 z + Z_1(Z_2 + R) + \omega^2 M^2}{z + Z_2 + R}.
\]

By reference to (20),

\[
d = Z_2 + R
\]

\[
N = \sqrt{d^2 - 1} = j\omega M.
\]

Therefore by (4)

\[
\frac{1}{\omega C_2} = \frac{\text{Re}(N d^*)}{\text{Im}(N)} = \omega L_2.
\]

Thus (for maximum coupled-circuit efficiency) the secondary is tuned to resonance.

B. Shunt Load

From (13),

\[
Z_b = \frac{|Z_1(Z_2 + R) + \omega^2 M^2| z + (Z_1 Z_2 + \omega^2 M^2) R}{(Z_2 + R) z + Z_2 R}
\]

Thereore by (4)

\[
\frac{1}{\omega C_2} = \omega L_2 \left(1 - \frac{R_2}{R + R_2}\right).
\]

Thus when \(R \ll R_2\) the tuning of \(C_2\) for maximum coupled-circuit efficiency is almost independent of \(R\), which is not the case for resonance of the secondary. It more nearly approximates the maximum-minimum tuning given by (14); however, the latter depends on the coupling and the \(Q\) of the primary and secondary coils as well as on the ratio of \(R_1\) to \(R\).

C. Series Load Taking into Account the Distributed Capacitance of the Secondary Coil

For this case

\[
Z_b = \frac{Z_1(Z_2 + Z_d) + \omega^2 M^2 z + (Z_1 Z_2 + \omega^2 M^2)(Z_d + R) + Z_1 Z_d R}{(Z_2 + Z_d) z + Z_2 Z_d + R}.
\]

Dividing numerator and denominator by \(Z_2 + Z_d\) and applying (4),

\[
\frac{1}{\omega C_2} = \omega L_2 \frac{1 + \frac{1}{Q_2} C_d R}{1 - \frac{1}{\omega^2 L_2 C_d}}.
\]
This is the tuning of $C_1$ for maximum coupled-circuit efficiency; clearly if $C_4$ is zero, this equation reduces to (22).

Note that in all three cases the tuning for maximum coupled circuit efficiency is independent of the coupling. If in (10), (14), and (15) for maximum-minimum of average plate current one sets $k = 0$ and $Q_1 = \infty$, these equations reduce respectively to (22), (24), and (26), the equations giving the tuning of $C_2$ for maximum coupled circuit efficiency.

**Practical Considerations**

The design of the output network of a transmitter depends on several considerations besides that of maximum coupled circuit efficiency; however, once the network is built, the primary factor in tuning the secondary capacitor is to achieve maximum efficiency of power transfer from the plate circuit to the antenna. The plate circuit efficiency of the tube is not to be considered, since it depends only on the load presented to the tube, which presumably can be adjusted to its optimum value by changing the coupling and adjusting the plate capacitor for unity power factor.

One sometimes uses the dc milliammeter of the plate circuit as an indicator for the proper tuning of the output circuit. This is done by alternately tuning the plate and antenna capacitors. Proper tuning is then usually taken to be that for maximum-minimum average plate current, the plate capacitor being tuned for minimum and the antenna capacitor for maximum current.

In this paper it is shown that the above maximum-minimum tuning yields maximum coupled circuit efficiency only if the $Q$ of the plate tank circuit is large and the coupling is very small. Moreover, only for a series load, neglecting distributed capacitance, does the adjustment of the antenna capacitor for resonance of the secondary circuit occur near the adjustment for maximum coupled circuit efficiency—or, for that matter, near the adjustment for maximum-minimum of average plate current.

**APPENDIX**

**Derivation of Equation (4)**

Equation (3) can be written

\[ Y_b = a + \frac{b - ad}{d + d^*}(1 + e^{i\phi}), \]  

(28)

where

\[ e^{i\phi} = \frac{z^* + d^*}{z + d}. \]

Variation of $\phi$ thus causes only a change in $\theta$ in (28); therefore the locus of $Y_b$ is a circle with center at

\[ a + \frac{b - ad}{d + d^*} \]

(29)

and radius

\[ R = \frac{|b - ad|}{d + d^*}. \]

The value of $Y_b$ having the greatest real part $G_b$, i.e., that corresponding to the point $P$ in Fig. 3, is as, seen from (28),

\[ Y_b = a + \frac{b - ad}{d + d^*} + \frac{|b - ad|}{d + d^*}. \]

(31)

Letting $\bar{z}$ be the corresponding value of $z$, i.e., that represented by the point $Q$ in Fig. 3, we have from (27) and (31) and the fact that $\bar{z}^* + \bar{z} = 0$,

\[ \bar{z} = \frac{d^*(b - ad) - d|b - ad|}{b - ad + |b - ad|}. \]

(32)

If $d$ is written as $d = d_1 + jd_2$, then

\[ \bar{z} = d_1 \frac{b - ad - |b - ad|}{b - ad + |b - ad|} - jd_2. \]

It is convenient to introduce a constant $N = N_1 + N_2j$, defined as one of the square roots of $ad - b$; that is

\[ - N^2 = - N_1^2 + N_2^2 - 2N_1N_2j = b - ad. \]

Our equation then simplifies to

\[ \bar{z} = - \frac{1}{\omega C_2} \left( - \frac{N_1}{N_2} d_1 j - d_2 j = - \frac{(N_1d_1 + N_2d_2)j}{N_2} \right). \]

This can be written

\[ \bar{z} = - \frac{1}{\omega C_2} \frac{\text{Real part of } Nd^*}{\text{Imaginary part of } N} = - \frac{\text{Re}(Nd^*)}{\text{Im}(N)}, \]

or, in terms of $C_2$,

\[ \frac{1}{\omega C_2} = \frac{\text{Re}(Nd^*)}{\text{Im}(N)}. \]

(33)

This is a surprisingly simple equation for maximum-minimum tuning of $C_2$, for any network in which a single capacitor $C_2$ is used to tune the secondary circuit.
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Mr. Dickinson joined the vacuum tube department on tube design and application, remaining until 1936 when he joined the General Engineering and Consulting Laboratory to work in the field of high-frequency measurements. From 1941 to 1946 he served as an officer in the Signal Corps, and upon return to the General Electric Company he was assigned to the high voltage and nucleonics division of the General Engineering and Consulting Laboratory to work on particle accelerators. At present he is project engineer for cyclotrons.

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Jean H. Felker (A'46) was born in 1919 at Centralia, Ill. In 1941 he received the B.S. degree in electrical engineering from Washington University, in St. Louis, Mo. After graduation he was associated with the Emerson Electric Co., and later he instructed at a Naval Training School in St. Louis, Mo. In 1942 he joined the U. S. Army, and served as a radar maintenance officer attached to the Royal Artillery in Great Britain, as a member of the Electronics Training Group. Later he returned to this country as an Army publications officer.

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Early in 1946, Dr. Kunz received an appointment as Research Fellow in communications at the Electronics Research Laboratory of Harvard University. One of his chief researches during this period was a mathematical study of the radiation from the open end of waveguides. In 1947 he was appointed lecturer in applied mathematics at the Computation Laboratory of Harvard University. His field of interest has been the development of finite difference methods for solving problems in applied mathematics. He is a member of Phi Beta Kappa, Sigma Xi, and the American Physical Society.

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Upon his return to the United States in 1945, Mr. Levine was assigned to the radar laboratories at Wright Field, Ohio, and continued his project work at that laboratory following his separation from the armed forces. His research interests have included radar systems design, propagation, and television transmitter problems.

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From 1936 to 1937, Mr. Parzen was a radio inspector for the Federal Communications Commission. He served as a civilian electrical engineer for the United States Navy from 1938 to 1944. Mr. Parzen joined Federal Telecommunication Laboratories in 1944, where he is now employed as a senior project engineer in the radio and radar components division.

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J. D. Ryder was born in 1907 in Columbus, Ohio. He received the B.E.E. degree in 1928 and the M.S. degree in 1929 from Ohio State University, where he was Robinson Fellow. In 1944 he was awarded the Ph.D. degree in electrical engineering by Iowa State College. From 1929 to 1931 he was associated with General Electric Company. He joined the Bailey Meter Company of Cleveland, Ohio, in 1931, as supervisor of the electronic and electrical section of the Research Laboratory. He holds twenty-four patents on this work, covering temperature-recording and automatic control applications of electronics. He was appointed assistant professor of electrical engineering at Iowa State College in 1941, was made professor in 1944, and in 1947 assumed his present position of assistant director of the Iowa Engineering Experiment Station. He is the author of two textbooks on electronics and networks, and of other technical papers. Dr. Ryder is a member of the American Institute of Electrical Engineers, American Association for the Advancement of Science, Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

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Abstracts and References

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ACOUSTICS AND AUDIO FREQUENCIES

534.21 On the Calculation of the Sound Field of Circular Membranes in a Rigid Wall—H. Stenzel. (Ann. Phys. (Lpz.), vol. 4, pp. 303-324; March 1, 1949.) Three developments in series are obtained, each with particular advantages for a certain portion of the field. A method is used involving characteristic functions which can be calculated analytically for various differential operators and which can be applied to radiation fields in general, including both acoustic and electromagnetic fields. Practical examples are given.

534.232:621.316.772:621.396.677 2410 Applications of Phase Devices—Tide. (See 2454.)

534.26 The Diffraction of Two-Dimensional Sound Pulses incident on an Infinite Uniform Slit in a Perfectly Reflecting Screen—R. E. Fox. (Phil. Trans., vol. 242, pp. 1-32; May 25, 1949.)


534.862.6:621.396 2420 Closed- and Open-Ridge Waveguide—T. G. Mihran. (Proc. I.R.E., vol. 37, pp. 640-644; June, 1949.) Expressions are derived for the voltage-current and voltage-power impedance, allowing for the discontinuity capacitance. Simplifying approximations and the conditions of their validity are discussed. A preliminary result can be obtained by using Cohn's curves (23 of 1948).

534.926+ 2420 Channel Section Waveguide Radiator—A. L. Cullen. (Phil. Mag., vol. 40, pp. 417-428; April, 1949.) The transmission characteristics are studied theoretically and the phase constant and attenuation coefficient are calculated. Simple approximate formulas are deduced; the validity of these has been verified by experiment. The use of the device as a radiator is limited because the attenuation rate is too great for narrow beams and the normal to the array is rather large.

535.025.2:621.392.52 2421 Crossover Filter for Disc Recording Heads—Rays. (See 2470.)

535.025.3 2422 Magnetic Recording Tapes—M. Camras. (Trans. A.I.F.E., vol. 67, part 1, pp. 503-510; 1948.) Magnetic properties, frequency response, signal-to-noise ratio, output level, sensitivity, and ease of erasure are considered for various types. A new coated material, Type 140-3, developed by the Armour Research Foundation, is found specially suitable for high-quality recording.

535.025.3:621.395.013 2423 An Evaluation of the Application of New and Old Techniques to the Improvement of Magnetic Recording Systems—L. C. Holmes. (Proc. N.E.C. (Chicago), vol. 4, pp. 46; 1948.) Summary only. Discussion of various factors contributing to high quality in such systems, and of data showing the performance characteristics obtained under the most favorable operating conditions is included.

535.026:621.396.44 2424 Channels of Equal Acoustic Quality with Different Frequency-Response Characteristics and Different Bandwidths—T. V. Ananiev (Radiotechnika (Moscow), vol. 4, no. 1, pp. 16-26; January and February, 1949. In Russian.) The necessary channel widths and the permissible deviations in the frequency characteristic for a wire broadcasting system are discussed. Experimental curves and results are given.

536.390.8:621.396.44 2424 Reflection from Corners in Rectangular Wave Guides—A. L. Cullen. (Phil. Mag., vol. 40, pp. 417-428; April, 1949.) The transmission characteristics are studied theoretically and the phase constant and attenuation coefficient are calculated. Simple approximate formulas are deduced; the validity of these has been verified by experiment. The use of the device as a radiator is limited because the attenuation rate is too great for narrow beams and the normal to the array is rather large.

536.392.20+ 2426 Reflection from Corners in Rectangular Wave Guides—Conformal Transformation—S. O. Rice. (Bell Sys. Tech. J., vol. 28, pp. 136-156; January, 1949.) When a bent waveguide is conformally transformed into a straight guide filled with different medium, propagation may either be studied by an integral equation method (2426 below) or by a more general method based upon a certain set of ordinary differential equations. The latter method is developed and applied to determine the reflection produced at the junction of a straight guide and a sectoral horn. This problem cannot be solved by the integral equation method. The WKB approximation, discussed for a single second-order differential equation by Schelleng (1570 of 1946), is extended to a set of equations and approximate expressions for the reflection coefficient are derived.

621.392.20† 2426 Closed- and Open-Ridge Waveguide—T. G. Mihran. (Proc. I.R.E., vol. 37, pp. 640-644; June, 1949.) Expressions are derived for the voltage-current and voltage-power impedance, allowing for the discontinuity capacitance. Simplifying approximations and the conditions of their validity are discussed. A preliminary result can be obtained by using Cohn's curves (23 of 1948).

621.392.30+ 2426 The Effect of Openings in the Walls of Metal Waveguides on the Wave Propagation—
Absorbing plates, the theory of an attenuating effect of openings of various sizes. By analogy with the theory of reflection from optically flat surfaces, the theory of an attenuating line is extended by the introduction of capacitance and inductance variations. Small openings in waveguide walls have no effect either on the propagation or on the attenuation, but with wide slits, the wavelength inside the guide is increased and energy is radiated from the slits. The results of measurements with slits of different widths in different positions in both closed- and open-ended rectangular waveguides are shown graphically and discussed. The effects of parasitic waves are also considered.

62.392.601:62.392.09
Propagation of TE_{0n} Waves in Curved Wave Guides—W. J. Albersheimer. (Bell Syst. Tech. Jour., vol. 28, pp. 1-32; January, 1949.) "TE_{0n} waves transmitted through curved wave guides lose power by conversion to other modes, especially to TM_{0n} modes."

62.392.601:62.392.30
A Waveguide with Phase Velocity u_{c} for the First TE_{01} Wave—K. Kettel. (Frequenz, vol. 3, pp. 73-75; March, 1949.) Propagation theory for a cylindrical waveguide with equally spaced annular partitions. A formula for the phase velocity is derived.

62.392.601:62.392.67
Some Properties of Radiation from Rectangular Wave Guides—J. T. Bolljahn. (Proc. I.R.E., vol. 37, pp. 617-621; June, 1949.) Exact relationships between the radiation pattern and impedance characteristics are valid only so far as the radiation patterns are comparable to those using wide walls of infinitesimal thickness. The ratio of the radiation intensities in certain preferred directions to the power gain in those directions is proportional to the coefficient of reflection inside the waveguide.

62.392.601:62.392.071
Some Properties of Radiation from Rectangular Wave Guides—J. T. Bolljahn. (Proc. I.R.E., vol. 37, pp. 617-621; June, 1949.) Exact relationships between the radiation pattern and impedance characteristics are valid only so far as the radiation patterns are comparable to those using wide walls of infinitesimal thickness. The ratio of the radiation intensities in certain preferred directions to the power gain in those directions is proportional to the coefficient of reflection inside the waveguide.

62.392.3

62.392.67
A High-Gain Cloverleaf Antenna—II. S. Seltzer and E. C. Bishop. (Chicago, vol. 4, pp. 497-504; 1948.) Design is discussed. Laboratory tests on a 1/10-scale model show that power gains of the order 10 to 12 can be obtained. The antenna is suitable for omnidirectional radiation of horizontally polarized waves, as required for FM broadcasting.

62.393
Microwave Antennas and Dielectric Surfaces—R. M. Keilhoffer. (Jour. Appl. Phys., vol. 20, pp. 397-411; April, 1949.) Discussion of the effect of placing a dielectric sheet in front of an antenna for the following cases: (a) reflecting and plane sheet, (b) transmitting antenna and plane sheet, (c) transmitting antenna and cylindrical shell. In case (a), results are explored in terms of an equivalent antenna reflection; phase, arbitrary incidence, and elliptical polarization are considered. In case (b) there is inverse-instantaneous attenuation for a distance of initial antenna mismatch is investigated. In case (c), reflection varies with angle according to the secondary power pattern. Formulas are given for the radiation from a circular cylinder, a narrow strip, a corrugated surface, or a series of strips, with a paraboloidal antenna. Applications to radomes and to the design of pressurizing sealings are discussed. All results are verified experimentally.

62.393.067
New Principle for Broad Band Antennas—M. W. Scheldorf and J. F. Bridges. (Tele-Tek, vol. 8, p. 8; May, 1949.) A description of the characteristics of antennas consisting of several similar elements, which may be of different lengths, arranged in the form of a fan, the inter-element band-width of over 10 percent was obtained with a symmetrical arrangement of 6 elements, the 3 pairs being of different lengths.

62.393.071
Radiation Resistance of Loaded Antennas—R. C. Raymond and W. Webb. (Jour. Appl. Phys., vol. 20, pp. 328-330; April, 1949.) Short cylindrical antennas were loaded by cylindrical or conical dielectric sheath or by external elements. Variations in the frequency below the normal values. Measured resonance resistances are compared with values calculated by the Poynting vector method for assumed sinusoidal current distribution. For a given current distribution at resonance, the antenna radiation resistance depends only on the antenna length. Particular loading methods alter the current distribution. Those which give more nearly uniform current distributions yield higher radiation resistances. See also 2439 below.

62.393.071
Current Distributions on Some Simple Antennas—J. A. Monteith. (Jour. Appl. Phys., vol. 20, pp. 330-333; April, 1949.) A method of measurement is described and measured distributions are compared with some frequently assumed curves. Some distributions were integrated numerically to determine the driving-point impedances. Deviations from curves usually assumed are significant, but do not bring about large errors in the calculated impedances. See also 2438 above.

62.393.071
Measured Impedance of Vertical Antennas over Finite Ground Planes—A. S. Meier and W. F. Rosen. (Jour. Appl. Phys., vol. 20, pp. 699-716; June, 1949.) For a ground plane of dimensions small compared with L, the impedance is a damped oscillating function of L. Thus, the ground-plane dimensions affect the impedance of the antennas. Impedance variations of -5 to -20 per cent were found with a circular ground plane; the corresponding variations for a square ground plane were about half as great, except where the ground plane was small. Impedance is relatively independent of antenna thickness.

At microwave frequencies, measurements were made of the ground-plane impedance following the (Chu) method. The merits of which are compared with those of the conventional shunt-line standing-wave method.

62.393.071

62.393.077
Applications of Phase Devices—Thiede. (See 2454.)

CIRCUITS AND CIRCUIT ELEMENTS

536.1:62.392.601:62.392.011.4
Narrow Gaps—Advantages include low input capacitance, low sensitivity to hum and to noise, and high dynamic range of input signals. Application to a balanced amplifier for a cro, a de, to a meter, amplifiers and tone control are considered.

62.394.25
A Cross-Coupled Input and Phase Inverter Circuit—J. T. Van Scyoc. (Radio and Telev. News, Radio-Electronic Eng. Supplement, vol. 11, pp. 6-9; November, 1948.) The basic circuit is discussed. Advantages include low input capacitance, low sensitivity to hum and noise, and high dynamic range of input signals. Application to a balanced amplifier for a cro, a de, to a meter, amplifiers and tone control are considered.

62.394.37
An Analysis of Interlinked Electric and Magnetic Networks with Application to Magnetic Amplifiers—I. W. ver Planck, M. Fishman, and D. C. Beaumariage. (Proc. N.E.C. (Chicago), vol. 4, p. 426; 1948.) Summary only. A general system of nonlinear equations is developed and applied to determine the stead-state behavior of 6 types of magnetic amplifier without feedback. The relationship between current and time is determined for given applied voltages and circuit elements. Results are confirmed experimentally. Amplifiers using two separate magnetic cores have important advantages over those using a single three-legged core. See also 2448 below.

62.394.37
An Analysis of Magnetic Amplifiers with Feedback—I. W. ver Planck, M. Fishman, and D. C. Beaumariage. (Proc. N.E.C. (Chicago), vol. 4, p. 426; 1948.) Summary only. Two methods are discussed: (a) external feedback, for which a bridge rectifier and separate coils for the feedback current are used, and (b) self feedback, for which two rectifier elements are arranged in series, so that separate feedback windings are not required. The general system of equations discussed in 2447 above is applied; wave shapes and the magnitudes of the currents
resulting from given applied voltages are obtained, and results are confirmed experimentally. The two methods of obtaining feedback have certain very similar characteristics.  

621.34.3  


621.34.3  

621.34 The Transducer—H. B. Rex. (Instruments, vol. 20, pp. 1102-1109; December, 1947.) Discussion of (a) the transducer without construction, (b) the self-excited transducer under various conditions of feedback and (c) the transducer with constrained magnetization. The article is based on papers by T. B. Buchanan presented at the April, 1942, to 1944, noted in 3547 of 1942, 363 of 1943, and 55 of 1944. See also 2451 below.

621.34.3:106  

621.34 Bibliography on Transducers, Magnetic Amplifiers, etc.—H. B. Rex. (Instruments, vol. 21, pp. 332-362; April, 1944.) A list of 213 references, with brief notes indicating the scope of many of the papers.

621.34.63  

621.34 A Practical Approach to Calculating Optimum Performance of Semiconductor Rectifiers—E. D. Wilson. (Trans. A.I.E.E., vol. 62, part 11, pp. 642-.) By applying Kirchhoff's laws to a conventional 4-ram rectifier bridge, an expression is derived for efficiency in terms of the reverse R, forward F, and load L resistances. Maximum efficiency is obtained when L/R = F - R; the forward and reverse losses are then equal. Efficiency versus temperature curves are given for various current densities.

621.3.6  

621.362.2  

621.361.722.4  

621.36 Controllable Voltage Divider—S. Freedman. (Radio and Television News, Radio-Electronic Eng. Supplement, vol. 11, pp. 7-9; 29; October, 1948.) The basic circuit is discussed for a device whose output depends only upon the ratio of two separate voltages applied to the input, and not upon the actual values of these voltages. The principle is applied in an AM-FM detector in which the output AM signal is eliminated during FM reception, and vice versa.

621.361.727:621.396.677:514.232  

621.36 Applications of Phase Devices—H. Fliege. (Funk und Ton, vol. 3, pp. 249-255; May, 1949.) Arrangements are described for obtaining directional acoustics, for the creation from a number of radiators spaced uniformly in a straight line. The phases of the voltages applied to consecutive transmitters differ by an amount δ which determines the direction of the beam relative to the line of radiators. Circuits including n crossed-coil transformers, each connected to two of the 2n radiators, permit the determination of δ for the phase difference δ so that the beam can be swept through a prescribed angle without varying the position of the radiators. Typical directional characteristics obtained with 8 radiators in line are illustrated. Similar devices can be used for reception. See also 1873 and 2738 of 1948.

621.3.71  

621.37.71 An Automatic Current Integrator—M. J. Poole. (Jour. Sci. Instr., vol. 26, pp. 113-114; April, 1949.) Circuit diagram and description of the detector for measuring mean current ranging from 2X10^-4 to A at 10^-4 A. Arrangement are included to enable counting apparatus recording nuclear disintegrations to be switched over on exact charge intervals.

621.3.78:512.99  


621.38:572:539.10.08  


621.39:109.9  


621.39.5:325.707  


621.39:832:535.877  

621.59 A Practical Approach to Calculating Optimum Performance of Semiconductor Rectifiers—E. D. Wilson. (Trans. A.I.E.E., vol. 62, part 11, pp. 642-.) By applying Kirchhoff's laws to a conventional 4-ram rectifier bridge, an expression is derived for efficiency in terms of the reverse R, forward F, and load L resistances. Maximum efficiency is obtained when L/R = F - R; the forward and reverse losses are then equal. Efficiency versus temperature curves are given for various current densities.

621.39:52:537:228.1  

621.59 Crystal Filters Using Ethylene Diamine Tetracetic Acid. —J. B. Willis. (Trans. A.I.E.E., vol. 67, part 1, pp. 552-556; 1948.) Performance is comparable to that of quartz filters. See also 2215 of 1948, for which the above Universal Decimal Classification is preferable, and 329 below.

621.39:202:621.396.261  

621.69 A Practical Approach to Calculating Optimum Performance of Semiconductor Rectifiers—E. D. Wilson. (Trans. A.I.E.E., vol. 62, part 11, pp. 642-.) By applying Kirchhoff's laws to a conventional 4-ram rectifier bridge, an expression is derived for efficiency in terms of the reverse R, forward F, and load L resistances. Maximum efficiency is obtained when L/R = F - R; the forward and reverse losses are then equal. Efficiency versus temperature curves are given for various current densities.

621.39:52:621.395.625.2  

621.69 Crossover Filter for Disc Recording Heads—H. E. Roys. (Audio Eng., vol. 33, pp. 18-21; June, 1949.) Description of a practical device designed for crossovers in cutters in cutter characteristics, both at the transition frequency and at the high-frequency end of the spectrum.

621.39:52:621.396.665  


621.39:661.1  

621.396.611.1:621.317.6:621.396.645 2473
The Response of a Tuned Amplituner to a Signal Varying Linearly in Frequency—W. H. Hamilton. (Proc. N.E.C. (Chicago), vol. 4, pp. 373-377; 1948.) An analysis of the mathematical properties of tuning the solution obtained from a differential analyzer for the integral

\[ U(2m, n) = \int_0^\infty \cos \gamma dy \]

and a similar integral involving \( \sin \gamma \). A corresponding integral is obtained for various values of the parameter \( \gamma = \lambda \times \text{bandwidth} \), being the interval during which the frequency of the input signal is within the half-power points of the desired tuned circuit. A curve of signal-to-noise ratio as a function of \( \gamma \) is included. Results for a 450-kc amplifier are shown graphically. See also 670 of N. (Barber and Urrat) and 671 of March (Hok).

621.396.611.21 2474
The Theory of a Self-Oscillator with a Quartz Crystal—S. I. Evtynov. (Radio tekhnika (Moscow), vol. 4, pp. 27-40: January and February, 1949. In Russian.) The quartz crystal is treated as a positive feedback element. A vector method is used in the analysis. The condition for self-excitation is established and formulas are derived for determining the amplitude of oscillations and the frequency change corresponding to a change in the tuning of the anode circuit. The stability of steady-state operation is examined, when the tuning of the anode circuit is allowed to vary slowly. A voltage-hysteresis effect takes place in the oscillator.

621.396.613:621.396.619.13 2475
The Application of Coupled Systems with Distributed Constants to Frequency Modulation in the Upper Microwave Region—V. N. Vlahov. (Radio tekhnika (Moscow), vol. 4, pp. 69-74: March and April, 1949. In Russian.) Modern FM methods for meter waves require manifold frequency modulation which is done even for sound broad broadcasting and it is doubtful whether these methods can be used at higher frequencies. Experiments were, therefore, conducted to study the effect of a secondary system with distributed constants on the frequency and amplitude of oscillations in a self-exciting system with distributed constants (Fig. 1). These experiments were made with two experimental circuits. It is impossible, in principle, to obtain direct modulation of the carrier of sufficient depth without using frequency multiplication.

621.396.611.4:621.392.26 2476
Tunable Waveguide Cavity Resonator for Broadband Operation of Reflex Klystrons—W. W. Harman. (Proc. N.E.C. (Chicago), vol. 4, pp. 233-252; 1948.) A study of the design of broad-band reflex resonators (2 to 1 tuning range) for use with reflex klystrons of the external-cavity type. Two groups are considered: (a) \( \lambda/4 \) or fundamental-mode resonators, whose high-frequency set is fixed by the physical size of the tube envelope, and (b) \( \lambda/4 \) resonators, which allow a variation up to the physical limit of the tube.

621.396.611.4.202.64:65 2477
Modified Cavity Oscillator for the Generation of Microwaves—G. G. Bruck. (Jour. Appl. Phys.), vol. 18, pp. 843-844; September, 1947.) A modification of a reflex-cavity oscillator to give frequency and amplitude stability. The inner cavity is of sufficient length so that the high-frequency set is determined by the physical size of the tube. This cavity is high enough to allow a small degree of freedom in the tuning of the cavity; hence, there is a slight variation in the frequency of the cavity. High output should be obtainable at frequencies above 30 Mkc; this type of oscillator is thought capable of working at 100 Mkc or higher.

621.396.615 2478

621.396.615 2479

621.396.615.029.64 2480
The Self-Excitation of a Triode Oscillator Loaded by a Line with Distributed Constants, at Microwave Frequencies—V. A. Zoren. (Zh. Tekh. Fiz., vol. 19, pp. 570-577; May, 1949. In Russian.) The theory of Gvozdev and Zoren (999 of May) is applied to the case of a triode oscillator loaded by a coaxial or 2-wire transmission line (Fig. 1). The operation of the oscillator is discussed and formulas (8) and (9) determining respectively the frequency and the condition of self-excitation are derived. The effect of the transient time of electrons on the condition for self-excitation is examined in detail and methods are indicated for determining the resonance length that can be generated by the oscillator.

621.396.615.17 2481
An Improved Regenerative Frequency Standard Application—F. E. Wyman. (Proc. N.E.C. (Chicago), vol. 4, pp. 406-413; 1948.) This arrangement uses two tuned sections in the mixer. In the frequency ratio (10 to 1), it requires no voltage-regulated power supply, and has satisfactory self-starting and locking qualities.

621.396.619 2482
Contribution to the General Theory of Modulation and Demodulation for Any Type of Characteristic—O. Heymann. (Arch. Elek. (Erbtragung), vol. 3, pp. 73-79; March, 1949.) The given characteristic is represented by a Fourier series. It is shown that the tube output voltage can then be obtained directly as a Fourier series. The coefficients of this series are definite integrals whose integrands are given by a Fourier series and two additional elementary functions. The form of the constant term of the series is particularly simple. This fact is very important in applications of the theory, since it enables the low-frequency behavior to be presented clearly and facilitates numerical evaluation. Two examples illustrate the usefulness of the general formula.

621.396.621 2483
Wideband Frequency-Discriminator Design C. Rideout. (Proc. N.E.C. (Chicago), vol. 4, pp. 414-424; 1948.) An analysis of the ordinary transformer-coupled frequency discriminator for a single channel. For a multi-channel frequency discriminator, the response can be calculated even for cases where the bandwidth is a large percentage of the center frequency. The output voltage drops to zero at a frequency above the transformer mfl-frequency; improvement in linearity can, therefore, be better effected by detuning than by overcoupling of the transformer; hence, no physical limitation of this type is present. The voltage on the accelerating grid is so related to the potential difference between the cylinders that orbits are initially circular. Microwave power is extracted from the inner cavity. High output should be obtainable at frequencies above 30 Mkc; this type of oscillator is thought capable of working at 100 Mkc or higher.

621.396.645 2484
A Concise Theory of Aperiodic Amplifiers—B. A. Khovan. (Radio tekhnika (Moscow), vol. 4, pp. 57-68; March and April, 1949. In Russian.) In studying the operation of a low-frequency amplification stage, it is customary to consider three different equivalent circuits corresponding respectively to the lower, middle and upper frequencies of the bandwidth and to derive three separate equations for the frequency and phase characteristics of the stage. A single formula is here derived containing a minimum number of generalized parameters; and the method of deriving the equations in a manner the characteristics of the main types of amplification circuits. The results obtained are tabulated.

621.396.645 2485
The Chain Amplifier—H. Feige. (Proc. IRE, vol. 3, pp. 291-301; March, 1949.) The principles and mode of operation of distributed-amplifier circuits are considered. The grid-to-cathode capacitances of the amplifier tubes on the one hand, and the anode-to-cathode capacitances on the other hand, form the cross capacitances of balanced transmission lines and are uniformly distributed along their length. The amplification and frequency characteristics of this type of amplifier are calculated from transmission-line theory; they are not subject to the limitations of the usual type of amplifier. An amplifier of frequency response curve can be obtained from 0 to 200 Mc, using ordinary tubes. The upper frequency limit is determined solely by the unavoidable grid-to-cathode capacitances of the tubes. See also 3755 of 1948 (Cinzott et al.).

621.396.645 2486
A New Type of High-Frequency Amplifier—J. R. Pierce and W. B. Heenestien. (Bell Sys. Tech. Jour., vol. 28, pp. 33-51; January, 1949.) Amplification is obtained by means of an electromechanical interaction within the electron flow, which consists of two streams with different average velocities. When the current in the charge detector of one of these streams is sufficient, the streams interact to give an exponentially increasing wave. Conditions for the gain and an increasing wave are determined for a particular geometry of the amplifier. Advantages include: (a) the electron flow need not pass close to complicated circuit elements, (b) for a sufficiently great length of electron flow, amplification can be obtained even though input and output circuits have very low impedance or poor coupling to the electron flow, and (c) close synchronization between electron velocity and input frequency is not essential. See also 2487 below.

621.396.645 2487
The Experimental Observation of Amplification by Interaction between Two Electron Streams—A. V. Hollenberg. (Bell Sys. Tech. Jour., vol. 28, pp. 52-58; January, 1949.) Discussion of the construction and performance of the amplifier used, which has identical input and output helices connected identically to a coaxial line several wavelengths long between the helices. Concentric tubular electron streams originate at the ring-shaped emitting surfaces of the two cathodes, pass through their respective control grids, and then through a common accelerating grid. A magnetic field of about 700 gauss is applied to maintain the definition of the streams. The gain produced depends on a difference in velocity between the interacting streams and is used to control the inner stream by the helix velocity that is that at which traveling-wave amplifier interaction between the stream and helix occurs. The outer stream travels at a lower velocity. A gain of 33 db at a center frequency of 255 Mc was observed, with bandwidth of 110 Mc between 3-db points. The gain produced by the interaction from the 255 Mc input stream travels at a lower velocity. The theory of Pierce and Heenestien (2486 above) is thus qualitatively confirmed, though the actual conditions in the amplifier differ from the theoretical and all assumptions.
New Developments in Preamplifiers—C. D. C.


A comprehensive research. Evidence is given of the existence of wave-propagation phenomena in the motion of discharges in discharge tubes; the propagation velocity is determined from measurements of the standing waves obtained with suitable excitation. A tentative theory of the observed phenomena is discussed. Results are given for the voltage at which a discharge suddenly occurs in He, using frequencies from 1 to 75 Mc and pressures from 10⁻⁶ to 1.5 mm Hg. The curves extend to the results of other workers concerning the existence of abrupt variations of the sudden-discharge voltage. The dimensions and arrangement of the electrodes are of fundamental importance. Equations are established between the various factors determining the amplitude of the abrupt voltage variation; a possible explanation of the effect is suggested. Continuous discharge from a kind of cut-off which is bounded by medium 3.

Frequency repetition frequency 50 to 2,000 per second.

The cut-off extends over a considerable wavelength in Ne at pressures above 0.2 mm Hg and in He above 5.0 mm Hg. Part 1, 1906 of 1948.

Secondary Electron Emission—H. Salow. (Funk und Tom, vol. 3, pp. 278–285; May, 1949.) Discussion of the cathode-follower circuit with (a) current feedback, (b) voltage feedback, and (c) mixed current and voltage feedback, with formulas for the load current, output voltage, and gain.

Circuit Design for Reduction of Humidity—F. Dickson. (Proc. N.E.C. (Chicago), vol. 4, p. 425; 1948.) Summary only. The chief sources of hum are electromagnetic fields, electrical leakage, badly arranged input circuit, and a bad front-end feedback network. Hum in high-frequency local oscillators is also discussed. Practical remedies are suggested, but no single device can eliminate all types of hum.

GENERAL PHYSICS

547.1: 5129

Tensors and Electricity—B. Thourhill. (Ser. 2541.)

547.52: 538.560.209:5:520


A comprehensive research. Evidence is given of the existence of wave-propagation phenomena in the motion of discharges in discharge tubes; the propagation velocity is determined from measurements of the standing waves obtained with suitable excitation. A tentative theory of the observed phenomena is discussed. Results are given for the voltage at which a discharge suddenly occurs in He, using frequencies from 1 to 75 Mc and pressures from 10⁻⁶ to 1.5 mm Hg. The curves extend to the results of other workers concerning the existence of abrupt variations of the sudden-discharge voltage. The dimensions and arrangement of the electrodes are of fundamental importance. Equations are established between the various factors determining the amplitude of the abrupt voltage variation; a possible explanation of the effect is suggested. Continuous discharge from a kind of cut-off which is bounded by medium 3.
### Abstracts and References

#### 525.73:523.85:621.396.822

Radio Astronomy—C. R. Burrows. (Sci. Mon., vol. 68, pp. 299-304; May, 1949.) Elementary discussion of: (a) the frequencies at which radio astronomical observatories are feasible, (b) the equipment and methods used for radio telescopes and for communication receivers, (c) the variation of solar apparent temperature with frequency, and (d) galactic radiation, and the location of its emission sources.

#### 523.72:523.64:621.396.822

Some Observations on Solar Noise—Chief Engineer's Branch, General Post Office, Wellington. (N.Z. Jour. Sci. Tech., vol. 29, Section B, p. 149; 1947.) Discussion of: (a) the appearance of the sunspots and the type of noise observed by the Wellington observatory at Avawaua during a period of high sunspot activity in February, 1946, at frequencies between 6.7 and 25.6 Mc. Simultaneous direction-finding observations prove conclusively the solar origin of the noise. Some of the surges lasted as long as one minute and reached high signal intensities.

#### 527.72:020.63:523.746:621.396.822

Relation between Decimetre-Wave Radar Reflections and Sunspots—J. F. Deniau. (Compt. Rend. Acad. Sci. (Paris), vol. 228, pp. 1571-1572; May 16, 1949.) The correlation coefficient between sunspots and the 10.6-cm solar echoes observed by the service observatory is found to be 0.6. The work of Secrest and Omark (523.13 of 1948) shows a progressive increase according as comparison is made with (a) the relative sunspot number $N$, (b) the sum $SA$ of the areas of individual spots, and (c) the daily number $SA/\pi A$, where $A$ is the magnetic field of a spot of area $A$. The correlation coefficients are respectively (a) 0.53, (b) 0.76, and (c) 0.87.

#### 525.75

Solar Flares and their Terrestrial Effects—M. E. Haines. (Nature (London), vol. 163, pp. 749-753; May 14, 1949.) Substance of lecture to the Cavendish Laboratory Radio Section. Flares occur within a radius of 1.5 million miles of the sun and are not infrequently between the leading and following elements of a bipolar group. The largest flares sometimes obscure the largest sunspots. An intense flare occurs on a sunspot of magnitude greater than magnitude 2.5. A flash of radiation followed by a slow decay. The development curve of such a flare is correlated with the corresponding observations on the earth. Intense radio fluxes by means of radio telescopes, incoming cosmic rays, and other phenomena are caused by ultraviolet radiation. Intense radio waves and short wave signals, and anomalous changes in the phase of long wave signals, are observed in the course of a solar eclipse. Delayed effects, due to particles, include geomagnetic storms and abnormal cosmic radiation. The theoretical work of Giovanelli (223.20 of 1946 and 376 of March) and of Uralov (2217 of September) is briefly mentioned.

#### 525.10:5:621.396.9

Radar as an Aid to the Study of the Atmosphere—Jones. (See 2526.)

#### 525.10:5:523.85:621.396.85:621.396.72

The Internal Reflection of Electromagnetic Waves in a Stratified Medium: Application to the Troposphere—Eckart and Kahan. (See 2605.)

#### 525.10:5:525.2:621.396.82

The Releasability of Ionospheric Height Determinations—J. P. M. MacAlister. (Proc. Phys. Soc. London, vol. 63, p. 205; May, 1949.) The determination of the height of the ionosphere by means of radio interferometer experiments is discussed. The height of the layer is determined from the difference in phase of the signal received at two different distances. Some of these signals, with a frequency of 10 Mc, are actually received at an angle of $10^\circ$.

#### 525.10:5:521.396.5:621.396.82

Short-Period Variations in the Ionosphere—G. H. Munro. (Nature (London), vol. 163, pp. 812-814; May 21, 1949.) Correlation between the phase of the sunspot cycle and the height of the ionosphere, which show both a downward trend and a horizontal progression. Results obtained on June 22, 1948, are shown graphically and discussed.

#### 525.10:5:521.396.5:621.396.82

The Ionosphere over Mid-Germany in March 1949—Dixmier. (Fonndmeftech. Z., vol. 2, p. 157; May, 1949.) Continuation of 1661 of July and 1934 of September. On various specified dates, the limiting frequencies were from 10 to 35 percent below the normal value. See also 2519 below.

#### 525.10:5:521.396.5:621.396.82

The Ionosphere over Mid-Germany in April 1949—Dixmier. (Fonndmeftech. Z., vol. 2, p. 172; June, 1949.) Continuation of 2518 above. In April, the change from winter to summer ionosphere conditions was complete. Disturbances and abnormal values of critical frequency are noted.

#### 525.10:5:521.396.5:621.396.11

The Analysis of Ionospheric Reflections: Part 2—de Voogt. (See 2606.)

#### 525.10:5:523.78:1949.04.28

Observations of the Atmospheric Electric Field during the partial Solar Eclipse of 28th April 1949—J. Rook. (Compt. Rend. Acad. Sci. Paris, vol. 229, pp. 547-549; May 16, 1949.) Field measurements in Monaco with a Wulf biauricular electrometer before, during, and after the eclipse are plotted, together with results from a similar experiment performed at Hyères, France, on the same day. Before the eclipse started, the field strength was abnormal for the season, with a value of 100 to 200 volts per meter. In the three-quarter hour period after the eclipse started, the field increased to 720 volts per meter. It remained high during the whole of the eclipse, with a peak reading $850$ volts per meter, after the eclipse, it oscillated around 550 volts per meter. On the following day, there was a fairly regular increase from 200 volts per meter to 500 volts per meter. The mean value for the month of April is 300 volts per meter and no values as high as 600 volts per meter were reached. A marked increase of the electric field was also observed during the eclipse of April 18, 1921. The random variations of the field are, however, so frequent and sometimes so large that it is possible for the large field increases observed, both in 1949 and 1921, to be chance coincidences.

#### 525.59:4:621.49:525.22

Auroral Displays at Saskatoon—W. Petrie, P. A. Forsyth, and E. McConkey. (Nature (London), vol. 163, p. 774; May 14, 1949.) Brief description of intense displays observed during auroral arcs on January 24 and 25, 1949, and February 21 and 22, 1949. All auroral forms were present at some time during both displays. A large sunspot group passed the solar meridian on January 22, 1949, and was present, though reduced in size, on February 21, 1949.

#### 523.72:525.64:621.396.82

Location and Aids to Navigation

534.88:623.06

An Electroacoustic System—E. A. Walker. (Trans. AIEEE, vol. 67, part 1, pp. 35-40; 1948. Discussion, pp. 40-41.) The location of torpedoes lost in firing tests is discussed. An ultrasonic generator is fitted in the target and the torpedo, the operator being able to place a portable hydrophone capable of giving the approximate bearing and depth of the source. The simple magnetostriiction generator is driven by two 3Q5-GT tubes, used as magnetic oscillator and power amplifier, and produces a signal audible above background noise at 1,000 yards. Power is obtained from storage batteries.

621.396.9

Radar—E. G. Schneider. (Bull. Schweiz. Elektrotech. Ver., vol. 39, pp. 192-196, 251-256, 290-291, 313-316, and 343-344; March 20 to May 15, 1949.) Radar is being used to investigate precipitation, to detect clouds dangerous to flying, and to measure cloud height and wind velocity. Techniques are described for measuring pressure, temperature, humidity, and atmospheric density in the upper air.

621.396.9:523.53

The Diffraction of Radio Waves from Meteor Trails and the Measurement of Meteor Velocities—Davies and Elliott. (See 2508.)

621.396.9:523.51:523.52

Radar as an Aid to the Study of the Atmosphere—F. E. Jones. (Jour. Roy. Aero. Soc., vol. 53, pp. 443-444; April 21, 1949.) A new technique for measuring precipitation, to detect clouds dangerous to flying, and to measure cloud height and wind velocity. Techniques are described for measuring pressure, temperature, humidity, and atmospheric density in the upper air.

621.396.933:94

Operational Trials of the Australian Distance-Measuring Equipment and Multiple-Track Radar Range—J. G. Downes. (Proc. I.R.E. (Australia), vol. 9, pp. 10-21; April, 1948.) Discussion, pp. 21-23.) The desirable features of short-range navigational aids for aircraft are discussed, and the MTR is a new Multiple-Track Radar Range, used in Australia. The presentation consists of track numbers indicated on a meter. The resulting Operational Trials of the Australian Distance-Measuring Equipment, described. Three ground stations were used, but the coverage was not complete over the 450-mile path.

#### MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7

An Investigation on Hot-Wire Vacuum

537.228.1 2520


537.228.1:548.5 2530


537.312.62 2531


546.212-16:621.3.011.5.029.64 2532

The Dielectric Properties of Ice at 1.25 cm Wavelength—J. Lamb and A. Turney. (Proc. Phys. Soc., vol. 62, pp. 272-273; April 1, 1949.) The dielectric constant is measured approximately at 3.18 between 0°C and -20°C. The value of tan δ falls continuously with decreasing temperature. Further dipole relaxation, apart from that giving a loss maximum at about 10⁶ cps is unlikely. A large atomic polarization may account for the high value of the permittivity.

546.431.82 2533


546.431.82 2534


621.315.62:621.396.602.029.64 2535


621.315.62 2536

Ceramic Materials with a High Dielectric Constant—S. Cremer. (Proc. A.I.E.E., vol. 67, part 1, pp. 55-57; March, 1948.) Discussion. The fundamental circuits and general principles of operation of a pressed-card machine are briefly discussed, with detailed diagrams. Simple electronic counters are briefly considered, with the automatic sequence controlled calculator discussed in 461, 468, and 787 of 1947.
Abstracts and References

5261.3.209.64

5261.3.323(08.74)

5261.3.331:621.319.4.015.3

5261.3.335.3.029.64:621.315.611.011.5
Microwave Dielectric Measurements—T. W. Dakin and C. N. Works. (J. Appl. Phys., vol. 18, pp. 789–796; September, 1947.) The method of Roberts and von HippeL (178 of 1947) was used with 10-cm and 3-cm waves. A simplified procedure for calculating the dielectric properties from the measurements is given. Results are tabulated.

5261.3.336
The Measurement of Antenna Impedance Using a Receiving Antenna—D. G. Wilson and R. W. P. King. (Proc. N.E.C. (Chicago), vol. 4, p. 496; 1948.) Summary only. Energy from the transmitteE receiver was detected by an antenna located above a large conducting plate and terminated in a slotted coaxial cable. The combined phase and damping functions for the two ends of the cavity were determined by the resonance-curve method. Since the phase and damping due to the lower end could be measured independently, the phase and damping due to the upper end could be found and used to calculate the antenna impedance. Curves of impedance as a function of length of antenna were obtained for several antennas. The agreement between the results determined in this way, was in reasonable agreement with the value measured when using the same apparatus as a transmitting system.

5261.3.337.336
Antenna Impedance Measurement by Reflection Method—E. Iswandy. (Proc. I.R.E., vol. 37, pp. 604–608; June, 1949.) The dipole whose impedance is to be measured is mounted on a small wooden carriage which can be moved along a track in front of transmitting and receiving parabolic reflectors mounted side by side. The amplitude of the received signal depends on the phase deflected by the dipole, and the target may be moved until maximum power is received in the phase of the dipole impedance. Results obtained for dipole of various lengths and diameters of 0.14, 0.48, 3, 13.6, and 40 mm are discussed.

5261.3.373.029.64
A Method of Measuring Phase at Microwave Frequencies—Stev. D. Robertson. (IEEE Symp. Tech. Jour., vol. 28, pp. 99–103; January, 1949.) One part of the output of a signal generator is modulated by an af signal in a balanced modulator. The resulting suppressed-carrier double-sideband signal is applied to the device to be measured, in which means are provided for sampling the signal at both input and output. The phase measurement was received when the generator output is fed through a calibrated phase shifter and applied to a crystal detector, to which the signal samples are then applied successively, the phase shifter being in each case adjusted for minimum af signal in the detector output. The difference between the two settings of the phase shifter gives the differential phase shift between the two samples. Compensation for amplitude inequalities of the two samples is thus unnecessary.

5261.7.087/088
Frequency Response Characteristics of Recording Instruments—T. D. Graybeal. (Trans. A.I.E.E., vol. 67, part 1, pp. 755–766; 1948.) Methodology for determining the errors due to the dynamic limitations of instrument movements, and which appear in records. The ac steady-state response may be obtained by using charts, given the undamped natural frequency and the relative damping. Test procedure is described for finding these two basic constants of a particular instrument. The importance of proper damping is emphasized; techniques are suggested whereby the frequency response may be made even better than that obtained by these methods. Analysis of servo-actuated recorders is included. Numerical examples are given.

5261.7.71
An Automatic Current Integrator—Poole. (See 2455.)

5261.7.72:621.306.645
D.C. Amplifier Stabilized for Zero and Gain—P. N. Etherington and W. R. Clark. (Trans. A.I.E.E., vol. 67, part 1, pp. 47–57; 1948.) An amplifier for dc measurements is described in detail. The zero is stabilized by means of a dc-evacuated vibrating at 60 cps. The over-all voltage gain is about 40X10^6. Great care is taken to ensure that no unwanted 60-cps component can affect the zero. Gain is locked in to a small fraction of a per cent of the use of overall dc feedback. Circuits are described for various measurements involving voltages of the order of 0.05 μv.

5261.7.726:621.319.4
A Wide-Range Variable Capacitor—C. W. Bowder. (J. Appl. Phys., vol. 26, pp. 117–119; April, 1949.) Description of a 3-terminal capacitor covering the range 0.01 to 10 pf. With a pair of Si rectifiers and a dc measuring instrument it can be used as a rf peaks volt meter for the range 1 to 10 kv at frequencies from 50 kc to 50 Mc.

5261.7.733:621.799
A Direct Reading D.C. Bridge for Microwave Power Measurements—H. J. Carlin and J. Blam. (Trans. A.I.E.E., vol. 67, part 1, pp. 311–315; 1945.) A self-balanced, soft iron, bolometer bridge circuit suitable for field work is described in detail. The bridge uses two bolometer elements, a Wheatstone wire for the range 0 to 1 milliwatt and a metalized glass element for the range 10 to 100 milliwatts. The design of a built-in attenuator with a range of 0.03 to 3.0 and maximum error 2.0 per cent is discussed.

5261.7.733.029.64
A Microwave Impedance Bridge—M. Chorhorow, E. L. Ginorn, and F. Kane. (Proc. I.R.E., vol. 37, pp. 634–639; June, 1949.) A simple, direct-reading, waveguide vacuum tube bridge circuit is discussed. The circuit uses only two vacuum tubes, one served as a modulator, the other as a detector. The phase and amplitude of the output from the bridge are measured to about the same accuracy as with a standing-wave detector. The standard impedances required are three variable reactances (movable shorting plungers) and A and Z terminations. The positions of the plungers are measured. The device can be used as a 4-terminal lattice section for filter design or other applications: greater flexibility is obtained than with conventional test procedures.

5261.7.740.029.44
Microscope Slotted Sections—S. A. Johnson. (Proc. N.E.C. (Chicago), vol. 4, pp. 222–232; 1948.) Instruments used to measure the magnitude and phase of standing waves in a transmission line or waveguide without introducing appreciable field disturbance. The electrical impedance of the line can be deduced. High mechanical precision is required in the construction of the stereomicroscope. Sections for use at frequencies up to 75 kMc are described. Tuning for matching the probe of the detector is considered.

5261.7.755

5261.7.755:621.701
A Polar Vector Indicator—A. H. Waynick, P. G. Sulzer, and E. A. Walker. (Proc. N.E.C. (Chicago), vol. 4, pp. 446–451; 1948.) An electronic device for demonstration purposes, which displays and permits determination of the magnitudes and phase angles of as many as three voltages in the field. Frequency range is 15 to 300 cps, amplitude accuracy within ±1 per cent, phase accuracy within ±3°.

5261.7.761
Contribution to the Technique of 'Density' Standard-Frequency Spectra—H. J. Grisela. (Ber. d. Phys. Ges., vol. 10, pp. 161–176; 1949.) The term 'density' is applied to spectra whose components are so close together in frequency that they can be separated from one another by means of simple electrical circuits. Special technique is required in applications involving such spectra. Methods of generating dense spectra by the use of dc and ac pulses, and arrangements for frequency multiplication and for pulse modulation are described. Special equipment discussed includes a (a) generator providing standard frequencies at 10-kc intervals, (b) a 100-kc interval, (c) a r-f power amplifier, (d) an intercompensation frequency meter, (e) a direct-reading frequency meter, (f) apparatus for frequency multiplication, (g) a receiving frequency, (h) frequency range is 15 to 300 cps, amplitude accuracy within ±1 per cent, phase accuracy within ±3°.

5261.7.772:621.306.67
Measurement of the Phase of Radiation from Antennas—J. N. Hines and C. H. Bohneker. (Proc. N.E.C. (Chicago), vol. 4, pp. 487–495; 1948.) The equipment described consists of a signal generator, attenuator, probe antenna, T-junction mixer, and receiver. Two signals are obtained in the T-junction which combine if in phase and cancel if out of phase. Phase contours thus measured for various antennas are shown. Prediction of amplitude and phase patterns of arrays by combining vectorially the patterns of individual antennas is discussed.

5261.7.784.087.44
A Square-Law Power-Level Recorder—W. R. Clark, W. R. Turner, and A. J. Williams, Jr. (Proc. N.E.C. (Chicago), vol. 4, pp. 446–451; 1948.) A rectangular scale ab and a datum level of 0.0002 milliwatts. The error is less than 0.1 db between 20 cps and 200 kc. The operation of the recorder is described. See also 488 of 1941 (Clark).
A Simple Method for Determining Distortion Factors—F. Enkel. (Fermi-Tech Z., vol. 2, pp. 153-154; May, 1949.) A signal of frequency 800 cps is applied to the system to be tested, and the amplitude of the second and third harmonics of the output is measured, using in succession band filters tuned to 1,000 and 2,400 cps. The nonlinear distortion can then be calculated from simple formulas which are valid for the condition \\
naming the variation of the distortion factor determined for (a) a triode amplifier, (b) a pentode amplifier, and (c) a magnetophone, as a function of the loading.

A Standard Method for Determining Distortion Factors—F. Enkel. (Fermi-Tech Z., vol. 2, pp. 153-154; May, 1949.) A signal of frequency 800 cps is applied to the system to be tested, and the amplitude of the second and third harmonics of the output is measured, using in succession band filters tuned to 1,000 and 2,400 cps. The nonlinear distortion can then be calculated from simple formulas which are valid for the condition \\
naming the variation of the distortion factor determined for (a) a triode amplifier, (b) a pentode amplifier, and (c) a magnetophone, as a function of the loading.
PROPAGATION OF WAVES

538.56 1949

The Approximate Solution of One-Dimensional Wave Equations—J. Botti (Rev. Mod. Phys., vol. 20, pp. 399-417; April, 1948.)

A general discussion, in terms of Hamiltonian functions, of the type of differential equation applicable to waves on a stretched string, sound waves in a tube of uniform cross section, the propagation of electric currents along a uniform wire, etc. The Fourier solution of the initial-value problem, group and phase velocity, the nature of the waves in the regions inside and outside the reflector, wave fronts, and unresolved waves are considered.

538.56 1950

The Sommerfeld's "Radiation Condition"—F. V. Atkinson (Phil. Mag., vol. 40, pp. 645-661; June, 1949.)

Discussion of the uniqueness of solution in problems of wave motion involving the Sommerfeld's condition as formulated under which the solution is proved to be unique; these include Sommerfeld's "condition of finiteness" as well as his "radiation condition.

538.56 1949


The media are assumed to be composed of many homogeneous isotropic layers bounded by parallel planes; the electromagnetic waves are plane and sinusoidal, with any orientation whatever. The problem reduces to the determination of the constants of integration for Maxwell's equations so as to satisfy the conditions at the limits. A system of 2p linear equations with 2p unknowns is solved, and the solution is determined by means of recurrence formulas.

538.56 1950


Discussion of the propagation of electromagnetic waves in the medium and not a discontinuity in the values of $\epsilon$ and $\mu$. The reflection coefficient for the stratified troposphere is calculated and also the variation of atmospheric conditions in the troposphere is shown. It is probable that this type of reflection can be observed with electromagnetic waves in the case of the troposphere, but since the variation of the refractive index is much greater than that of the values of $\epsilon$ and $\mu$, acoustic experiments might be more successful.

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538.56 1949


Polarization curves for different collision frequencies are given, together with hypothetical curves for electron-density distribution and collision frequency during the day and in the evening. Path lengths are derived for positive and negative pulses. Discussion of attenuation leads to the view that recently published values of the collision frequency are too high and that the greater part of the friction losses occurs in the E-layer, or even the D-layer. Results obtained with twin-loop receiving equipment 1,000 meters from a 2-kw transmitter are discussed. The effect of circulation of the signals, though even these were subject to fading during evening periods. Part 1, for which corrections are given, is 1381 of June.

538.56 1950


538.56 1949


An account of results obtained at Frederikshavn during the period 1942 to 1943. Records were obtained, on quickly moving film, of direct, and reverse signals from various distant stations. In some cases, the transmission path included both the north and south polar zones. The signal observed for frequencies above 1,000 kc may be explained by interference between two or more waves of slightly different frequency due to a Doppler effect. When two waves of nearly equal amplitude are combined, the pronounced beat minimum is obtained. The observed Doppler frequency shift of 5 to 30 cps at frequencies between 10 and 20 Mc indicates large movement of the reflecting layers in the polar zone.

538.56 1949


The low-level atmospheric conditions discussed in 1167 of May (McPete and Starneck) have little effect on the propagation of centimeter waves when the transmitter or the receiving path is high enough. The effect on the propagation of meter waves for any height of transmitter or receiver. Experimental results supporting this are shown graphically and discussed.

538.56 1949

Preliminary Analysis of Microwave Transmission Data Obtained on the San Diego Coast under Conditions of a Surface Duct—C. L. Pekeris and E. B. Davis (Jour. Appl. Phys., vol. 18, pp. 838-842; September, 1947.)

Transmission data for 61 and 170 Mc are analyzed by wave theory. Good agreement between theory and experiment was obtained for a range of 12 miles. Between 32 and 70 miles variations of intensity with height are as expected, but the theoretical horizontal decrement is less than that observed. Beyond 70 miles the received field tends to become uniform with height, and shows little horizontal attenuation. See also 521 of 1948,

538.56 1949

1229

RECEPTION

538.56 1950


The selectivity of a superregenerator is determined by the variation of instantaneous sensitivity due to modulation of the resonator conductor during a quench cycle. For optimum selectivity, a conductance wave form may be selected to give narrow skirt selectivity, or the noise of the second reflected tone may be narrowed to a degree determined by the rejection required for signals outside the pass band.

538.56 1951


An arrangement in which a band filter with feedback is used to increase the selectivity and sensitivity of a small receiver for the reception of distant station signals.

538.56 1949


The method is based on the fact that with quadrature rectification of single-sideband oscillations, only the first-order interference tones occur as nonlinear distortion, the amplitude of which is independent of the carrier amplitude. Since only the first-order interference tones remain after quadrature rectification of single-sideband oscillations without carrier, they can be completely eliminated in the reconstruction arrangement described, which consists of two balanced receiver units and a double-sideband carrier. The single-sideband oscillations with carrier are applied to one receiver and equal signals without carrier to the second; the two outputs are applied in opposition in the common output circuit.

Nonlinear distortion is thus eliminated so long as the modulation of the transmitter does not greatly exceed 50 percent. Signals of 50 percent modulation should have wide application in single-sideband technique, including multichannel cable communication, medium and long-wave broadcasting, and television.

538.56 1949

Wideband Frequency Discriminator Design—R. Senf (See 248.)

1229

Abstracts and References

621.396.621:621.396.82 2618 The Effect of Interference on Superregenerative Receivers—L. S. Gutkin. (Radioelektronika, vol. 2, pp. 62–76; February, 1949. In Russian.) Continuation of 1738 of 1948. The effect of interference, consisting of single and repeated impulses, on the output of a superregenerative receiver is investigated theoretically. The interference voltages at the output of the superregenerative circuit and of the low-frequency amplifier, and the effect of nonlinear operating conditions are discussed. A superregenerative receiver possesses a higher degree of discrimination with respect to impulse interference than an ordinary receiver. A general summary is given of types of interference.

621.396.621:621.396.911.13 2619 On Theoretical Signal-to-Noise Ratios in F.M. Receivers: A Comparison with Amplitude Modulation—D. Middleton. (J. Appl. Phys., vol. 19, pp. 334–351; April, 1948.) Signal-to-noise ratios at the output of a F.M. receiver are determined as functions of the input signal-to-noise ratio, clipping level, and filter characteristics when random (fluctuating) noise dominates the signal. Both narrow-band and broad-band FM are examined. Calculations are made for sinusoidal FM. The concept of signal-to-noise ratio is redefined. A comparison is made with AM reception using a half-wave linear rectifier. For signals less than 3 db above noise AM requires less input signal than FM for a given noise background and a given output. Only for signals at least 10 db above noise is FM superior to AM. Broad-band FM with heavy limiting is required in this case. Narrow-band FM, even with no limiting, is not better than AM at high signal levels. Limiting is detrimental for broad-band operation, whereas it is essential for broad-band operation, in which case it is of the utmost importance if response quality is important. Curves illustrating the average and mean-square signal and noise outputs, and signal-to-noise ratios for various conditions of operation, are included.

621.396.82 2620 Tolerable Mutual Interference of Two F.M. Broadcasting Transmitters—T. J. Wejers, (Tijdschr. ned. Radioenoot., vol. 14, pp. 61–72; May, 1949. In Dutch.) Selectivity measurements for three different FM transmitters are shown graphically and discussed. Conclusions as to the relative strengths and frequency characteristics of signals from two neighboring transmitters are presented with satisfaction of reception of the signals from one of them, are summarized.

621.396.82:537.523.3 2621 Corona Interference with Radio Reception in Aircraft—M. M. Newman. (Proc. N.E.C. (Chicago), vol. 4, pp. 91–95; 1948.) Oscillograms show typical corona pulses received on bare and on insulated antennas, and the serious effect on performance of both antennas and receiver circuits due to the sharp rate of rise of the pulses. Insulated antennas are recommended, but, even for serious interference, may be caused by other bare antennas and by charges developed on various parts of the aircraft structure.

621.396.82:551.594:551.577 2622 The Effect of Air Speed upon Precipitation Charging of an Airplane—H. J. Dana. (Proc. N.E.C. (Chicago), vol. 4, pp. 96–101; 1948.) Charging rate varies as the cube of air speed. This law was confirmed by experiment. Two types of discharge are discussed briefly.

621.396.822.029.63 2623 Impairment of Intelligibility by Noise in Decimetre-Wave Links—E. Dietrich. (Fermi-Forum, Z., vol. 2, pp. 173–178; June, 1949.) Discussion of the various factors involved establishes that speech intelligibility suffers little reduction due to the circuit, multipath, or interference usually met with. Decimetre-wave links can thus provide a satisfactory means of communication even without noise suppression.

621.396.828 2624 Interference-Free Reception by means of Aerials with Increased Down-Leads—W. Hormuth. (Frequenz, vol. 3, pp. 61–73; March, 1949.) Description of equipment suitable for either private dwellings or apartment houses. Particular attention is given to earthing arrangements and distribution systems.

621.396.828 2625 Circuit Design for Reduction of Hum—Dickerson. (See 2496.)


621.396(94) 2627 Some Aspects of the Overseas Telecommunications Services Operated by the Overseas Telecommunications Commission, Australia—A. S. McDonald. (Proc. I.R.E. (Australia), vol. 9, pp. 4–8; Discussion, pp. 8–9; April, 1948.) An outline is given of the formation and functions of the Commission. A review of the present facilities includes a reference to the overseas facsimile service. Equipment and techniques are briefly described.

621.396.41:621.396.19.16 2628 Terminal Equipment for Pulse-Time Multipler—A. M. Levine and D. G. Grier. (Proc. N.E.C. (Chicago), vol. 4, p. 131; 1948.) Summary only. Discussion of: (a) design of terminals using pulse-amplitude, pulse-time or pulse-width modulation; (b) ways of maintaining bandwidth while improving signal-to-noise ratio by arduous cross talk and distortion at the transmitting end and removing them at the receiving end; (c) methods of establishing base rates and timing of the channels; (d) modulation circuits for timed channels, (e) methods of maintaining synchronization and framing, (f) channel selection and demodulation, and (g) mechanical and electrical construction. A general description of various terminal units, with photographs, is included.

621.396.41:020.63:621.396.97 2629 Experimental Ultra-High-Frequency Multiplex Broadcasting System—A. G. Kandilion and A. M. Levine. (Proc. I.R.E., vol. 37, pp. 694–701; June, 1949.) An 8-channel high-fidelity system in which multiplex operation is achieved by adding parallel modulations. The main components, including modulator, transmitter antenna system and receiver, are described and their operating characteristics discussed.


621.396.619 2631 Contribution to the General Theory of Modulation and Demodulation for Any Type of Characteristic—Heymann. (See 2482.)

621.396.619:621.396.61 2632 "Auto-Anode" Modulation for Radio Broadcasting Transmitters—N. G. Kruglov. (Radioelektronika (Moscow) vol. 4, pp. 7–24; March and April, 1949. In Russian.) Description of a new modulation system developed by the author and now used at several broadcasting stations in the U.S.S.R. No high-power modulator is used, but certain tubes which generate radiofrequency oscillations also serve as anode modulation. The rf output is not less than that obtained with the usual anode-modulation methods. The theory of the method is discussed and various practical circuits are suggested. The modulation efficiency is almost double that for grid modulation; an overall transmitter efficiency of 45 per cent is possible.

621.396.619.13 2633 Some Developments in Frequency-Modulation Techniques—D. A. Bell. (Stoner J. Phys., vol. 18, pp. 62–71; April, 1949.) Discussion of implementation and modulation methods by which FM is obtained, the minimum value of frequency deviation, the method of obtaining the best performance, and a comparison of several practical methods and systems.


621.396.619.13:621.396.611.3 2635 The Application of Coupled Systems with Distributed Compensation to Frequency Modulation in the Ultra-High-Frequency Range—Tolstikov. (See 2475.)


621.396.65:621.396.3 2637 A 150-kc/s Carrier System for Radio Relay Applications—J. E. Broughton. (Trans. A.I.E.E., vol. 61, pp. 577–582; 1948.) Development of the Western Union Telegraph Co. Thirty-two voice-band communication channels, each 3,000 cps wide, are carried in the frequency range 4 to 147 kc. Tandem stages of group modulators maintain high spectrum efficiency without crystal filters. A crystal-controlled harmonic carrier supply ensures the frequency range 4 to 147 kc. narrow-band FM telephone circuits. See also 2471 above.


SUBSIDIARY APPARATUS


621.526 Electronic Circuits for Control of Clutch-Type Servomechanisms—F. E. Edwards, Jr. (Proc. N. E. C. (Chicago), vol. 4, pp. 54—58; 1948.) Brief description of clutch-type servo-mechanisms and desirable control-characteristics, with details of a circuit using the minimum time-lag and good stability.

621.526:621.317:621.396.2639.97 (43) 2639

1949

621.397.5:555.88 2650

An Improved Schmidt Plate—D. B. G. Hawkins. (Phil. Mag., vol. 40, pp. 670—679; June, 1949.) The profile of an improved aspheric plate is calculated which minimizes the off-axis errors. Detailed description of the given field. The color error is small compared with the off-axis errors for wide angular fields.

621.397.5:555.88 2651

Large-Screen Television—R. V. Little, Jr. (Proc. N. E. C. (Chicago), vol. 4, pp. 352—361; 1948.) Discussion of the basic elements of the direct-viewing system and the intermediate-filter system.

621.397.5:555.88 2652


621.397.5:535.88:791.45 2653


621.397.6:621.385.832 2654


621.397.6:778.3 2655

Video Recording Techniques—G. H. Gordon. (Tele-Tech, vol. 8, pp. 31—33, 63 and 29—31, 55; May and June, 1949.) A detailed description of equipment using 16-mm film for photographing direct from a television cathode-ray tube. Film speed is 24 frames per second. The arrangements for conversion from the standard American television rate of 30 frames per second are discussed. Standard films are satisfactory.

621.397.61 2656


621.397.62 2657


621.397.62 2658

New Television Receiver without Transformers, Design with Interchangeable Units—R. Achen. (TSP Pour Tous, vol. 25, pp. 129—133, 169—173, 205—212, and 244—246; April and August, 1949.) Details of a receiver comprising four units, each with its own power supply. Interaction between the four units can easily be avoided by operating them sufficiently far apart and any unit can be changed without affecting the others. The units are respectively (a) video receiver, (b) sound receiver, (c) time base unit, and (d) 7-kyv supply unit for the cathode-ray tube. Stagger tuning is used in the video circuit and blocking oscillators in the line and image timebases. Detailed circuit diagrams are given in all cases. Pertinent circuit for pulsals and line-phase refinements are discussed in an appendix.

TRANSMISSION

621.392.53:621.396.645 2659

The Surge Testing of High-Vacuum Tubes—H. J. Dailey. (Proc. N. E. C. (Chicago), vol. 4, pp. 187—199; 1948.) A method for initiating flash arcs and the results of such arcs under various conditions are described. Damage is a function of tube gas pressure, electrode configuration, and arc current. Suggestions are made for minimizing arc effects.

621.385 2662

Operating Conditions for the Optimum Working of Output Valves in High-Power Broadcast Amplifiers—N. I. Straladinov. (Radioelektronika (Moscow), vol. 4, pp. 5—15; January and February, 1949. In Russian.) The optimum working of a tube is secured when a 7-kv supply is used and A tangent current is obtained. Where E is the ratio of peak voltage swing to steady anode voltage, k is the ratio of maximum anode current to saturation current, and A is the ratio of anode current to permissible anode dissipation. Each of these conditions is examined separately and the necessary tube parameters and operating conditions for satisfying them are discussed. Practical suggestions are given.

621.385 2663


621.385.713 2654

A Transmitting Valve Cooler with Increased Turbulence of the Cooling Water—M. Smichowski. (Philips Tech. Rev., vol. 10, pp. 239—240; February, 1949.) Details of a cooling system using rings of jets round the anodes of high-power tubes.

621.385.029.62 2665

Glass Transmitting Valves of High Efficiency—R. L. White. (Proc. I. E. E., vol. 93, pp. 823—831; May, 1949.) A new tube for use in the 100-Mc. Range—E. G. Dornew (Philips Tech. Rev., vol. 10, pp. 273—281; March, 1949.) Reasons for preferring a coaxial cylindrical construction are given. Special features include a special spiral tube node of thundertan, a nonemissive grid, a graphite anode with horizontal cooling fins and shaped like a cotton reel, and a shield to reduce the temperature of the latter part of the all-glass envelope. In the trioedel, the shield is connected to the grid, so that the tube can be
used in grounded-grid circuits without neutralization, is 100 Mc. In the tuning of the shield the grid is connected to the screen-grid and neutralizing is necessary only above about 100 Mc. Details are given of the triode TB 250B. The anode dissipation is 135 watts and 125 watts respectively and an efficiency of 65 to 70 per cent at 100 Mc. Larger tubes, with dissipation up to 540 watts, are in course of development.

2621.385.02.21

2670

2621.385.02.21


2621.385.03.216

Pulse Emission Decay Phenomenon in Oxidized Tungsten Cathodes—T. Giovanella. (Proc. N.E.C. (Chicago), vol. 20, 1944-1947.) A cathode fatigue effect which sometimes occurs during high-power pulse emission from electrons in glass cathodes is attributed to sputtering. Experimental evidence of this is discussed. The decay is only obtained when electrons bombard a portion of the anode containing a poisoning agent. The question whether oxygen can cause the decay is still being investigated.

2621.385.2:537.525.92

Note on Space-Charge Considerations in Systems Including Cathodes—A. Coomes and J. C. Buck. (Proc. N.E.C. (Chicago), vol. 37, pp. 626-627; June 1949.) For test diodes of axial symmetry, the minimum variation in the slope of the space-charge line, and the values of the anode radius $r_a$ and cathode radius $r_c$ occur when $r_a/r_c = 3.16$.

2621.386.822

Current Fluctuations in a Plane Diode, taking account of Space Charge and Transit Time—A. Perez. (Comp. Rend., Acad. Sci. (Paris), vol. 228, 1949.) Starting from published tables (2977 of 1948), new tables have been derived covering the low-frequency range. In the dc-voltage region the fluctuations are attributed to space-charge. The results are shown graphically. The noise level is determined.

2621.386.232 2674

Testing Cathode Materials in Factory Production—J. T. Acker. (Proc. N.E.C. (Chicago), vol. 37, pp. 263-264; June 1949.) The determination of standard forms for reporting the results of tests of initial shrinkage, initial tube characteristics, and life tests, in order to determine a figure of merit for the material. See also 2675 below.

2621.385.232 2675

A Standard of Oxide-Coated Cathode-Core Material-Approved Test—R. L. McCormack. (Proc. N.E.C. (Chicago), vol. 37, pp. 683-687; June 1949.) Report of the work of a committee set up by the American Society for Testing Materials to prepare a specification for a standard diode test. Construction details of the standard diode are tabulated and illustrated; standards of test samples have been made available to interested companies. A cathode-temperature emission characteristic for each test lot is taken at anode voltage 40 volts, and compared with the same characteristic for a standard material. A figure of merit is deduced. Satisfactory correlation between tests by different companies was thus achieved; in one case surface contamination of a particular sample was established as a result of the tests. Additional tests are briefly discussed. See also 2674 above.

2621.385.232 2676


2621.385.832:537.291+538.691

Electron Beam Deflection. Part 2—Applications of the Small-Angle Deflection Theory—R. E. Hutter. (Jour. Appl. Phys., vol. 18, pp. 797-799, 1947.) The theory discussed in part 1 (609 of 1948) is applied to deflection fields produced by parallel plates, parallel wires, semi-infinite coplanar sheets, and bent plates. Results are shown graphically. An electrolyte-tank potential-drive device used for experimental verification is described. The deflection depends on the initial conditions and its relationship with field strength is slightly nonlinear. The resulting distortions of the spot and the crt pattern are calculated for several deflection fields and beam shapes, and methods of reducing such distortion are considered.

2621.385.832:621.397.6


2621.396.615.142

Pulsed Reflex Oscillator—(Electronics, vol. 22, p. 130; April, 1949.) A brief description of a rugged, miniature, v.m., external-cavity oscillator, with (Type F 5721), which can generate frequencies from 2,000 to 12,000 Mc. Operating characterstics are tabulated.

2621.396.645:537.311.33:621.315.59


MISCELLANEOUS

2621.387/39(43)

German Electronics in World War II—A. H. Sullivan, Jr. (Eng. Elect., vol. 68, pp. 403-409; May, 1949.) Much effort was devoted after 1943 to radar jamming devices, but with little success. Radio, radar, infrared, proximity fuses, tubes, ceramics, components, and navigational devices were also investigated. Apparatus of various kinds listed below is briefly discussed: (a) radar; Freya, Würzburg and Jagdflieger types, (b) radio; Type FuG 98 (early experimental), which was extremely selective and stable; Type FuG 24, a mass-produced equipment weighing only 3.5 lb, with frequency range 37 to 50 Mc; the Peil G 6 standard direction-finding receiver; (c) V-1 and V-2 rockets, (d) Tenschreiber and magnetophon magnetic tape recorders; (e) metal-lens, iron-core, and multi-element dielectric antennas; (f) a cathode-ray tube having a color center with different afterglow periods, (g) the Blauschicht, Mocos, and Krawinkel storage tubes; (h) carbon-filn resistors, (i) inductors formed by depositing metalization on ceramics, and (j) synthetic-nica and metalized-paper capacitors.

2621.396

Its practical cost astonishes users almost as much as its distinctive characteristics

**CHEMICAL COMPOSITION**
The nominal composition of commercially pure wrought Nickel is:
- Nickel* ..... 99.4%
- Copper ..... 0.15
- Iron ..... 0.15
- Manganese ..... 0.2
- Silicon ..... 0.05
- Carbon ..... 0.1
*Including cobalt

**PHYSICAL CONSTANTS**
- Specific Gravity: 8.89
- Density, lb. per cu. in.: 0.321
- Melting Point: 2615-2635°F, 1435-1445°C
- Specific Heat at (80-212°F): 0.130
- Heat Expansion Coefficient at (80-212°F): 0.000072
- Thermal Conductivity at (80-212°F): 420
- Electrical Resistivity at 32°F: 63 ohms/circ. mil. ft.
- Temperature Coefficient of Electrical Resistivity: 0.0022-0.0028
- Modulus of Elasticity: in tension, psi: 30,000,000, in torsion, psi: 11,000,000
- Poisson's Ratio: 0.31

**MECHANICAL PROPERTIES**
The following figures for Standard Cold Rolled Sheet are typical, though the figures will vary for different forms and tempers.
- Tensile Strength: 55,000-80,000 psi
- Yield Strength: (2% offset) 15,000-45,000 psi
- Elongation in 2 in.: 50-35%
- Rockwell B Hardness: 40-70

**AVAILABLE FORMS**
- Wire
- Bar
- Plate
- Pipe
- Rod
- Sheets
- Strips
- Seamless and Welded Tubing
- Sand and Precision Castings
- Clad-Steel Plate and Strip

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**It is a strange and interesting metal, Pure Nickel. A kin of both the base metals and the precious metals. Among all the elements, no other metal possesses its unique combination of so many different and uncommon properties.**

- It is highly resistant to corrosives that destroy many other metals—alkalies, many acids, salts, organic compounds, fumes.
- It has mechanical properties like those of structural steel.
- Yet it is so ductile that it can be worked into the most intricate and delicate shapes that are practical in metal.
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- Its special electronic properties make it a standard metal for electronic uses.
- It offers rare electrical and magnetostrictive characteristics that often give theoretical ideas a birth of practical value.
- It can be exposed to temperatures ranging into yellow heat and even hotter in the absence of sulphur.
- At sub-zero temperatures its strength increases without change in ductility and toughness.
- It is a standard metal for the cladding of steel, and as a base for gold, palladium and silver-clad products.

And one of the most valuable of all its features is the fact that Pure Nickel is a practical metal at a practical price.

Does it stimulate an idea of how you may find an easy answer to a difficult problem?

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COSMALITE COIL FORMS WITH COLLARS, made from our #96 COSMALITE, are a fibre base phenolic tubing, also of the highest quality at the lowest production cost. Specify that the collars be included and positioned on the core and thus secure a snug fit and an electrically stronger assembly.

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COSMALITE FOR SAFETY ECONOMY CAPACITY

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* (left) For high voltage power supply circuits of television receivers, specify our #102 COSMALITE COIL FORM. Note the sturdy, heavy-wall construction... the clean-cut, accurate, punched holes and notches. Ask about the many other advantages of these outstanding Cosmalite Coil Forms.

(right) COSMALITE COIL FORMS WITH COLLARS, made from our #96 Cosmalite, are a fibre base phenolic tubing, also of the highest quality at the lowest production cost. Specify that the collars be included and positioned on the core and thus secure a snug fit and an electrically stronger assembly.

COSMALITE is a proven product backed by over 25 years of experience.
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WIS completely equips all ground stations and aircraft with WILCOX radio

VHF AIR-BORNE COMMUNICATIONS
WILCOX TYPE 361A—50 watt transmitter, high sensitivity receiver, and compact power supply—each contained in a separate ½ ATR chassis. Receiver and transmitter contain frequency selector with provisions for 70 channels...ample for both present and future needs.

VHF GROUND STATION PACKAGED RADIO
WILCOX TYPE 378A—Complete with Type 364A, 50 watt transmitter, 305A Receiver, common antenna, telephone handset and loudspeaker, desk front, message rock and typewriter well. Type 411A LF Transmitter may be installed in the same cabinet for radiobeacon facilities.

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WILCOX TYPE 99A—Provides simultaneous transmission on LF, MHF, and VHF, frequencies. Housed in a single steel cabinet, the rectifier, modulator, remote control, and 4 transmitting channels combine to make the most compact multi-frequency transmitter in the 400 watt field.

WRITE TODAY...for complete information on all types of point-to-point, air-borne, ground station, or shore-to-ship communications equipment.

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KANSAS CITY MISSOURI

PROCEEDINGS OF THE I.R.E. October, 1949
**Type H-12 SIGNAL GENERATOR 900-2100 MEGACYCLES**

- Simplified
- Compact
- Portable

- 900-2100 megacycles, single band
- Directly calibrated, single dial frequency control
- Directly calibrated attenuator, 0 to -120 dbm
- CW or AM pulse modulation
- Internal pulse generator with controls for width, delay, and rate. Provision for external pulsing
- Controls planned and grouped for ease of operation
- Weight: 42 lbs. Easily portable—ideal for airborne installations
- Immediate delivery

*Built to Navy Specifications for research and production testing*

Write for specifications—investigate the advantages of this outstanding new instrument.

**STUDENT BRANCH MEETINGS**

(Continued from page 34A)

**CALIFORNIA STATE POLYTECHNIC COLLEGE—IRE BRANCH**

Election of Officers; April 11, 1949.
*Applications of Digital Computers,* by S. P. Frankel, Faculty of California State Polytechnic College; May 16, 1949

**UNIVERSITY OF DAYTON—IRE BRANCH**

"Electronic Control of Magnetic Coupling," by Tom Holloran, Student; April 5, 1949.
"Electronically Operated Voltage Regulators," by Bob Burtner, Student; April 19, 1949.
"Properties of Wide-Band Phase-Splitting Networks," by Frank Rancier, Student; April 26, 1949.
Election of Officers; May 10, 1949.

**UNIVERSITY OF LOUISVILLE—IRE BRANCH**

"Field-Strength Measurements," by D. C. Summerford, Technical Director of Radio Station WXLO; May 26, 1949.
Film: Crystal Clear, Industrial Measurements, and Stepping Along with Television; July 28, 1949

**MEMBERSHIP**

The following transfers and admissions were approved and will be effective as of October 1, 1949:

**Transfer to Senior Member**
Alberston, F. W., Dow, Lohner and Alberston, Munsey Bldg., Washington 4, D. C.
Brown, M. H., 14851 Wadkins Ave., Gardena, Calif.
Bunday, D. L., 2416 Carnation, Fort Worth 11, Tex.
Gihring, H. E., 5103 Westwood Lane, Merchantville, N. J.
Kerr, E. J., C.S.I.R., Radiophysics Laboratory, University Grounds, City Rd., Chippendale, Sydney, N.S.W., Australia
Muniz, R., 134 Mortimer Ave., Rutherford, N. J.
Stevens, S. S., 225 Overdale St., St. James, Man.
Canada
Teachman, A. E., 102 Byrner St., Jamaica Plain, Boston, Mass.
Zelle, J. F., 1227 Addison Rd., Cleveland 3, Ohio

**Admission to Senior Member**
Ainsworth, M. J., 17514 Rayner St., Northridge, Calif.
Gormley, R. S., 40 High St., Glen Ridge, N. J.
Holli, W. C., 458 77 St., Brooklyn 9, N. Y.

**Transfer to Member**
Anderson, A., 101 Osborne Ave., Baltimore 28, Md.
Cline, D. E., 75 Morris Ave., Manasquan, N. J.
Dasgupta, S. M., Government Engineering College, Jubulpore, C. P., India
Ganguly, S. R., 19 Chakrabora Rd., S. Calcutta 25, India
Harmatuk, S. N., 1575 Odell St., New York 62, N. Y.

**DEPENDABLE ELECTRONIC EQUIPMENT SINCE 1928**

**Aircraft Radio Corporation**
BOONTON, New Jersey
The following admissions to Associate were approved, to be effective as of September 1, 1949:

Allow J. 2212 Aerial St., Dayton 9, Ohio
Allen, C. A., 403 S. Third St., Homer, La.
Allen, R. C., Jr., 131 Washington Pl., Ridgewood, N. J.
Anderson, W. L., 185 Gundy Dr., Falls Church, Va.
Andrew, R. T., 121 East St., Wrentham, Mass.
Arch, J. R., 4145 Sheridan Rd., Chicago 13, Ill.
Bartolomei, H., 2043 Central Ave., Alameda, Calif.
Bellis, R. P., 50 Ganesvood Blvd., Staten Island, N. Y.
Best, J. G., 1142 W. Beatydale Ave., Elkhart, Ind.
Boyce, M. J., 4245 Magna Ave., East Chicago, Ind.
Butler, G. T., Jr., Rm. 1301, VMCA, Dayton 2, Ohio
Chester, C. L., 3902 Santa Ana, South Gate, Calif.
Clark, R. V., 14 Apple Tree Lane, Great Neck, L. I., N. Y.
Cohen, G. S., 100 Hashbrook St., Newburgh, N. Y.
Cook, K. H., 118 W. 67 St., Kansas City, Mo.
Ducastel, G., Rue D'Equirdeline No. 64, Leumont, Nord, France.
Ederly, J. J., 5810 Ave., Charleston 30, S. C.
Felix, E. A., Imperial Bank of India, Trichinopoly, Madras Province, India.
Filipe, P. P., 1313 S. Maple Ave., Berwyn, Ill.
Fox, M. L., 1212 W. Pratt Blvd., Chicago 26, Ill.
Garlock, J. W., 28 Westfield St., Providence 7, R. I.
Garrett, J. C., 2610 Wimoa St., Chicago 25, Ill.
George, J. E., 130 W. 38 St., Kansas City 2, Mo.
Gien, J. A., 411 N. Jackson St., Glendale 6, Calif.
Girling, K. W., Barnes Bldg., Bridgetown, Barbados, B.W.
Gormley, P. M., Kirkhill Farm, R.D. 2, German.
Guile, E. W., Florence Ave., and Teale St., Culver City, Calif.

Admission to Member
Bowles, M. E., 3056 McFarlin Blvd., Dallas 5, Tex.
Grobler, J. J., Box 31, Benoni, Transvaal, South Africa
Jacobson, R. E., Jr., 2814 E. Silver Ave., Albuquerque, N. Mex.
McGoy, F. G., 23A Savage St., Charleston 4, S. C.
Mitra, A. K., Communication Officer, Aero Communication Station, Dum Dum Airport, India
Secan, H., Kollman Instrument Division, 80-08 45 Ave., Elmhurst, L. I., N. Y.
Sherman, R. L., 324th Strategic Reconnaissance Squadron, McGuire Air Force Base, N. J.
Srivastava, S. S., Bayswater Hall, 46 Kennington Gardens Sq., London W. 2, England

The PROCEEDINGS of the I.R.E. October, 1949

THEIR THREE NEW DEVELOPMENTS ARE
Keeping
ASTATIC
Out in Front
IN THE
MANUFACTURE OF
PICKUP CARTRIDGES

1 GC CERAMIC CARTRIDGE
First major engineering stride in phonograph pickup cartridges employing ceramic elements since Astatic pioneered in this type unit last year. The GC is the first cartridge of its kind with replaceable needle. Takes the special new Astatic "Type G" needle—with either one or three-nil tip radius, precious metal or sapphire—which slips from its rubber chuck with a quarter turn sideways. Resistance of the ceramic element to high temperatures and humidity is not the only additional advantage of this new development. Output has been increased over that of any ceramic cartridge available. Its light weight and low minimum needle pressure make it ideal for a great variety of modern applications.

2 CQ CRYSTAL CARTRIDGE
An entirely new Astatic design, featuring miniature size and five-gram weight. Model CQ-1 fits standard 1/2" mounting and RCA 45 RPM record changers. Model CQ-1 fits RCA No. 2 Specifications for top mounting 45" mounting centers. Needle pressure five grams. Output 0.7 volts at 1,000 c.p.s. Employs one-nil tip radius, Q-33 needle. Cast aluminum housing.

3 LQD Double-Needle Crystal Cartridge
The LQD Cartridge—for 45, 33 1/3 and 78 RPM Records—quickly became the first choice of many of the nation's largest users, on the basis of comparative listening tests, and is, today, the PROVED TOP PERFORMER for turnover type pickups. Outstanding for excellence of frequency response, particularly at low frequencies. A gentle press with penknife removes ONE needle for replacement . . . without disturbing the other needle, without removing cartridge from tone arm. Gentle pressure snaps new needle into place. Available with or without needle guards. Stamped aluminum housing.

THE ASTATIC CORPORATION CONNEAUT, OHIO
IN CANADA: CANADIAN ASTATIC LTD., KERONIO, ONTARIO
Astatic Crystal Devices manufactured under U.S. Patent.

(Continued from page 36A)
MEASUREMENTS CORPORATION
BOSTON • NEW JERSEY

20 CYCLES TO 50 MC.
IN ONE INSTRUMENT!

THIS new Laboratory Standard is designed for the extremely wide frequency coverage of 20 cycles to 50 megacycles, employing two specially designed oscillators.

A low frequency oscillator, in the range from 20 cycles to 200 kilocycles, provides continuously variable, metered output from 0 to 50 volts across 7,500 ohms. This is sufficient for most measurements at audio and supersonic frequencies. It may also be used as the modulator for the radio frequency oscillator.

A radio frequency oscillator covers the range from 80 kilocycles to 50 megacycles. It provides metered output, continuously variable with an improved mutual inductance type attenuator, from 0.1 microvolt to 1 volt. This voltage range makes possible most receiver measurements including the determination of a.v.c. characteristics and interference susceptibility.

SPECIFICATIONS:

- **Frequency Range**: 20 cycles to 50 megacycles. (20 cycles to 200 kilocycles in four ranges; 80 kilocycles to 50 megacycles in seven ranges; plus one blank range.)
- **Frequency Calibration**: Direct reading dial, individually calibrated for each range.
- **Frequency Accuracy**: 20 cycles to 200 kilocycles, accurate to ± 1%. 80 kilocycles to 50 megacycles, accurate to ± 1.5%.
- **Output Voltage and Impedance**: 0 to 50 volts across 7,500 ohms from 20 cycles to 200 kilocycles. 0.1 microvolt to 1 volt across 50 ohms from 80 kilocycles to 50 megacycles. (Improved mutual inductance type attenuator.) The output voltage or impedance of either range can be changed by the use of external pots.
- **Modulation**: (80 KC-50 MC range) Continuously variable from 0 to 50% from 20 cycles to 200 kilocycles by internal low frequency oscillator or external source.
- **Harmonic Output**: Less than 1% from 20 cycles to 200 kilocycles, 3% or less from 20 kilocycles to 50 megacycles.
- **Leakage and Stray Field**: Less than 1 microvolt from 80 kilocycles to 50 megacycles.
- **Power Supply**: 117 volts, 50 to 60 cycles. 75 watts.
- **Dimensions**: 15" high x 19" wide x 12" deep, overall.
- **Weight**: 50 lbs.
(Continued from page 38A)

The following transfers to the Associate grade were approved to be effective as of August 1, 1949:

Berkenshamp, F. J., 732 W. Onondaga St., Syracuse, N. Y.
Carrell, R. M., 15 W. Coulter, Collingwood 7, N. J.
Drew, H. S., 34 West 175 St., New York 24, N. Y.
Geary, L. W., 5506 Parkland Ct., Washington 19, D. C.
Hajek, A. C., 120 S. Randall Ave., Madison 5, Wis.
Levy, B., 2025 Washington Ave., New York 37, N. Y.
Meyer, R. C., Jr., 9 Currius Pl., New Brunswick, N. J.
Meyerson, M., 150 Leslie St., Newark 8, N. J.
Mitchell, C. T., 202 Harrison St., La Porte, Ind.
Mrozowski, E. F., OTD, The Ordinance School, Aberdeen, Md.
Salisbury, L. J., 925 W. Seventh, N. S., Salt Lake City, Utah
Schoeder, D. W., 748 Fifth St., Columbus, Ind.
Strain, D. C., 4635 S.E. Hawthorne, Portland 15, Ore.
Thompson, L. H., 212 Birmingham Ave., Norfolk 5, Va.
Vincent, W. R., c/o Paul Hedler, Chestnut Ridge Rd., R. F. D., 2 Lockport, N. Y.
Wong, R. Y., 901 Talbot Ave., Santa Rosa, Calif.

(Continued from page 18A)

News—New Products

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

New Whip-Type Antenna

A new mobile antenna to cover the 75-meter band has been introduced by Premax Products, Niagara Falls, N. Y.

This design incorporates a special base-loading coil and a graduated or tapering whip of about 6 feet in length, giving a total over-all length of about 88 inches.

It is claimed that this type of antenna shows about a 6-th gain over the conventional "whip" types. Such a gain is of importance to the user as it equals the quadrupling of transmitter power and increases the effectiveness and range of radio operations, both on transmission and on reception, without involving any great expense for equipment. Reports state the use of this antenna has gone a long way in offsetting many of the difficulties encountered in the 75-meter band.

(Continued on page 41A)

PROCEEDINGS OF THE I.R.E. October, 1949

Shallcross Precision ATTENUATORS

MANY TYPES AVAILABLE FROM STOCK and through Shallcross parts distributors

Perhaps you've noticed how frequently Shallcross attenuators now appear in the finest audio or communications equipment? Or how often they are chosen for replacement purposes?

There's a reason! Improved design, materials and production techniques have resulted in a line that sets new, higher standards of attenuation performance for practically every audio and communications use.

Shallcross Attenuation Engineering Bulletin 4 gladly sent on request.

Shallcross Manufacturing Co.
Dept. PR-109 Collingdale, Pa.
3 CENTIMETER

(STD. 1”x ½” GUIDE UNLESS OTHERWISE SPECIFIED)

723 A/B Klystron mixer section with crystal mount, choke flange and iris flange put $32.50 ea.

TR/ATR Section for above with 724 ATR Cavity $7.50 ea.

TWIST 90 deg. 6” long $8.00

90 deg. Twist, 6” long $8.00

723 A/B Mixer—Beam Dual Oscillator Mount $12.00

2 Way Waveguide Coupler, type N, 11/16” x 11/16” $18.50

CG 100 (Waveguide Section 1½”) $10.00

TR/ATR Section, APS 15, for 724, with 724 ATR Cavity with 1824 and 724 tubes. Complete $21.00

Crystal mount

Stabilizer Cavity with bellows $2.50

3 cm. 180° bend with pressuring nipple $3.50 ea.

3 cm. 10° bend, 14” long 90° twist with pressuring nipple $4.00 ea.

3 cm. “S” curve 18” long $5.50 ea.

3 cm. “S” curve 18” long $5.50 ea.

3 cm. right angle bend, “E” plane 18” long $5.50 ea.

3 cm. Culler feed dipole, 11” from crystal mount to feed back $8.50 ea.

TWIST 45 deg. 3” long or specific length $1.50 ea.


DUPLExER section for 2K4 $1.00

CIRCULAR CHOLNE PLATES, gold $5

SQ. FLANGES, FLAT BRASS $1.50

APS-10 TR/ATR DUPLExER section with additional iris flange $3.00

FLAT WAVEGUIDE $4.00/ft.

TRANSITION 1½” to 3/4” 3/4”, 14” in $8.00

“X” BAND PREAMPLIFIER, consisting of 2723 A/B mixer stage feeding waveguide and TR/ATR Duplexer, section 1824. Complete $57.50

Random Lengths: waveguide, 6” to 18” L $1.00/ft.

WAVEGUIDE RUN, ½” x ⅞” guide, consisting of one 90° bend, 15” long $8.00

WAVEGUIDE RUN, ⅞” x ⅞” guide, 10” long $10.00

WAVEGUIDE Pressuring gauge section with 1/16th. gauge $1.00/ft. $1.50

45 DEG. TWIST 6” Long $8.00

12° SECTION 45 deg. Twist 90 deg. bend $4.00

15 DEG. BEND 10” long to cover $4.50

5 FT. SECTIONS cover to 90 deg., Silver Plated $14.50

FLEXIBLE SECTION $17.50

“E” PLANE BEND $12.50

“X” BAND WAVEGUIDE ½” x ⅞” OD 1½” ID $6.50

WAVEGUIDE 1½” x ⅞” OD 5” long, with 1” ⅛” b/t, 1.75 $7.50

WAVEGUIDE E 1½” x ⅞” OD, For 30 ft. $15.00

3 FLEX. SEC. sq. flange to Circ. Flange Adapter $7.50

724 TR TUBE (41 TR) $7.50

SWR MEAS. SECTION, with 2 type “N” output ports, full wave output, full wave apart. Bell style, Silver plated $10.00

ROTARY JOINT with slotted section and center output, 12” long to cover $4.50

WAVEGUIDE SECTION, 12” long to cover to 90 deg. bend with slotted section and center output $7.50

SLUG TUNER/ATTENUATOR, W. E. guide, gold plated $4.50

TWIST 90 deg. 5” long to cover 1½” x ⅞” $6.50

WAVEGUIDE SECTIONS ⅞” x ⅞” long: Silver plated 3 inch $7.50

ROTARY JOINT cover to 90 deg. $17.50

ROTARY JOINT cover to 90 deg. with deck mounting $17.50

3 cm. mitered elbow “E” plane unslotted $4.50 ea.
The Model A-1 crystal cartridge is newly developed miniature in size and ideally adapted for tone arms of modern styling and function. It mounts either a 1-mil or 3-mil point stylus or both, making it applicable to all types of recordings in use today. Tracking pressure is only 7 grams — meeting the requirements of 33⅓ and 45 RPM as well as the standard 78 RPM records. Adaptor brackets supplied for mounting in arms originally designed for standard cartridges.

News—New Products

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

Midget-Can Electrolytic Capacitors

The latest type PRS midget-can electrolytic capacitor in reduced sizes are being marketed by Aerovox Corp., 740 Belleville Ave., New Bedford, Mass.

Illustrated is the type PRS 450 volt, 8 μf "Dandee," which is placed next to a cigarette for comparison. The size is 1½ inches long by 13/16 inch diameter.

These electrolytics are available in single-section ratings from 25 to 700 volts dc working, 4 to 100 μf and in dual-section from 25 to 450 volts dc working, 8-8 to 100-100 μf.

In the high-capacitance low-voltage series, they have ratings from 6 to 25 volts dc working, 100 to 2,000 μf.

TV Lead-in Clamps

A new line of TENNA-CLAMPS, stand-off insulators, designed to clamp on to masts, cross-arms, gutters, and guy-wires for supporting TV lead-in lines is now being marketed by Mueller Electric Co., 1583 E. 31st St., Cleveland 14, Ohio.
**Correction Notice**

A dc microammeter and magnetic amplifier, Type 100, has been designed to measure low dc by W. S. MacDonald Co., Inc., 33 University Rd., Cambridge 38, Mass.

Incorrectly described in our August issue, the Type 100 has an input resistance of 50 ohms and a sensitivity of 1 microampere full scale. Input may be overloaded 2 amperes without damage.

This instrument may be used as a dc amplifier, as such it will actuate a 1-ma, 1,400-ohm recorder directly.

**Recent Catalogs**

- In the May issue of *The Experimenter*, a high power, low-speed (600 rpm) stroboscope, and a versatile voltage divider are discussed. The organ is published by General Radio Co., 275 Massachusetts Ave., Cambridge 39, Mass.

- A desk-size Army-Navy Connector Specifications (AN-C-501) Chart with the latest insert arrangements shown in detail at half scale for use by aircraft, radio, communication engineers, and designers, by Cannon Electric Development Co., Catalog Dept., 3209 Humbolt St., Los Angeles 31, Calif.

- An 8-page bulletin with information and technical data on dc motors for timing applications, and the performance characteristics of Model 9200 dc motor, may be obtained by writing to E. B. Hamlin, Haydon Mfg. Co., Torrington, Conn. Specify Engineering Bulletin #1.

- A 35-page catalog August #854 in color with complete design specifications and dimensions of all their screw products, by The Bristol Co., Mill Supply Div., Waterbury 91, Conn.

(Continued on page 41A)
NEW Andrew MULTI-V FM ANTENNA

<table>
<thead>
<tr>
<th>TYPE</th>
<th>NO. OF BAYS</th>
<th>POWER GAIN</th>
<th>PRICE</th>
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<tr>
<td>1308</td>
<td>8</td>
<td>7.3</td>
<td>$2800</td>
</tr>
<tr>
<td>1304</td>
<td>4</td>
<td>3.7</td>
<td>850</td>
</tr>
<tr>
<td>1302</td>
<td>2</td>
<td>1.6</td>
<td>320</td>
</tr>
</tbody>
</table>

This table shows you why the new Andrew Multi-V is your best FM antenna buy!

NOW! Minimize your investment in equipment. Get top performance for only half the cost. The new Andrew Multi-V FM antenna is made and guaranteed by the World’s Largest Antenna Equipment Specialists. It’s another Andrew “First.”

**FEATURES**

* Twice as much power gain per dollar as any other FM transmitting antenna!
* Top performance, yet half the cost of competitive antennas.
* Side mounting construction permits installation on towers too light to support heavier antennas.
* Circular radiation pattern.
* Factory tuned to required frequency — no further adjustments necessary.

It will pay you to use the Andrew Multi-V Antenna on your FM station. Write for Bulletins 86 and 186 for complete details TODAY.

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 434)

• • • A new Engineering Data Sheet #558, which describes the latest types of coil tube fasteners, and shows typical assemblies, may be obtained from The Palnut Co., 94 Cordt St., Irvington, N. J.

• • • A 24-page booklet, 2IP8-46, describing insulation resistance testing with a “Megger,” and a new color catalog, Bulletin 21-05-46, to be used in selecting from 10 types of “Megger” instruments covering ranges from 0.01 ohm to 10,000 megohms, by James G. Biddle Co., 1316 Arch St., Philadelphia 7, Pa.

• • • An illustrated color folder with a description of a wide line of crystals, and the FS-344 Frequency Standard, is at present available from James Knights Co., Sandwich, Ill.

• • • A 36-page bound catalog of instrumentation for radioactivity measurement, describing a wide range of instruments and accessories with information about application is ready for distribution by Nuclear Instrument & Chemical Corp., 223-233 W. Erie St., Chicago 10, Ill.

New Voltmeters Check Up
To 30,000 Volts

Inexpensive, high-voltage voltmeters checking voltages up to 30,000 volts are now being manufactured by Industrial Devices, Inc., Edgewater, N. J.

The Hi-Volt is available in two models. Model 500 is designed for checking the voltage output of transformers, such as for oil burner ignition, gas-discharge display signs, etc. Model 520 is for electronic high-voltage uses, such as television, oscillographs, etc.

The High-Volt utilizes a neon-lamp indicator in place of the usual meter movement. The knob is turned until neon lamp extinguishes. Voltage is then read directly off scale where pointer rests.

Heavily insulated and using a multi-megohm multiplier, the High-Volt test prod is 7 inches long, thus assuring user of sufficient reach to keep away from “hot” leads. Model 500 draws less than 1 milliampere, while Model 520 draws less than 300 microamperes at full-scale reading.

(Continued on page 454)
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 44A)

Package Circuit Assembly
For TV Receivers

Following the trend of component manufacturers endeavoring to reduce production time and save space, a new package circuit assembly, known as the "Vertical Integrator Network," has been produced by Centralab, Div. Globe-Union, Inc., 900 F. Keefe Ave., Milwaukee, 1, Wis.

This printed electronic circuit consists of four capacitors and four resistors, terminating in three leads, as against sixteen connections when using standard components. In size the assembly is 1 9/16 inches long, 1 1/4 inches wide, and 3/16 inch thick.

A condensed version, which covers the balance of TV circuits, is a smaller model with three capacitors and three resistors.

High-Voltage Controls

Increased insulation, necessary with controls used in TV, oscillograph, and other high-voltage circuits, may be obtained (on special order) by use of an improved coupler feature with most types of controls manufactured by Clarostar Mfg. Co., Washington St., Dover, N. H.
OSCILLOSYNCHROSCOPE
Model OL-15B
Designed for maximum usefulness in laboratories doing a variety of research work, this instrument is suited to radar, television, communication, facsimile, and applications involving extremely short pulses or transients. It provides a variety of time bases, triggers, phasing and delay circuits, and extended-range amplifiers in combination with all standard oscilloscope functions.

These features are important to you
- Extended range amplifiers: vertical, flat within 3 db 5 cycles to 5 megacycles. Full tube deflection; horizontal, flat within 1 db 5 cycles to 1 megacycle.
- High sensitivity: vertical, 0.05 RMS volts per inch; horizontal 0.1 RMS volts per inch.
- Single-sweep triggered time base permits observation of transients or irregularly recurring phenomena.
- Variable delay circuit usable with external or internal trigger or separate from scope.
- Sawtooth sweep range covers 5 cycles to 500 Kilocycles per second.
- 4,000-volt acceleration gives superior intensity and definition.

For complete data, request Bulletin RO-910

SWEEP CALIBRATOR
Model GL-22A
This versatile source of timing markers provides these requisites for accurate time and frequency measurements with an oscilloscope:
- Positive and negative markers at 0.1, 1.0, 10, and 100 micro-seconds.
- Marker amplitude variable to 50 volts.
- Gate having variable width and amplitude for blanking or timing.
- Trigger generator with positive and negative outputs. Further details are given in Bulletin RC-910.

STANDING WAVE RATIO METER AND HIGH GAIN AUDIO AMPLIFIER
Model TAA-16
- Standing wave voltage ratios are read directly on the panel meter of this sensitive, accurate measuring instrument.
- Frequency range 500 to 5,000 cycles per second.
- Two input channels with separate gain control for each.
- "Wide-band" sensitivity 15 microvolts full scale.
- "Selective" sensitivity 10 microvolts full scale.
- Bolometer/crystal switch adjusts input circuit to signal source.

Write for Bulletin RA-910 containing full details of this useful instrument.

BROWNING Laboratories, Inc.
Winchester, Mass.
ENGINEERED FOR ENGINEERS

News—New Products

Designated as Type 56-125 high-voltage coupler, this control has a plastic straight-through shaft in place of the previous insulating strip joining separate sections of the metal shaft. An insulation tube further isolates the control proper from its mounting bushing.

Control-to-ground breakdown rating is better than 10,000 volts.

Adding Component for Analogue Computers

The K3-A adding component, which delivers from one to four input signals, and which has an adjustable additive steady level up to 20 per cent of the range, is being offered by George A. Philbrick Researches, Inc., 230 Congress St., Boston, Mass.

Either or both outputs may be employed, giving plus or minus the direct sum of the inputs in use. Through tandem interconnection, any number of signals may be combined in sums or differences.

In the K3 analogue line, each component is self-contained and performs an individual function as part of a computing system. Each measures approximately 4X5X7 inches, weighs about 5 lbs., has 5-pin input and output connectors for 110 volts, 60 cps.

New Impedance Meter

For the measurement of impedances from 0.1 to 100,000 ohms over a wide frequency range, the "Impedometer" has been developed by Edward S. Shepard, Sr. of Boston College and is being manufactured by Electrodyne Co., 899 Boylston St., Boston 15, Mass.

(Continued on page 47A)
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.
(Continued from page 46A)

The "Impedometer" provides a means for comparing the voltage drop across the unknown impedance and across a resistive standard, when the same current is present in both circuits. In making this comparison, the meter is used in conjunction with a suitable oscillator and vacuum-tube voltmeter.

Microscope for Disk Recording Analysis

The Model 231 microscope, with self light source and reticle for groove analysis and surface quality testing, has been placed on the market by Clarkstan Corp., 11927 W. Pico Blvd., Los Angeles 34, Calif.

Magnification of either 20 or 40 times is offered. For calculation of lines per inch, width of groove, or groove-to-land ratio, a reticle has been incorporated for direct measurement with divisions of 0.004 inch or 0.002 inch for the respective powers.
(Continued on page 48A)

PROCEEDINGS OF THE I.R.E. October, 1949
NOW.
R. F. ATTENUATION NETWORK
FOR YOUR WORK

To meet the increasing needs for accurate, dependable instruments to attenuate UHF, The Daven Company now offers RF attenuation boxes. These units are notably compact, provide a wide range of attenuation and are moderately priced.

---SPECIFICATIONS---

CIRCUIT: Pi network.
STANDARD IMPEDANCES: 50 and 73 ohms. Other impedances on request.
RESISTOR ACCURACY: ± 2½% at D.C.
IMPEDANCE ACCURACY: Terminal impedance of loss network essentially flat from 0—225 MC.
MOUNTING: Cabinet Type or Rack Mounting Available.
NO. OF STEPS: Types: 640, 641, 642, 643 8 Push Buttons
Types: 650 and 651 10 Push Buttons
RECEPTACLES: Army-Navy Types UG-58/U or UG-185/U Supplied

<table>
<thead>
<tr>
<th>TYPES</th>
<th>IMPEDANCE</th>
<th>LOSS</th>
</tr>
</thead>
<tbody>
<tr>
<td>RFA 640 &amp; 641</td>
<td>50 &amp; 73</td>
<td>1, 2, 3, 4, 10, 20, 20, 20, 20, DB STEPS (80 DB TOTAL IN 1 DB STEPS)</td>
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<tr>
<td>RFA 642 &amp; 643</td>
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<td>2, 4, 6, 8, 20, 20, 20, 20, DB STEPS 100 DB TOTAL IN 2 DB STEPS</td>
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<tr>
<td>RFA 650 &amp; 651</td>
<td>50 &amp; 73</td>
<td>1, 2, 3, 4, 10, 10, 10, 20, 20, 20, DB 100 DB TOTAL IN 1 DB STEPS</td>
</tr>
</tbody>
</table>

---APPLICATIONS---

- In signal and sweep generators.
- In field strength measuring equipment.
- Nucleonic and atomic research.
- Television receiver testing.
- Wide-band amplifiers.
- Pulse amplifiers.
- Any application where attenuation of UHF is required.

For additional information write to Dept. 1E-8

THE DAVEN CO.
191 Central Avenue Newark, N.J.

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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 47A)

Power Relay

Described as having been manufactured to meet aircraft specifications and, therefore, capable of meeting rigid control requirements, the Type “DO” power relay is available from American Relay & Controls, Inc., 4925 W. Flournoy St., Chicago 44, Ill.

This relay is available in contact combinations up to 4-pole, double-throw. The terminals and contact arms are mounted on moulded phenolic panels. The coil is of cellulose acetate sealed construction.

Contact rating is 10 amperes at 115 volts ac, noninductive, or 10 amperes at 32 volts dc, noninductive. Coils are available for continuous duty for operation up to 230 volts ac, 60 cps, or 115 volts dc.

Magnetic Tape Eraser

A device for erasure of reels of magnetic tape without running the tape past the erase head has been developed by the Accessories Div., Amplifier Corp. of America, 598-1 Broadway, New York 13, N. Y.

Described as the “Magnerasor,” the manufacturer claims that this eraser removes all normal and overloaded signal from all types of tape, and is able to lower the residual noise level as much as 3 to 6 db below that of unused tape.

The erasure is accomplished by placing the “Magnerasor” on the reel and moving it around the circumference of the reel. The life of the tape is lengthened by elimination of physical contact with the erase head.

(Continued on page 49A)
Wobbulator Signal Generator for Television, Radar, and UHF

The Model 705, a sweep-frequency type which incorporates a 5-inch cathode-ray tube with associated sweep and gate circuits, traveling detector, signal amplifier, and 100-db attenuator for control of the amount of signal into the circuit under test, is being produced by Canoga Corp., 14315 Bessemer St., Van Nuys, Calif.

A high-pass filter following the crystal mixer removes the direct rectified 100-kc modulation frequency and leaves the desired beat frequency. The output is developed across a 50-ohm cable termination, and is about 0.1 volt when the total rectified crystal current is 5 ma.

The Model 705 has a three-stage amplifier with a 50-kc bandwidth. The deflection sensitivity of the amplifier is such that 100-microvolt signal input produces a 1-inch deflection on the screen.

Flexible Inert-Arc Welder

A miniature Inert-arc electrode holder, featuring a flexible front-end assembly made of malleable copper tubing surrounded by a sheath of silicone rubber so that it can be bent in any direction to reach difficult places, has been announced by Welding Div., General Electric Co., Schenectady 5, N. Y.

Specifically designed for the fluxless welding of nonferrous metals in gauges from #16 to #40 (0.0625 to 0.003 inch), the holder is available in two models: one for 0.01- and 0.02-inch tungsten electrodes, and the other for 0.04- and 0.06-inch tungsten electrodes. It is rated at 40 amperes continuous, and can be used with either ac or dc supply.

(Continued on page 554)
PROJECT ENGINEERS
Real opportunities exist for Graduate Engineers with design and development experience in any of the following: Servo-mechanisms, radar, microwave techniques, microwave antenna design, communications equipment, electron optics, pulse transformers, fractional h.p. motors.
SEND COMPLETE RESUME TO EMPLOYMENT OFFICE.

SPERRY GYROSCOPE CO.
DIVISION OF THE SPERRY CORP.
GREAT NECK, LONG ISLAND

Positions Available for
ELECTRONIC ENGINEERS
with Development & Design Experience in
MAGNETIC TAPE RECORDING
MICROWAVE COMMUNICATIONS
SONAR EQUIPMENTS
Opportunity For Advancement Limited Only By Individual Ability
Send complete résumé to:
Personnel Department
MELPAR, INC.
452 Swann Avenue
Alexandria, Virginia

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.

ENGINEER
Large active midwestern quartz crystal plant needs first class quartz crystal engineer, well grounded in theory and with complete experience in manufacturing and testing procedures all types quartz units. Send complete detailed information on experience and background in first reply. Salary open. Our employees know of this ad. Box 574.

ENGINEER
Excellent opportunity for a man with experience in servo mechanisms and circuit theory work. Opportunity for graduate work available. Write Box 30, State College, Pennsylvania.

DEVELOPMENT PHYSICIST ENGINEER

DIRECTOR OF TRAINING
Southern, G. I. approved, private vocational school desires capable man to take charge of teaching courses in radio and television. Salary $300 to $350 per month. Scholastic hours. Teaching experience preferred. Write Mr. J. D. Muse, P.O. Box 505, Atlanta, Georgia.

MECHANICAL ENGINEERS AND DESIGNERS
Fully qualified. Acquainted with large electromechanical devices. Write Mr. D. A. Murray, Mount Denis, Toronto 15, Ontario.

ELECTRONIC ENGINEER
Fully qualified. At least five years laboratory experience in circuit design and radio physics. Canadian national preferred. Write Mr. D. A. Murray, Mount Denis, Toronto 15, Ontario.

DEVELOPMENT ENGINEERS
All grades with degrees and experience in design and development of high quality instruments for research in physics, chemistry, etc. Applicant will be required to design and develop electrical, electronic and mechanical instruments for the nuclear field. Salary commensurate with ability to produce a final working model from the idea state. Box 577.

(Continued on page 51A)

WANTED
ELECTRICAL ENGINEER
for DEAN of ELECTRICAL SCHOOL of a SMALL MID-WESTERN COLLEGE

• An interesting and challenging appointment is available to a man having the desire and experience to enter or broaden his opportunities in the field of Education.
• The locality and facilities are the finest with good housing available on the campus and the position is one which can be permanent and satisfy a desire for personal achievement.
• In applying for this appointment outline in complete detail—teaching and administrative experience, scope of research, educational background, degrees. Experience in electronics will be given extra consideration.
• An interview can be expedited if you will include personal information such as family status, and salary desired. Our faculty knows of this staff opening and your reply will be held in full confidence.

Write care
Box 581
Proceedings of the I.R.E.
1 East 79th St.
New York 21, N.Y.

(Continued on page 51A)

PROCEEDINGS OF THE I.R.E., October, 1949
ELECTRONIC TECHNICIANS

For work in laboratory, assembling, wiring and testing of precision electronic design models. Applicants must have at least three years of similar experience and must be capable of producing the highest quality work. Box 577.

ENGINEER

Wanted, new electronic ideas company with capital and manufacturing facilities is seeking new electronic products, inventions, or ideas to expand commercial business. Liberal arrangements with inventors. Box 578.

PHYSICISTS AND ENGINEERS

This expanding scientist-operated organization offers excellent opportunities to alert physicists and engineers who are interested in exploring new fields. We desire applicants with experience in the design of electronic circuits (either pulse or c. w.), computers, gyros, or precision mechanical instruments. A few openings for Junior Engineers and Technicians also exist. This company specializes in research and development work. Laboratories are located in suburbs of Washington, D.C.

JACOBS INSTRUMENT CO.
4718 Bethesda Ave.
Bethesda 14, Maryland

Give enough!

PROCEEDINGS OF THE I.R.E. October, 1949

Sorensen and company, inc.
375 Fairfield Ave., Stamford, Connecticut
Positions Wanted
By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants, and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The Institute publishes free of charge notices of positions wanted by I.R.E. members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is one per year. The Institute necessarily reserves the right to decline any announcement without assignment of reason.

ENGINEER

ENGINEER

TECHNICAL WRITER
Development engineer available for writing assignments in radio, television, and the allied arts. Technical articles, instruction booklets and similar work treated with strictest confidence on commission basis. Box 300 W.

FIELD ENGINEER

ELECTRICAL ENGINEER
B.S. in Radio Engineering, and graduate work in Electrical Engineering. Former communications and fighter control officer. Desires research in electronics or electrical engineering. Married. 1 child. Location immaterial. Box 302 W.

ENGINEER MANAGER
Harvard MS in communication engineering. Directed important and extensive wartime Loran project in Pacific. Recently Development Division Engineer Manager. Desires permanent and responsible position with progressive organization. Box 303 W.

*(Continued on page 53A)*

PROCEEDINGS OF THE I.R.E., October, 1949
Positions Wanted

(Continued from page 524)

RADIO ENGINEER
B.S. in radio engineering, one course to complete for B.S.E.E. Married, no children. Age 28. 3½ years experience in telephone carrier and VHF installation and maintenance. Desires research or design position anywhere in United States. Box 305 W.

TELEVISION ENGINEER
Graduate American Television Institute of Technology May 1949 with B.S.T.E. Age 35. Married, 1st class F.C.C. license. 5 years electrical maintenance, 3 years radar maintenance, 3 years radio servicing, desires position as TV station engineer. Prefer east or midwest. Box 306 W.

ELECTRONICS ENGINEER
B.S. physics, Age 27. Married, no children. 2 years industrial electronics research and development. Some guided-missile development. 3 years Army Radar developing and maintenance. Anywhere in United States; will also consider foreign position. Box 307 W.

ELECTRONIC DESIGN ENGINEER—ELECTRONICS INSTRUCTOR
B.S., 1947 University of Chicago. Age 32. Married, 3 years experience in design of radar equipment with eastern manufacturer. 2½ years experience in Navy as radar maintenance officer. Have had teaching experience in physics. Desires position as electronic design engineer in midwest, or electronics instructor in midwestern college. Box 308 W.

ENGINEER
B.S.E.E. 1948 electronics option Georgia School of Technology, Age 22. 1½ years experience electronic technician U.S. Navy. 1st class radio telephone F.C.C. license. 1 year industrial experience in design and development of radar components. Member Eta Kappa Nu and Tau Beta Pi. Desires work in development. Box 309 W.

JUNIOR ENGINEER OR TECHNICAL WRITER

RADIO BROADCAST) ENGINEER
B.S.E.E. 1939. Over 8 years' experience in broadcast work includes AM, FM, Television, 50kw, transmitters. Last position chief engineer construction large station with complex directional antenna. Signal Corps radar staff officer during war. Box 311 W.

DEVELOPMENT ENGINEER
Eight years broad experience, research and development. Servos, auto controls, analog computers, industrial electronics. Excellent theoretical ability well balanced by laboratory and experimental work. High scholastic standing in college and research position as Senior Development Engineer or Assistant Director of Research. Box 321 W.

A Complete Line of
PRODUCTION TEST EQUIPMENT
for TV Manufacturers

Tel-Instrument has designed and provided the production test equipment for many of the major TV manufacturers. A complete line of instruments designed to be unusually critical in the testing of TV receivers is available. They are the result of the wide practical experience of Tel-Instrument engineers plus a complete understanding of the production problems of TV manufacturing.

TYPE 2120
R.F. PICTURE SIGNAL GENERATOR

TYPE 1200
12 CHANNEL
R.F. SWEEP GENERATOR
Intended for precise adjustment of R.F. head oscillator coils and R.F. band pass circuits. Pulse type markers at picture and sound carrier frequencies extend to zero signal reference base line. Accuracy of markers 0.02% of carrier frequency. 12 to 15Mc, sweep on all channels. Max. 1V peak output across a 75 ohm line. Provisions for balanced input receiver. Instant selection by push button.

TYPE 1900
CRYSTAL CONTROLLED
MULTI-FREQUENCY GENERATOR
A 10 frequency, 400 cps. modulated crystal controlled oscillator, ideal for production line adjustment of stagger tuned I.F. amplifiers. Available with crystals ranging from 4.5 to 40 Mc. Output frequency accurate to 0.01%. Immediate push button selection of frequency. Output attenuator range 5V to 500 microvolts. Self contained regulated power supply.

Write for Detailed Engineering Data Sheets.
Quality Electronic Equipment Deserves

TRIAD "HS" TRANSFORMERS

Volume production of TRIAD HS (hermetically sealed) Transformers to JAN specifications has enabled TRIAD to lower costs to little more than that of ordinary cased types. TRIAD HS Series Transformers feature:

TRIAD Hermetic Seals—sturdy brass studs, molded in low-loss plastic, eliminate mechanical weaknesses often found in other designs.

Wide Frequency Range—Nickel alloy laminations, low capacity and low leakage reactance windings, plus balanced designs, result in a frequency range from 20-20,000 cycles ± 1 db.

Reduced Field Pickup—Triad GP series cases, drawn from annealed nickel alloy, reduce stray field pickup by as much as 95 db.

Small Size—HS-1 line input transformers with 95 db shielding and 20-20,000 cycle frequency response, in case only 1½" x 1½" (base dimensions) x 2½" high above chassis.

Low Distortion—Triad output coils employ large cores of the best magnetic alloys, with coils of low resistance and low leakage reactance, to approach full output at all frequencies with low distortion. Output transformers may be included in feedback loops using 30 db of feedback.

Complete Line—All types of audio coils, power coils, reactors, supplied in matching HS Series construction.

Write for Catalog TR-49

TRIAD TRANSFORMER MFG CO

2254 Sepulveda Blvd.
Los Angeles 64, Calif.

Positions Wanted

(Continued from page 53A)

ENGINEER
B. E. E., Cum Laude, C. C. N. Y., February 1948; Tau Beta Pi and Eta Kappa Nu. Age: 24. One and a half years design, development and production experience as project engineer. Main field—Antennas. Desires position in New York Metropolitan Area. Box 322 W.

ENGINEER

JUNIOR ENGINEER OR LABORATORY TECHNICIAN
Graduate of R. C. A. Institute Technology Course. Four years’ commercial experience in electronic laboratory technique. Worked for Alexander Fowler and Allen B. Dumont Laboratories. Former Air Corps instrument instructor and specialist, Hold 1st class F. C. C. radio and telephone license. Experienced in all phases of laboratory work on video or electronic circuitry. Age: 27. Married. Desires New York City area or Long Island. Box 325 W.

RADIO ENGINEER

MICROWAVE ENGINEER
B. E. E. 1943, graduate evening student. Married, age 27. 3½ years’ research and development experience with microwave transmission components and systems. 2 years Army P. P. M. radar link work. Desires research or development work in New York City. Box 328 W.

ENGINEER
B. S. in radio engineering. 3 years airborne radio and radar maintenance with U. S. M. C. 19 months audio repair with Sound Scriber Distributor. Age 27, married, 1 child. Desires work in U. H. F. field. Box 329 W.

ELECTRICAL ENGINEER

JUNIOR ENGINEER
B. S. E. E. Columbia University, June 1949, Age 28, single. Desires promising starting position in design development or production, anywhere in United States. Box 331 W.
Positions Wanted

(Continued from page 54A)

COMMUNICATIONS ENGINEER
B.S.E.E. 1947. 2 years carrier telephony, Signal Corps radio-link. Age 27, married, 2 children. Now employed in Boston, wants research, design, station construction, sales engineering, teaching, or technical writing in central to southern Maine. 2 years design of high-frequency and microwave antennas. Box 332 W.

ELECTRICAL ENGINEER
B.E. registration applied. Age 28, single, 3 years' communications experience. 1 year servo-mechanism experience. Desires position in servo-mechanism in growing concern. Location, midwest or east. Box 333 W.

ELECTRONIC TECHNICIAN
High school graduate. 2 years U.S. Coast Guard radio and radar school. 2 years at RCA Institutes, 5 years' experience in radio and radar maintenance and installation with U.S. Coast Guard. 3 years with American Airlines as radar technician in transmitter and experimental radar laboratory. Box 334 W.

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

(Continued from page 90A)

Increased Wattage for Soldering Guns

With the aid of a new method of transformer core winding, a more efficient soldering gun is now being manufactured by Weller Mfg. Co., 808 Parker St., Easton, Pa.

Previously held to 135 watts, due to bulk and weight of the transformer, the new gun has a 250-watt capacity by use of a machine that winds the transformer core from strip steel onto the primary and secondary coils.

Described as Model WD-250, this gun has no increase in size, a few ounces increase in weight, and a heating time of 5 seconds.

Universal Bridge

The #1150 Universal Bridge has been added to the laboratory instrument line of electronic equipment of the Freed Transformer Co., 1718-36 Weirfield St., Brooklyn 27, N.Y.

(Continued on page 56A)

PROCEEDINGS OF THE I.R.E. October, 1939

DESIGN TIPS FROM CHIEF ENGINEER FLEXY

S.S.WHITE FLEXIBLE SHAFT

HERE'S HOW TO GAIN GREATER LEEWAY IN YOUR EQUIPMENT DESIGN

"Study that illustration a minute. It won't take you long to see the many design possibilities you gain by using S.S.White remote control flexible shafts to connect variable elements to their controls.

"For instance, the flexible shaft coupling gives you a free hand in locating the elements independently of their controls. This is mighty important when it comes to meeting space, wiring and servicing requirements or when you're working for top circuit efficiency.

"As for the control knobs... you get the same freedom in positioning them. This means that you remove many limitations on your cabinet designs and can provide more convenient tuning.

"So, when your circuit design includes variable elements, think of S.S.White flexible shafts. This is a tip many designers of electronic equipment have used to good advantage."

WRITE FOR THIS FLEXIBLE SHAFT HANDBOOK

It contains 260 pages of facts and technical data on flexible shafts and how to select and apply them. Write for a copy.

S.S.WHITE INDUSTRIAL DIVISION
THE S.S.WHITET DENTAL MFG. CO. DEPT. G 10 EAST 40TH ST., NEW YORK 16, N.Y.

FLEXIBLE SHAFTS * FLEXIBLE SHAFT TOOLS * AIRCRAFT ACCESSORIES
SMALL CUTOFF AND SERRIFIC TOOLS * SPECIAL FORMED SHAFTS
RESIN MOUNTED * METAL SPACERS * CONTACT PLASTIC MOUNTING

One of America's AAAA Industrial Enterprises

55A
It's CP for COAXIAL HALF-WAVE DIPOLE ANTENNAS for two-way mobile service

PROVEN RELIABLE AND EFFECTIVE OVER THE YEARS

CP Antennas use one of the most effective—and elemental—forms of antennas serving radio communications. Simple physically and electrically, CP Antennas have earned a reputation for reliability.

CP Coaxial Antennas are ruggedly constructed of selected materials to give satisfactory commercial service under severe operating conditions. They are recommended for both transmission and reception. Power handling capacity is limited only by the rating of the feed line. All antennas feature rust-proof construction throughout with necessary accessories for simple, positive installation.

Ask for your copy of CP Bulletin 106

This new bulletin, compiled by the engineering staff of Communication Products, provides complete data on CP antenna construction, as well as detailed specifications and operational information. Fully illustrated with photographs and schematic sketches. Call or write CP, now, for Bulletin 106 or literature on any of the products listed below.

YOU CAN DEPEND ON THESE CP PRODUCTS

✓ LO-LOSS SWITCHES
✓ TEFLOX TRANSMISSION LINE
✓ AUTO-DRYAIRÉ DEHYDRATORS
✓ COAXIAL DIPOLE ANTENNAS
✓ TOWER HARDWARE

Communication Products Company, Inc.
KEYPORT NEW JERSEY

News—New Products

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

(Continued from page 55A)

Designed for measurements of inductors, capacitors, and determination of resistive and reactive components of impedances, the #1150 has a range of 20 to 20,000 cps, with accuracy to 1 per cent. It is used as a Maxwell, Hay, resonance, series-resistance capacitor, and parallel-resistance capacitor bridge.

Further information is available from Freed.

Six-Inch Log Slide Rule

A 6-inch log slide rule with computational accuracy comparable to a 10-inch rule, incorporating 16 purpose-scales, may be obtained from Pickett & Eckel, Inc., 5 S. Wabash Ave., Chicago 3, Ill.

Described as Model 300, this rule is constructed of a magnesium alloy with optical tongues and grooves machined to 0.001 inch for permanent alignment.

On the front, scales L.L. A-B, T, S, C-D, and L.L.2 are inscribed, on the reverse are L.L.1, DF-CF, CIF, CI, C-D, and L.

An instruction manual, written by Prof. M. L. Hartung of the University of Chicago, is included.

Metal Screen Shielding

Metal screen, described as Lectromesh, suitable for shielding purposes may now be obtained in standard "counts" or designed to your specifications from C. O. Jellif Mfg. Corp., Southport, Conn.

Lectromesh is a screen formed by electroplating copper, nickel, or a combination of both onto a rotating cylindrical matrix.

Standard production includes from 25 to 400 "counts" per square inch. Widths range up to 36 inches, lengths to 100 feet.

(Continued on page 57a)
Cramer Type SX Synchronous Motors

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

(Continued from page 564)

Cathode-Ray Tube Sealing Machine

A new machine to be used in cathode-ray-tube manufacture has been placed on the market by Kahie Engineering Co., 1309 Seventh St., North Bergen, N. J.

This machine is constructed to handle interchangeable adaptors which enable it to seal 16 tubes up to 124 inches, or 12 up to 16 inches per cycle of operation.

The manufacturer claims the close tolerances to which these sealers are built holds shrinkage to a minimum.

Complete information on all types of tube machinery and consultation information on individual problems is available from Kahie.

Thin Mica End-Window Counter

A new Model D81 end-window type Geiger-Müller counter, with the mica window available in thicknesses from 2 to 3.9 milligrams per square centimeter, is available from Nuclear Instrument & Chemical Corp., 223 W. Erie St., Chicago 10, III.

The D81 has a plateau length of 200 to 300 volts, with a slope of only 3 per cent per hundred volts. Threshold voltage is between 1,050 and 1,300 volts, and counting life is at least 106 counts.

An identical counter, D32, with window thicknesses between 1.5 and 2.0 milligrams per square centimeter is also available.

(Continued on page 58A)
Better Reception

with

ACME ELECTRIC

TRANSFORMERS

The definitely better reception of sets powered by Acme Electric transformers, is perceptible to the eye and audible to the ear. This better performance can well be a major sales feature in a competitive market. Our engineering department will assist you in all your transformer needs.

ACME ELECTRIC CORPORATION
4410 Water St.  Cuba, N.Y.

Kilovoltmeter Range Extension Device

A method, which does not require any internal meter changes, for extending the range of portable kilovoltmeters of Shallercross Mfg. Co., 520 Pinney Ave., Collingdale, Pa., has just been devised.

A fitting is attached to the highest-range binding post supporting from 1 to 6 Shallercross 505 Corona resistors, increasing the range as desired.

When a meter movement with full-scale current range of 1 ma is used, each resistor added increases the range by 5 kv.

Audio Sweep Generator

A new automatic audio sweep generator, with a frequency range from 25 cps to 32 kc, has just been announced by Clough Brengle Co., 6014 Broadway, Chicago 40, Ill.

Within the range described, the automatic sweep may be adjusted to any spread from 500 cps to 10 kc, or it may be operated manually.

The manufacturer claims that distortion is less than 0.5 per cent. Sweep calibration is linear, and adjustable from 2 to 10 sweeps per second. Panel calibration is direct on a 17-inch Verni-Vider dial.

Complete data is available on request of Bulletin 27A.

Heterodyne Eliminator

Described as type MCL-4 Signal Splitter, a heterodyne eliminator has been de-
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 59A)

developed for commercial purposes by J. L. A. McLaughlin, P.O. Box 529, La Jolla, Calif.

Intended for use with communications receivers having an if of approximately 455 kc, the MCL-4 has off frequency interference attenuation approximately 100 db down at 1.5 kc removed from carrier or telephone reception.

Approximately 2 watts output power are available for telephone reception; 4 watts output for CW. Line and speaker are provided with 500-ohm and 5,000-ohm terminals. Power supply is 105 to 125 volts, 60 cps, self-contained.

Reversible Polarity DC Power Supply

The Model 203, 0- to 30-kv dc power supply with reversible polarity, is being produced by Beta Electric Corp., 1762 Third Ave., New York 29, N. Y.

The output is variable-controlled; a kilovoltmeter is included on the panel to indicate its magnitude. Voltage is continuously variable from 0 to 30 kv. Current rating is 2 ma maximum (determined by rectifier rating). Approximately 300 microamperes at 30 kv.

Input is 117 volts, 50 to 60 cps, 225 volt-amperes maximum.

Microwave Pulse Amplifier

A new wide-band amplifier designed for amplification of microwave pulses has been introduced by Hewlett-Packard Co., 395 Page Mill Rd., Palo Alto, Calif.

Described as -hp-460A, this amplifier has a pulse rise time of 0.003 microsecond, and provides a gain of 20 db. Five instruments may be cascaded to provide additional gain.

When used in conjunction with -hp-410A vacuum-tube voltmeter, the -hp-460A increases voltmeter sensitivity 10 times at frequencies up to 200 Mc. This makes reading of voltages as low as 0.01 volts possible.

Television Tube Improvement

An improved line of cathode-ray tubes of the bent gun type utilizing a single ion trap magnet has been developed by the Tube Div., Allen B. DuMont Labs., Inc., 2 Main Ave., Passaic, N. J.

(Continued on page 60A)
Portable Precision for the Field Engineer!

STODDART NM-20A
RADIO INTERFERENCE AND FIELD INTENSITY METER
• A portable unit that you can DEPEND upon! Designed especially to withstand the rigors of all-weather field operation and yet provide reliable performance.
• Measures FIELD INTENSITIES of radio signals and r.f. disturbances using either a rod antenna or a rotatable loop antenna.
• May be used as a two-terminal r.f. voltmeter (balanced or unbalanced), frequency selective over the CONTINUOUS RANGE 150 kc to 25 mc.
• ONE MICROVOLT SENSITIVITY as a two-terminal voltmeter; 2 microvolts-per-meter using rod antenna.
• Operates from self-contained dry batteries or external A.C. power unit providing well-regulated filament and plate supplies.

STODDART AIRCRAFT RADIO CO.
Main office and plant:
6644 Santa Monica Blvd.
Hollywood 38, Calif.
Phone: Hollywood 9294

1244 Connecticut Ave.
Washington 6, D. C.
Phone: Trinity 1-9260

8-247 General Motors Bldg.
Detroit 2, Michigan
Phone: Hudson 7313

STODDART NM-20A
RADIO INTERFERENCE AND FIELD INTENSITY METER

EACH SOLVED A SPECIAL PROBLEM

HEAD AND FORMED WIRE CONTACTS

YET these are "STANDARD" NEY PRECIOUS METAL COMPONENTS

Chart shows form and overall dimensions of a few of the many types of contacts made from Ney Precious Metal Alloys for brush or wiping contact applications. Full technical and test data are available on request. Other Ney Precious Metal Alloys have solved many special industrial application problems. Consult us freely without obligation.

THE J. M. NEY COMPANY
171 ELM STREET • HARTFORD 1, CONN.
SPECIALISTS IN PRECIOUS METAL METALLURGY SINCE 1812

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 59A)

The manufacturer states that this type of ion-trap design eliminates ion-spot blemishes, and since it bends the beam only once, results in an undistorted spot.

The new types 12RP4 and 15DP4 replace respectively types 12JF4 and 15AP4. These tubes serve as direct replacements except for the single beam-binding magnet which may be added at low cost.

New Model Oscillator

A new oscillator, Model M, with a continuously variable frequency from 0.25 cps to 120 kc has been developed by Southwestern Industrial Electronic Co., 2831 Post Oak Rd., Houston 19, Texas.

Two positions are provided on the range selector switch for fixed frequencies which are determined by units which plug into the central tuning assembly. Plug-in units may be obtained for frequencies as low as 0.1 cps. Special oscillators which are modified slightly to allow operation with good wave form down to 0.1 cps can be obtained, but their dynamic characteristics are not as good as the standard model, because a longer time is required for the level to stabilize after a frequency change has been made.

Portable Projection Oscilloscope

A new Model 701, portable oscilloscope employing the "Norelo" projection system with a 16X12 inch screen, has been developed by Beta Electric Corp., 1762 Third Ave., New York 29, N. Y.

Very useful for educational and demonstrative purposes, the Model 701 when closed is a cabinet 13 inches wide, 16 inches high, and 19 inches deep, weighing 60 lbs. with the screen inside the case.

Provisions are made for external 60 cps, or internal sweep synchronization; 6.3 (Continued on page 62A)
S.S. WHITE RESISTORS

ARE USED IN HIGH VOLTAGE "HIPOT" COUPLERS

S.S. White resistors are connected in series to permit a current flow to ground, when the "Hipot" Coupler is used to measure or to synchronize voltage of high voltage lines.

Canadian Line Materials, Ltd.—maker of "Hipot" Couplers and other transmission, distribution and lighting equipment—says: "We have always found S.S. White resistors of the highest quality." This checks with the experience of the many other producers of electrical and electronic equipment who use S.S. White resistors.

WRITE FOR BULLETIN 4906

It gives details of S.S. White Resistors including construction, characteristics, dimensions, etc. Copy with price list on request.

S.S. WHITE INDUSTRIAL DIVISION

FLEXIBLE SHAFTS AND ACCESSORIES
MOLDED PLASTICS PRODUCTS—MOLDED RESISTORS

One of America’s AAAA Industrial Enterprises

---

Which of these books do you want to examine 10 DAYS FREE?

1. WAVEFORMS

Vol. 19, MIT Rad. Lab. Series. Edited by Britton Chance, E. F. MacNichol, University of Penn.; F. C. Williams, Manchester University; V. W. Hughes, Columbia University; and D. Sayre, Alabama Polytechnic Institute. 776 pages, illustrated, $10.00.

A detailed description of the generation and use of precisely controlled voltages and currents, introducing methods of wave shaping by linear circuits elements and negative feedback amplifiers. The properties of vacuum tubes as nonlinear circuit elements and their application to waveform manipulation are presented in detail.

2. VACUUM TUBE AMPLIFIERS


Here is a complete analysis of important types of amplifiers together with their design principles and constructional techniques. The amplifiers discussed provide special characteristics such as very high gain, large bandwidth, or precise response.

3. COMPONENTS HANDBOOK


This book compiles information on the properties and characteristics of most electronic components. The first part lists fixed components such as wires, cables, resistors, etc.; the second deals with electronic devices, and the third section is devoted to vacuum tubes and cathode ray tubes.

4. ULTRASONICS

By Benson Carl, Hillyer Instrument Company; formerly Product Research Supervisor, Sperry Products; 264 pages, 162 illustrations, $5.00.

Here is the first engineering consideration of the ultrasonic field — the theory plus practical information never before published! This new book reviews electronic considerations and outlines of circuits. Mechanical and electrical design and construction techniques of ultrasonic systems are included.

It brings you valuable information on: material testing, agitation, ultrasonic transducers, ultrasonic systems. It explains clearly the characteristics of ultrasonic waves that are important in practical applications: curves, waves, and complex waves; Fourier's theorem, wave trains and the law of angular transmission; the ways ultrasonic waves may be produced, and the electro-mechanical converting systems.

Send No Money

MAIL COUPON FOR FREE TRIAL

McGraw-Hill Book Co., 330 W. 42nd St., N.Y. 18

I send one book (enclosing in numbers ordered below) for 10 days' examination on approval. If in 10 days I will return one book (1 copy, plus a few cents for delivery, and return unbound book(s) postpaid. I will pay for delivery if you remit with this coupon; same return privilege.

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Mail coupon to Dept. 61A

This offer applies to U. S. only
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)

volts ac signal is available from binding posts. Vertical deflection is approximately 60 millivolts rms per inch, or 0.6 volt full scale. Horizontal deflection is approximately 0.65 volt rms per inch, or 1 volt full scale.

New Type Permanent Magnet Materials

Two new types of materials for the fabrication of permanent magnets, Alnico 5 DG (directional grain), and Alnico 7, are announced by Chemical Dept., General Electric Co., Pittsfield, Mass.

A change in the manufacturing process causes the crystal structure of the magnets to be aligned in the direction of magnetization. It is claimed, as a result of this, that smaller magnets may now be used to perform as larger magnets formerly did; and, that Alnico 5 DG will provide the highest external energy and residual induction of any permanent magnet material known today.

The other product, Alnico 7, has also been developed specifically for applications where a high demagnetization force is present such as in motors, generators, and variable air gap devices. This new magnet shows a higher coercive force than any other grade of Alnico.

Beta Gauge for Industrial Applications

The Model SM-2, the first of a series of industrial measuring and control instruments using the isotope Strontium-90 as a source of radiation is announced by Tracerlab, Inc., 55 Oliver St., Boston 10, Mass.

The sheet material to be measured is interposed between the source and the detector and a part of the radiation is absorbed by the sheet material in proportion to its weight per unit area. Weight per unit area or thickness is read on a properly calibrated recorder connected to the detector. The recorder scale can be calibrated in terms of a plus or minus deviation from specifications or as an absolute thickness or weight reading.

One of the advantages of these gauges is the fact that no physical contact is made with the material being measured, causing no marking of delicate or easily marred surfaces, as is the case with mechanical and other contacting gauges.

(Continued on page 63A)
**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)

**Oscilloscope Time Calibrator**

An instrument for measurement of elapsed time between any two points on the oscilloscope trace is being manufactured by Owen Laboratories, 9130 Orion St., San Fernando, Calif.

Designated as the Type 160, this accessory is inserted in the lead from signal source to the input, and allows either the signal to be observed or places time markers along the sweep. These markers appear as the crests of a damped sine wave, with a frequency of 1, 10, or 100 kc, or 1 Mc. Thus a choice of markers having intervals of 1 millisecond, 10 microseconds, 10 microseconds, or 1 microsecond is possible.

**New Socket-Turrets**

A new terminal structure on which the circuit components associated with a vacuum tube may be connected directly at the socket, has been designed by Vector Electronic Co. 1101 Riverside Dr., Los Angeles 31, Calif.

By this means sub-assemblies are readily formed and these can be installed with a minimum of connections, thus simplifying the construction of electronic equipment.

Stray capacitance is generally reduced since the number and length of circuit leads is minimized. By the use of space under the socket which is usually wasted, a compact arrangement can be made. While Socket-Turrets are advantageous for experimental work, they also make possible many economies in production. These mountings are supplied in a variety of sizes and styles having octal, kathrein, miniature, or noval sockets of standard design.

(Continued on page 64A)

**KLYSTRON POWER SUPPLY**

**MODEL 910**

4 Section, Rack Mounted Unit Supplies All Voltage and Current Requirements for Most Types of Klystrons.

**OUTPUTS**

<table>
<thead>
<tr>
<th>Beam Supply</th>
<th>Control Electrode Supply</th>
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<tbody>
<tr>
<td>0 to 300 ma.</td>
<td>0 to —300 V.D.C.</td>
</tr>
<tr>
<td>10 mv. ripple</td>
<td>0 to 1 ma.</td>
</tr>
<tr>
<td>Regulated</td>
<td>0 to 175 V.D.C.</td>
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<tr>
<td>Continuously Variable</td>
<td>0 to 30 ma.</td>
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<tr>
<td>15 mv. ripple</td>
<td>15 mv. ripple</td>
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<tr>
<td>Regulated</td>
<td>Continuously Variable</td>
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Reflector Supply

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<tr>
<th>—50 to —1500 V.D.C.</th>
<th>Filament Supply</th>
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<tr>
<td>0 to 110 ma.</td>
<td>0 to 10 V.A.C.</td>
</tr>
<tr>
<td>10 mv. ripple</td>
<td>0 to 330 ma.</td>
</tr>
<tr>
<td>Regulated</td>
<td>Continuously Variable</td>
</tr>
<tr>
<td>Continuously Variable</td>
<td>Continuously Variable</td>
</tr>
</tbody>
</table>

All sections insulated for 5000 V.D.C.

All Outputs Metered for Both Voltage and Current

Time Delay Relay—Interlock

*Write for Bulletin 910 for Complete Description and Price.*

**FURST ELECTRONICS**

14 S. Jefferson St., Chicago 6, Ill.

**FREE: 32,000 RESISTORS**

These Kits are in use by industrial, design, and research labs.; colleges; trade schools; amateurs.

**PRINTED CIRCUITS**

*You can reduce costs on your present products with the aid of Conducting, Resistance, and Magnetic Coatings.*

**PRINTED CIRCUITS**

can be used in EVERY branch of electricity

**HAND PAINTED CIRCUIT**

You do not have to switch completely to printed circuits to realize part of their savings. Your present product designs can benefit from brush or spray-applied conducting and resistance coatings. A few dabs or short painted or sprayed lines may save you pennies and dimes on any unit. Current uses are for shielding, elimination of resistors and wires, grounding of static charges, increasing call and conductor inductance, and many others.

**521 SUPER KIT**

The 521, Super Kit, is a complete experimental-printed-circuit-paint laboratory. Contains Silver Conducting Paint; Copper Conducting Paint; Low, Medium, and High Resistance Paints; Solvent; and Lacquer. Manual "Design and Repair of Printed Circuits" included free with each Kit.

From your distributor or our factory only $7.27.

*One Kit contains the equivalent of 33,000 1/4-watt resistors.*

**DO THIS:**

1. Order one of our convenient 521 Kits of air-drying conducting and resistance paints.
2. Use it to simplify your circuits or to make complete experimental printed circuits.
3. Tell us what kind of paint you will need in quantity. We will supply it.

**MICROCIRCUITS COMPANY**

Dept. 10K New Buffalo, Michigan
FM SIGNAL GENERATOR
TYPE 204-B
54-216 Megacycles

Specifications:
-0.5% accuracy. Also covers 0.4 mc. to 25 mc. with accessory 203-B Univertex.
VERNIER DIAL: 24:1 gear ratio with main frequency dial.
FREQUENCY DEVIATION RANGES: 0-24 kc., 0-50 kc., 0-240 kc.
AMPLITUDE MODULATION: Continuously variable 0-50%, calibrated at 30% and 50% points.
MODULATING OSCILLATOR: Eight internalmodulating frequencies, from 50 cycles to 15 kc., available for FM or AM.
RF OUTPUT VOLTAGE: 0.2 volt to 0.1 micro-volt. Output impedance 26.5 ohms.
FM DISTORTION: Less than 0.2% at 75 kc. deviation.
SPURIOUS RF OUTPUT: All spurious RF voltages 30 db or more below fundamental.

200 MC BAND WIDTH
40 db GAIN

MODEL 204 WIDE BAND CHAIN AMPLIFIER
Band Width: 100 KC to 200 MC. Gain: 40 db.
With the Model 204: transients, pulses and other high frequency voltages can now be amplified with the advantage of audio amplifiers.
With the Model 204: vacuum tube voltmeters and oscilloscopes are 100 times more sensitive.
With the Model 204: the output of signal sweep and pulse generators and crystal and mercury delay lines are 100 times greater.
With the Model 204: television signals are 100 times stronger.

OTHER WIDE BAND CHAIN AMPLIFIERS AVAILABLE:
Model 200A—10 db gain. Model 202—20 db gain.
Makers of chain amplifiers, temperature controls, variable electronic filters, high frequency probes, and power supplies.

SKL SPENCER-KENNEDY LABORATORIES, INC.
186 MASSACHUSETTS AVE., CAMBRIDGE 39, MASS.

News—New Products
These manufacturers have invited PROCEEDINGS readers to write for literature and other technical information. Please mention your I.R.E. affiliation.

(Continued from page 65A)

- A 26-page booklet entitled "Inco Nickel Alloys for Electronic Uses" has been made available by International Nickel Co., Inc., 67 Wall St., New York 5, N.Y.
- On a bulletin describing four types of soldering guns with interchangeable tips, from Weller Mfg. Co., 808 Packer St., Easton, Pa.
- A new 50-page catalog in color, describing production methods, many measuring instruments, and a list of representatives and repair stations may be obtained by writing to Simpson Electric Co., 5200-18 W. Kinzie St., Chicago 44, Ill.
- A measurement manual, available to chief engineers of radio stations, to assist in making FCC required station performance measurements for FM and AM, may be obtained from Hewlett-Packard Co., 395 Page Mill Rd., Palo Alto, Calif.
- A catalog, #OP-AR50, describing a heavy-duty splash-proof switch with a "plate" type actuator, rated at 15 amperes, 125, 250, or 460 volts ac, single-pole, double-throw, may be obtained from Micro Switch Corp., 11 W. Spring St., Freepori, III.

TV Power Supply
A lightweight power supply capable of providing a source of regulated dc at loads from 200 to 300-ma has been designed by the Television Section, RCA Victor Div., Radio Corp. of America, Camden, N.J.

Suitable for broadcast, laboratory, industrial, and communications applications, the type TY-25A is adapted for use as either a rack-mounted, or as illustrated, a portable unit.
The output is adjustable between 260 and 290 volts, with variations of less than 0.5 per cent from minimum to maximum load. The ac ripple is less than 0.01 per cent from peak-to-peak. Power requirement is 120 volts, 60 cps, 300 watts.

(Continued on page 65A)
STURDY! JOHNSON BANANA SPRING PLUGS AND JACKS

Studs extend full length of springs for added support. High grade nickel plated brass screw machine parts with accurate threads and milled nuts. All plugs can be furnished with nickel, cadmium or silver plating if required. JOHNSON also manufactures spring sleeve types, insulating and head tip jacks, molded round head tip jacks, insulated combination jacks, metal head tip jacks, twin tip jacks and shorting type twin tip jacks.

A.D.C.'s VHF Communication and Navigation Equipment is a

REVELATION

Get static-free communication and the added reliability of omni range navigation with A.D.C.'s Type 17 2-way VHF Communication and Type 15B Omni Range Navigation Equipment. With the 15B tuned to VHF omni stations, you fly directly in less time. You can receive weather broadcasts simultaneously with navigation signals —static free! It simplifies navigation and gives long, trouble-free life. The Type 17 adds an independent communication system for use while the 15B is providing navigational information. Installations for both single and multi-engined planes are made only by authorized agencies.

Improved Sound Pressure Measurement Equipment

The Model GA-1007 sound pressure measurement equipment, with the microphone permanently attached to the tip of the flexible probe, is now available from the manufacturer, Massa Laboratories, Inc., 3808 Carnegie Ave., Cleveland 15, Ohio.

On the left, the co-axial cable to waveguide transition has VSWR of less than 1.25 from 2,700 to 3,200 Mc, and a connectorless coupling accommodating RG-5/U, RG-8/U, or RG-21/U flexible co-axial cable directly, utilizing the inner conductor as the probe. Range of the variable attenuator is from 0.5 to 10 db, with power rating of 1 watt coverage, 1 kw peak.

This instrument will directly measure sound pressure from a few to several million dynes/cm² in a range from 50 cps to 250 kc. The output signal is delivered by a 25-foot cable at an impedance level of 500 ohms. Extension cables have no appreciable effect on the frequency characteristics of the system.

The manufacturer further states that wave fronts whose pressure changes occur in periods of only a few microseconds are easily measured with the Model GA-1007.
The IRE Promotion Package gives three-way coverage, in a balanced campaign to sell the technical radio-electronic market.

Promotional advertising in the monthly magazine of radio engineers, "Proceedings of the I.R.E." aggressively presents the products of your firm to design engineers in the "pre-specification" period. These men are extraordinarily hard to reach, yet control buying through their engineering knowledge. They are the men you have to sell if you want your product or material specified, or designed into the equipment or stations they engineer and plan.

Reference advertising in the IRE YEARBOOK, gives your program year-long service in the buying data book every engineer has handy. Here, you sell when the engineer wants the facts. Your ad faces your product classification or company listings. Your story is told precisely when, and every time it is needed. YEARBOOK advertisers get all the breaks, because they serve the engineer.

Product Presentation is accomplished in the annual Radio Engineering Show to which 15,710 radio engineers came in March 1949. Here, in four days you can do more contact work, at lower cost than in any other way!

Write us for "Electronic Market No. 1" file.
ACCURATE PHASING WITH T.I.C.
METHOD OF GANGLING PRECISION POTENTIOMETERS

MECHANICAL SPECIFICATIONS
- Precision machined aluminum base and cover
  2" diameter, 1" depth.
- Precision phosphor bronze bushing.
- Centerless ground stainless steel shaft.
- No set screws.
- Mechanical rotation—360°.
- Clamping method of ganging permits individual
  adjustment of angular position.
- Temperature range —85 to +165°F.
- Rotational Life—At least 1 million complete cycles
  of revolution.

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- Winding—both linear to 0.2% and nonlinear to
  1% accuracies.
- Palney contact to winding; two-brush rotor take-off
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- High, uniform resolution provided by our method
  of winding non-linear resistances.
- Electrical rotation maximum 320°.
- All soldered connections (except sliding contacts.)

Write today for bulletins on other T.I.C.
products: Z-Angle Meter . . . R.F. Z
Angle Meter . . . R.F. Power Oscillator
Translatory Variable Resistors
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This general line of precision potentiometers was developed in collabora-
tion with the Fire Control Section of the Glenn L. Martin Company.

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Hollywood, Calif.—Hollywood 6-6306

You can do every kind of soldering with this new 250 watt Weller Gun.
Power-packed, it handles heavy
work with ease—yet the compact,
lightweight design makes it equally
suited for delicate soldering and
getting into tight spots.

Pull the trigger switch and you
solder. Release the trigger, and off
goes the heat—automatically. No
wasted time. No wasted current. No
need to unplug the gun between
jobs. "Over and under" position of
terminals provides greater visibility
with built-in spotlight. Extra 5/8"
length and new RIGID-TIP mean
real soldering efficiency.

Chisel-shape RIGID-TIP offers
more soldering area for faster heat
transfer, and new design gives brac-
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get features not found in any other
soldering tool...advantages that
save hours and dollars. Your Weller
Gun pays for itself in a few months.
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soldering—20 pages fully illustrated.
Price 1c of your distributor, or or-
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STODDART NM-10A
RADIO INTERFERENCE AND FIELD INTENSITY METER
- MEASURES radiated and conducted
signals, including pulse or random
interference.
- RANGE—14 kc to 250 kc.
- SENSITIVITY—Field strength
using rad antennas one microvolt-per-meter
to 2 volts-per-meter. Field strength
using shielded loop antennas 10
microvolts-per-meter to 100 volts-per-
meter. As a two-terminal voltmeter,
either balanced or unbalanced, one
microvolt to one volt.
- READS directly in microvolts and db.
- A.C. POWER SUPPLY REQUIREMENTS
105 to 125 volts or 210 to 250
volts A.C. Single phase source may
be ANY FREQUENCY BETWEEN 50
CPS and 1600 CPS. No shock hazard.
- GRAPHIC RECORDER included with
versatile complement of accessories.

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**THE MU-BETA EFFECT CALCULATOR**

described in this issue is available as a precision instrument in permanent form.

**PRICE $5.00 (with case)**

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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 65A)

**New Mutual Conductance Tube Tester**

A portable tube tester, Model 600, with dynamic mutual conductance circuits for accurate test of radio or TV tubes, has been designed by Hickok Electrical Instrument Co., 10551 Dupont Ave., Cleveland 8, Ohio.

Scale readings are directly in micromhos, with ranges of 0-3,000-6,000-15,000 micromhos.

The Model 600 is housed in a portable case, 7½ inches by 11½ inches by 16 inches. Weight, 15 lbs. For complete information, write H. D. Johnson at Hickok.

New Freed Instruments

Introduced for electronic laboratory use, the Freed Transformer Co., 1718-36 W. 42nd St., Brooklyn 27, N. Y. announces a new power supply and a null detector voltmeter.

The A.C. Power Supply Freed is a laboratory instrument with a continuously variable output from .1 volt to 100 volts with variable 60 cycles supply. It is illustrated here.

The #1210 Null Detector and Vacuum Tube Voltmeter is presented as precision equipment designed for a.c. Bridge measurements. It provides simultaneous measurement of the voltage across the unknown and the balance of the bridge. According to the manufacturer, sensitivity to .1. 1. 10. 100 volts and the Input Impedance is 50 megohms shunted by 20 megohms. Frequency range is 20 cycles to 20,000 cycles. Null detector gain is 94 db. Selective circuits for 60 cycles, 400 cycles and 1000 cycles. Frequency range is 20 cycles to 30,000 cycles.

Exhibit Reservations Early and Heavy In 1950 Show

1,835 exhibit reservations have already been made and assigned in the 1950 Radio Engineering Show to be held in Grand Central Palace, March 6-9, in conjunction with the IRE National Convention.

Acting surprisingly fast on the first industry announcement, most of the "famous regulars" who have been in every show since IRE moved to Grand Central Palace, have taken their usual space.

But 33 new companies have been quick to reserve space too. Special interest has been shown in the combination booths with sound demonstration theatres provided for audio exhibitors.

"Island Style" grouping and wider aisles on the second floor have been so well received that this floor is entirely sold out. For full information and Floor plans, write Mr. Robert Marcott, Reservations Manager, Room 707, 303 West 42nd Street, New York 18, N. Y.
Look in the IRE Yearbook—

3347 Electronic Supply Firms Listed:

An alphabetical index of over 3000 firms of interest to Radio Engineers is given. All engineering products for each firm are shown in code numbers so that a complete picture of that firm is presented. Cross indexing is provided for advertisers to save you time in making direct reference to the advertisers' full story.

Index of 75 Products and Services

This year, complete for companies who answered our requests. The Product Index shows ALL firms. Advertisers are given with complete addresses. A turn-to-alphabetical directory gives full data on the briefest listings.

Look It Up In
Your IRE Yearbook

Use Its Engineering
Products Listings!

Choose right and make Big Savings on
SMALL METAL PARTS

COSTS HALVED: Instead of turning and drilling parts like these from solid rod, or stamping and forming them, the BEAD CHAIN MULTI-SWAGE Process automatically swages them from flat stock. By doubling the production rate and eliminating scrap, this advanced process can save you as much as fifty percent of the cost of other methods.

The BEAD CHAIN MULTI-SWAGE Process produces a wide variety of hollow or solid metal parts—beaded, grooved, shouldered—from flat stock, tubing, rod, or wire—of any metal. Sizes to ¼” dia. and 1½” length.

GET COST COMPARISON ON YOUR PARTS—If you use small metal parts in quantities of about 100,000, don't overlook the almost certain savings of this high-speed, precision process. Send sketch, blueprint or sample part and our engineers will furnish facts about Multi-Swage economy. Or, write for Catalog. The Bead Chain Manufacturing Co., 60 Mountain Grove St., Bridgeport, Conn.
ACCURATE OBSERVATION OF WAVEFORMS FROM 10 CYCLES TO 50 MC PER SECOND

50 MC WIDEBAND VIDEO OSCILLOSCOPE FTL-32A

- Vertical amplifier bandwidth of 10 cps to 50 mc.
- High deflection sensitivity over the entire bandwidth.
- Low-capacity probe maintaining high sensitivity.
- Sweep time as fast as one micro-second.

Write for complete FTL-32A brochure.

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Paul Godley 70A
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Samuel Gubin 70A
Helipot Corp. 47A
Hewlett-Packard Co. 3A
International Nickel Co., Inc. 33A
International Resistance Co. 8A & 9A
Jacobs Instrument Co. 51A
E. F. Johnson Co. 59A, 65A
Karp Metal Products Co., Inc. 7A
James Knights Co. 62A
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United Transformer Co. Cov. II
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LOWEST EVER CAPACITANCE & ATTENUATION

We are specially organized to handle direct enquiries from overseas and can give IMMEDIATE DELIVERIES to U.S.A.

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That is just exactly what 180 manufacturers of technical-radio and electronic products and services have been doing by advertising in the PROCEEDINGS of the I.R.E. They have "rolled up their sleeves" and come to market with factual copy which renders a real service to engineer-readers. Their messages have given specifications and working data. They have worked hard to put into their advertising exactly the information that engineers need. Logically, such advertising gets good results and many of these firms have been our customers, year after year. Proudly we present this list of advertisers—all "industry leaders."

The "PROCEEDINGS of the I.R.E."—Published by

THE INSTITUTE OF RADIO ENGINEERS

William C. Copp, Advertising Mgr.
303 West 42nd St., N.Y.C. 18, N.Y.

Circle 6-6357

PROCEEDINGS OF THE I.R.E. October, 1949
Typical of the C-D line of capacitors with built-in quality characteristics is the

**TYPE UP**

SPECIAL TV ELECTROLYTIC

The only etched foil electrolytic approved by one of the country's largest manufacturers of television receivers after a full year's entirely satisfactory experience. Available in round aluminum containers in sizes to meet your specifications.

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Performance makes the difference

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**CAPACITORS**

Long life, free from the troubles that beset run-of-the-mill capacitors, is engineered into every C-D unit. When you specify C-D's you align yourself with the overwhelming majority of engineers who agree that none can match C-D. Your inquiry will receive prompt and intelligent attention.

JAN Catalog No. 400 on paper capacitors also available. Only requests on company letterheads can be filled.

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CONSISTENTLY DEPENDABLE

☆ CAPACITORS
☆ VIBRATORS
☆ ANTENNAS
☆ CONVERTERS
FREQUENCY CONTROL AND CALIBRATION

The main dial is direct-reading on two approximately logarithmic scales for 50 cycles to 40 kilocycles and 10 kilocycles to 5 megacycles. The incremental frequency control (above) is calibrated between -100 and +100 cycles and -10 and +10 kilocycles for the two respective ranges.

50 CYCLES TO 5 MEGACYCLES

This wide-range beat-frequency oscillator has a number of novel circuit arrangements which make it very valuable for use not only as a general-purpose laboratory oscillator, but also for testing all sorts of wide-band circuits and systems.

1. The exceptionally wide range is obtained with a single control knob and a two-position range selector switch.
2. The output voltage, by means of an a-v-c circuit, is held constant within ±1.5 decibels over the entire range.
3. Frequency drift is held to a very small value through carefully designed thermal distribution and ventilation systems.
4. Any small drift remaining may be eliminated by resetting the oscillator to zero beat.
5. A degenerative amplifier minimizes hum and distortion and also equalizes the frequency response.
6. The output voltage is measured by a v-t voltmeter across the output terminals.
7. Calibration may be standardized at any time to within 5 cycles and 500 cycles on the low and the high ranges, respectively.

For taking selectivity curves on tuned circuits over a wide range of frequencies this oscillator is especially useful in that these measurements may be made very rapidly and accurately with this instrument.

TYPE 700—A WIDE-RANGE BEAT-FREQUENCY OSCILLATOR . . . . $700.00

WRITE FOR COMPLETE DATA